5 GHz WIRELESS CHANNEL CHARACTERISTICS ON THE OHIO UNIVERSITY CAMPUS

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Abstract

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This thesis presents a wideband statistical wireless channel characterization for a campus environment in the band around 5120 MHz (just below the Unlicensed National Information Infrastructure (UNII) bands), based upon measurements. To accomplish these measurements, the stationary transmitter was placed on the rooftop of three separate buildings, and the receiver was moved along the ground at pedestrian velocities. The measurement locations chosen were segregated into line of sight (LOS) and non line of sight (NLOS) regions. Important channel parameters, including delay spread and coherence bandwidth, were obtained from power delay profiles, collected using the 50 MHz bandwidth signal. From the measured data, the thesis provides channel models for a 50 MHz bandwidth and other smaller values of bandwidth, by specifying the number of “taps” in the common tapped delay line model, and the tap’s amplitude statistics and relative energies. The channel models are validated by comparing our simulated channel output statistics to those of the measured data; agreement between model and data is reasonable.

Approved: ____________________________________________________________

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Chapter 1

Introduction

1.1 Introduction

This chapter begins with a brief overview of the recent development of wireless personal communications. We then review the importance of channel characterization in this context, and also describe the unlicensed national information infrastructure (UNII) bands. We conclude the chapter by describing the scope of this thesis.

1.2 Growth of Wireless Personal Communications

Wireless communication systems and services have undergone significant advancement since the first mobile radio telephone was invented in 1924 [1] and more so during the past 15 years. The freedom from having a physical ("wired") connection among various communication networks and devices lead to a rapid growth of wireless personal communications. This growth has been mainly fueled by the desire to have wireless communication capabilities within multiple settings. Example settings include the following: a campus, a building or a house where the system may take the form of a Wireless Local Area Network (WLAN); in and around houses using cordless telephones; within a town or city by employing paging or cellular radios; within a state or a country by using cellular and satellite systems. These systems are capable of providing voice, data, messaging, and video as services to the end user.

The cellular systems that were deployed in the early 1980s are known as the 1st Generation (1G) cellular systems [3]. These systems made use of analog frequency
modulation (FM) and frequency division multiple access (FDMA) as the primary modulation and multiple access schemes. Several popular 1G radio standards deployed at this time included the Advanced Mobile Phone System (AMPS) in the US (which is still in use today), Nordic Mobile Telephone system (NMT), European Total Access Cellular System (ETACS), Radio Telephone Mobile System (RTMS), etc. in Europe and Nippon Telegraph and Telephone (NTT) cellular system, Japanese Total Access Cellular System/Narrowband Total Access Cellular System (JTACS/NTACS) in Japan [2].

Whereas 1G cellular systems were analog in nature, the 2nd Generation or 2G cellular systems deployed in the early 1990s use digital modulation schemes and time division multiple access (TDMA) and FDMA multiplexing schemes [2]. Popular 2G standards include a) Global System for Mobile Communications (GSM), b) Interim Standard 136 (IS-136), c) Personal Digital Cellular (PDC), and d) Interim Standard 95 (IS-95) which is more commonly known as cdmaOne or simply code division multiple access (CDMA) [3]. The former three use TDMA, while the latter uses CDMA as the multiple access technique. Services provided by 2G systems are commonly known as personal communication services (PCS), and these operate in the 1900 MHz frequency range in the US and Canada.

In order to accommodate a larger number of users who could make use of voice and higher data rates at substantially lower costs when compared to 2G systems, the International Telecommunications Union (ITU) defined the global standard IMT-2000 (International Mobile Telephone by year 2000) for 3G systems [2]. Some of the advantages of 3G cellular systems include increased capacity, better spectral efficiency, compatibility with already existing 2G systems, worldwide roaming facility thus
providing universal access to the end user, etc. The radio interfaces proposed by IMT-2000 include, a) CDMA2000, b) Wideband CDMA (W-CDMA), also known as Universal Mobile Telecommunications System (UMTS), c) Universal Wireless Communications 136 (UWC-136) and d) Digital Enhanced Cordless Telecommunications (DECT) [2].

As the number of worldwide subscribers keeps increasing, and with this the users’ need for wireless broadband services such as multimedia and Peer to Peer (P2P) services, most companies around the world are deploying high speed Wireless Local Area Networks (WLANs). Universities are no exception to this. The WLAN devices available right now mainly make use of either the IEEE 802.11a or 802.11b standards [4]. Reduction in equipment prices as well as increases in data rates has made WLANs increasingly popular. WLAN devices typically use omnidirectional antennas to cover local regions, and this enables many receivers to connect to the network through the access point. Thus WLANs find extensive use in point to multipoint type networks.

The 802.11a standard is designed to operate in the 5 GHz band, whereas the 802.11b standard operates in the 2.4 GHz band. Since the 5 GHz band isn’t as widely used yet, the systems operating there have a definite advantage over 802.11b as far as interference is concerned, since cordless phones, Bluetooth devices and others also operate in the 2.4 GHz band. In the United States, in both these bands, a license may not be required to legally operate WLAN equipment.

Currently fourth generation or 4G networks have been receiving wide spread attention not only due to their proposed very high data rates, but also with their ability to integrate existing wireless technologies by having a single standard to satisfy users’
needs. Thus these systems are being designed to have a seamless interconnection with existing 2G and 3G systems, ad hoc networks, fixed type networks as well as WLANs (both 802.11a and 802.11b) [5].

### 1.3 Importance of Channel Characterization

Even though the current 3G systems are capable of providing increased data rates, the key to their successful operation lies in utilizing the mobile radio channel in a very efficient manner. Hence providing an accurate characterization of the channel would help understand, for example, the channel’s ability to have the available bandwidth and gain to sustain high data rate services such as multimedia services. These multimedia services require frequent synchronization between audio and video, with reduction in channel bandwidths possibly leading to a loss of audio-video synchronization. As is commonly seen in CDMA and WCDMA devices (such as cell phones and WLAN devices), a rake receiver is able to collect the energy distributed among multipath components and thus is used to mitigate the effects of multipath fading [2]. Design and operation of this rake receiver requires having a thorough knowledge of the number of apparent multipath components, their relative energies, the time they persist, the fading rate or Doppler spread, etc. Channel characterization provides this information, thus helping system engineers design the various correlators in a rake receiver, with each correlator synchronized to an individual multipath component. An accurate characterization of the physical layer can yield a better design and performance of the subsequent higher layers in the OSI protocol stack.
Hence channel characterization is one of the important steps required to understand a communication system’s performance [6]. Accurate channel models will help in utilizing the available frequency spectrum in a most efficient and effective manner. This should lead to a reduced cost in designing a communication system, and the system performance and communication coverage can also be improved. Chapter 2 provides further details on the importance of statistical channel characterization, when we discuss channel models.

1.4 UNII Bands

The Unlicensed National Information Infrastructure (UNII) bands were mainly allocated to enable short distance high-speed wireless communications at a reasonable cost in the 5 GHz frequency range. In 1997, the Federal Communications Commission (FCC) allocated 300 MHz of bandwidth that would primarily help schools connect wirelessly to the Internet, both indoors as well as outdoors, without having to acquire a license [7]. The 300 MHz of bandwidth was separated into bands of 200 MHz (from 5.15-5.35 GHz) and 100 MHz (from 5.725-5.825 GHz) for UNII devices, with these devices expected to share this spectrum with already existing services. As a result, to avoid radio interference, FCC specified different effective isotropic radiated power levels (EIRP) for each of the UNII bands. Table 1.1 lists the EIRP for the three bands [7].

<table>
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<td>5.15-5.25</td>
<td>0.2</td>
</tr>
<tr>
<td>5.25-5.35</td>
<td>1</td>
</tr>
<tr>
<td>5.725-5.825</td>
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In November of 2003, the FCC allocated an additional 255 MHz of bandwidth for UNII devices, in the 5.470-5.725 GHz band [8]. This was done not only to increase the available spectrum, but also to have a common spectrum allocation in most parts of the world for UNII devices. In addition, devices operating in the 5.25-5.35 GHz and 5.470-5.725 GHz range were expected to make use of dynamic frequency selection (DFS) [8], a mechanism wherein a transmitter can dynamically switch to a different channel, when the existing channel’s interference level exceeds a certain predefined threshold.

1.5 Thesis Scope

The goal of this thesis is to provide a wideband statistical wireless channel characterization in the band around 5120 MHz (which is just below the UNII (unlicensed) band), on the Ohio University campus, Athens, OH. For our channel measurements, the stationary transmitter was placed on the rooftop of three separate buildings for three separate measurement “runs,” and the receiver was moved along the ground at pedestrian velocities. As is common, these measurement locations were segregated into line of sight (LOS) and non line of sight (NLOS) regions. Power delay profiles (PDPs) collected using a 50 MHz bandwidth signal were used to extract important channel parameters such as the delay spread and frequency correlation estimates.

Based upon these measurements, we also provide a model for the channel impulse response in the form of a tapped delay line. This resulting channel model accounts for the number of channel taps, their probability of existence, the energy associated with each of these taps, and the tap amplitude statistics. We also provide a brief overview on how
these models can be obtained for different signal bandwidths. Finally we provide a channel model validation for our empirical models by generating simulated channel impulse responses (CIRs), and then comparing the delay spread statistics of these CIRs with that of the data.

1.6 Thesis Outline

This first chapter briefly describes the growth of wireless personal communications, various UNII bands, as well as the thesis scope. Chapter 2 provides a description for the channel impulse response, important channel parameters, and a discussion on the popular fading amplitude distributions used in channel modeling. In Chapter 3, we describe the measurement technique employed in carrying out the wireless channel measurements, along with sample measurement outputs. In Chapter 4, we discuss the channel model results for several different bandwidths and also provide channel model validation. Finally in Chapter 5, we summarize this thesis, provide conclusions and recommendations for future work to supplement this research.
Chapter 2

Overview of Channel Characterization

2.1 Introduction

This chapter provides a discussion of wireless channel modeling, useful in understanding the channel modeling results presented in Chapter 4. The chapter begins with a literature review that presents previous research carried out for wireless channel characterization for campus environments and other applications in the 5 GHz band, followed by a brief description of the channel impulse response (CIR) and important channel parameters. We then provide a description of the CIR form for the mobile channel. Finally we provide a discussion of fading amplitude distributions commonly employed in statistical channel modeling.

2.2 Literature Review

The area of campus channel modeling has received less attention than many other areas, and relatively little attention for the 5 GHz band. For the literature review we note both texts and papers that were used in this thesis. First we cite relevant texts that discuss channel modeling. Chapter 2 in [3] provides a fairly detailed coverage of multipath fading (both frequency selective and non-selective), characterization of the impulse response, and different fading and path loss models for a terrestrial environment. Reference [2] is a useful text that emphasizes wireless personal communications and also covers mobile radio propagation in fair detail. Another useful text is [9], which discusses in depth the mobile radio channel in terms of propagation phenomena, the effect
multipath has on narrowband and wideband signals, and channel modeling concepts for different types of radio channels.

Some recent references that address wireless channel characterization in the 5 GHz band for other applications and for campus environments include [10]-[14]. In [10], the authors provided statistical analysis for time variant mobile channels in the 5.2 GHz band for different measurement scenarios termed “campus,” “crossroads,” “suburban,” and “aircraft hangars.” The transmitter was mobile and the receiver, which employed a uniform linear array of eight antenna elements, was fixed. The campus measurements were limited to a single run, and were done within a roofed courtyard of small dimensions. The two channel statistics reported were the rms delay spread (RMS-DS) and rms azimuth spread.

In [11], the authors developed a statistical channel model for an indoor environment at 5.2 GHz by modeling the correlation between spatial and temporal domains. The measurements were made of single input, multiple output channels, with a single omnidirectional antenna at the transmitter (Tx) and an 8-element array at the receiver (Rx). This reference is cited because of its relative similarity in measurement set-up to ours; its environment though is an indoor setting whereas ours is outdoor with a single antenna at both Tx and Rx.

Reference [12] describes channel statistics and path loss modeling for indoor environments at 5.3 GHz with a 53.75 MHz bandwidth signal. Vertically polarized omnidirectional antennas were used at both the Tx and Rx, with the Rx antenna height fixed and the Tx antenna height varying. Thus the measurement equipment used is similar to ours, in terms of measurement bandwidth and antennas. Measurements were
carried out in a university’s old and new office buildings, and also in large halls inside an airport, covering both LOS and NLOS regions. Delay spreads were quantified in the form of cumulative distribution functions (CDFs), tapped delay line models were used to develop small-scale models, spatial and frequency correlation functions were shown for NLOS regions, and finally Rayleigh and Ricean fading models were used for tap amplitude distributions.

In [13], the authors discuss channel models and parameters obtained in an outdoor setting, at 5.3 GHz. The outdoor environments were segregated into “urban,” “suburban,” and “rural,” with different Tx antenna heights for each case. Path loss models, mean excess delay and rms delay spreads were quantified. For measurement distances limited to 300 m, the maximum mean value of rms delay spread was found to be 88 ns. For urban environments, it was found that the received power not only decreased with an increase in the horn antenna’s angle of elevation, but also that the received power wasn’t scattered over the entire azimuth, which indicated the scattering to be non-isotropic.

Finally in [14], the authors reported wideband results for outdoor multiple input multiple output (MIMO) measurements at 5.2 GHz using a signal bandwidth of 120 MHz for LOS conditions over short distances. The Tx and Rx both employed a linear array of 8 antenna elements and a fast switching technique was employed for channel measurement between each pair of Tx and Rx. These measurements were carried out on a university campus and in a suburban location. Some of the channel parameters listed were Doppler spread, mean excess delay and rms delay spread. CDFs for Ricean factor $K$ were also shown for all measurement locations, with $K$ decreasing with increasing link distance.
In our work, in addition to computing channel parameters such as delay spread statistics and frequency correlation functions, we also develop models for the channel impulse response. These models account for the number of channel taps, their probability of existence, the energy associated with each of these taps, and tap amplitude distributions. For these amplitude distributions, we employ Weibull and Nakagami fading models for 50 MHz and lower signal bandwidths. These models thus provide a complete description for the outdoor campus channel in the 5 GHz band.

2.3 Channel Impulse Response

The channel impulse response is defined as the output of the channel to an impulse input. For a wireless channel (indoor or outdoor), the transmitter and receiver locations, as well as the positions of objects that make up the surrounding environment can be changing. At any time instant, the channel can comprise a number of multipath components that are well modeled as random [9], each having an amplitude, phase and delay. This leads to the impulse response being time varying as well as spatially varying. Hence the random channel can be statistically nonstationary both in space and in time. Figure 2.1 shows a conceptual diagram of the CIR measured at two different time instants. The time varying nature of the channel can be readily observed at $t = t_0$ and $t = t_1$. 
2.4 Important Channel Parameter Definitions

In this section we discuss several well-known channel parameters, such as delay spread, Doppler spread, coherence bandwidth, and attenuation [3]. Most of these parameters can be obtained from power delay profiles (PDPs) [3]. A PDP is a plot of received power as a function of time delay, which can show different multipath components, as in Figure 2.1. These parameters not only characterize the communications channel, but also help system designers to estimate physical link features such as achievable link distance, and required antenna characteristics and transmitter and receiver locations [15]. The channel is accurately modeled as a time-varying, linear filter, hence it is completely characterized by its impulse response [3].
2.4.1 Multipath Delay Spread

Multipath delay spread $T_M$ quantifies the dispersion of an impulse input into the channel, in the time domain. It occurs when different multipath components arrive at the receiver (either by direct transmission, reflection, diffraction, or scattering) with different time delays. Typically the root mean square (RMS) value of delay spread (DS) is used in practice, as it correlates well with digital communication system performance [15]. It also enables design engineers to select an appropriate transmitted symbol rate or the length of an equalizer to eliminate intersymbol interference (ISI) [16].

2.4.2 Coherence Bandwidth

The coherence bandwidth $B_c$ is reciprocally related to the multipath delay spread. It can be defined as the range of frequencies beyond which the channel affects inputs differently. An approximate relation is [16],

$$B_c \approx \frac{1}{T_M}$$

The coherence bandwidth can also be defined as the bandwidth over which channel amplitude correlation is likely to be the same or falls to some value (say 0.7). Commonly used approximations for $B_c$ are $B_c \approx \frac{1}{50T_M}$ and $B_c \approx \frac{1}{5T_M}$ for correlation values above 0.9 and above 0.5 respectively [2].
2.4.3 Doppler Spread

Doppler spread $f_D$ can be defined as the amount of spectral broadening that occurs on a transmitted tone when it traverses the channel. In its simplest form, the Doppler shift of a single transmitted tone is given by the following [2]

$$f_D = \frac{v \cos(\theta)}{\lambda}$$

(2.2)

where $v$ is the maximum relative velocity of the transmitter-receiver pair, $\lambda$ is the wavelength and $\theta$ is angle of arrival between the velocity vector and the propagation vector. Figure 2.2 illustrates the geometry for Doppler shift.

![Doppler shift Illustration](image)

Figure 2.2. Doppler shift Illustration.

The Doppler spread is the range of Doppler shifts. When the transmitted signal bandwidth $B_r$ is large enough in comparison to the Doppler spread, the channel can be considered to be slowly fading, in which case the Doppler spread effects can often be ignored at the receiver. Typically when $B_r > 100f_D$, the fading can be considered slow fading.
2.4.4 Coherence Time

Coherence time $t_c$ is inversely proportional to Doppler spread and can be defined as the time over which the channel’s amplitude correlation falls to a specified value. The approximate value is [2],

$$t_c \approx \frac{1}{f_D} \quad (2.3)$$

The coherence time is well approximated by the time duration over which the channel impulse response is approximately constant. For a distortionless channel, the coherence time must be greater than the transmitted symbol time.

2.4.5 Attenuation

The free space channel is one of the simplest channels, as it assumes the radio frequency (RF) energy that is transmitted through the channel is dependent only on the distance from the transmitter and the frequency of the propagating wave. It tacitly assumes line of sight (LOS) conditions to exist between the transmitter and the receiver throughout the propagation time. For the free space case, attenuation (also known as power loss or path loss) follows the inverse square law, and is given by the following [16]:

$$\alpha = 20 \log \left( \frac{4\pi d}{\lambda} \right) \text{ dB} \quad (2.4)$$

where $d$ is the distance from the transmitter to receiver, and $\lambda$ is the carrier wavelength. Here transmitter and receiver antennas are considered to be isotropic ($G_t = G_r = 1$). This loss is the ratio of the transmitted power $P_t$ to the received power $P_r$ and is specified in decibels (dB).
2.5 Channel Impulse Response Form

For a linear system, the impulse response can be used to completely characterize the system. The output of the linear system is obtained by convolving the impulse response with the applied input. For a free space channel, the impulse response is given by [17]:

\[ h(\tau; t) = \alpha(t)\delta[\tau - \tau_0(t)] \]

(2.5)

where \( h(\tau; t) \) denotes channel impulse response at time \( t \) with an impulse input applied at time \( t - \tau \). In this form, convolution is taken with respect to the delay variable \( \tau \) to obtain the channel output. The quantity \( \alpha \) is a time dependent attenuation factor. For time invariant channels, the impulse response is independent of \( t \), i.e.

\[ h_{TI}(\tau) = \alpha\delta(\tau - \tau_0) \]

(2.6)

The channel transfer function, denoted by \( H(f; t) \), is obtained by taking the Fourier transform of (2.5). Thus [17],

\[ H(f; t) = F\{h(\tau; t)\} = \int_{-\infty}^{\infty} \alpha(t)\delta[\tau - \tau_0(t)]e^{-j2\pi f\tau} d\tau = \alpha(t)e^{-j2\pi f\tau_0(t)} \]

(2.7)

For the time invariant case, the channel transfer function \( H(f) \) is obtained by the Fourier transform of (2.6). Hence, \( H_{TI}(f) = F\{h_{TI}(\tau)\} = \alpha e^{-j2\pi f\tau_0} \). The form of (2.6) indicates that the channel is distortionless, as the shape of the transmitted signal is not altered by the channel. For \( h(\tau; t) \) to be distortionless, the variations of \( \alpha \) and \( \tau \) with time must be small enough in comparison to the signal duration [17]. Distortionless channels are also known as frequency non-selective channels and if they are fading channels, flat fading channels.
Since a mobile channel can comprise several multipath components arriving with
different delays, a generic representation of the channel impulse response can be given by [17]

\[ h(\tau; t) = \sum_{k=0}^{L} z_k(t) \alpha_k(t) \exp \{ j[\omega_{D,k}(t - \tau_k(t)) - \omega_k(t) \tau_k(t)] \} \]  \hspace{1cm} (2.8)

This channel model is very well suited for wideband signals, such as the one used in our
channel measurements. As can be seen from the above equation, the channel impulse
response is a function of attenuations, delays and phase shifts. We also account for a
“persistence” process, described below. The terms in (2.8) are defined as follows [3]:

1. \( h(\tau; t) \) again corresponds to the channel output at time \( t \) from an impulse input
   applied at time \( t - \tau \).
2. \( L \) is the number of multipath components or echoes, known as “taps”. The value
   of \( L \) represents the duration of the CIR and is dependent on bit, symbol or chip
duration. It is given by [15]
   \[ L = \left\lceil \frac{T_M}{T} \right\rceil + 1 \]  \hspace{1cm} (2.9)
   where \( T \) is the signaling duration, \( T_M \) the delay spread and \( \lceil x \rceil \) denotes the
   ceiling function, the smallest integer greater than or equal to \( x \).
3. \( z_k(t) \) is the tap persistence process for the \( k^{th} \) received multipath echo, which
   assumes an “on” (1) value for the presence of a multipath component and an “off”
   (0) value for its absence. Typically we employ a threshold (20 or 25 dB) below
   the main tap, to determine the tap’s presence or absence.
4. \( \alpha_k(t) \) is the amplitude of the \( k^{th} \) echo path at time \( t \).
5. The $\delta$ function is a Dirac delta, with the $k^{th}$ echo path arriving at the receiver with time dependent delay $\tau_k(t)$.

6. The exponential term’s argument denotes the phase of the $k^{th}$ received echo. It can be written as

$$\phi_k(t) = \omega_c(t)\tau_k(t) - \omega_{D,k}(t - \tau_k(t))$$

(2.10)

where $\omega_c(t) = 2\pi f_c(t)$ is the carrier frequency in radians, with $f_c$ the carrier frequency in Hz, and $\omega_{D,k}(t)$ is the radian Doppler shift of the $k^{th}$ multipath echo, given by $\omega_{D,k}(t) = 2\pi f_{D,k}(t)$. As seen earlier, $f_{D,k}(t) = \frac{v(t)\cos[\theta_k(t)]f_c}{c}$,

where $v(t)$ is the relative velocity, $\theta_k(t)$ is the angle between the $k^{th}$ received vector and the velocity vector and $c$ is the speed of light in free space.

Since $v(t)$ for our case is of the order of 1-2 m/s, which is very small compared to $c$, $f_{D,k}(t)$ varies very slowly, hence in (2.10) the second term, $\omega_{D,k}(t - \tau_k(t))$ is very slowly changing with respect to the first term. For our case, the carrier frequency is in the 5 GHz range, which means that changes in $\tau_k$ on the order of a wavelength of 6 centimeters can cause large ($2\pi$) phase shifts.

When the range of values of $\tau$ is small in comparison to the symbol duration, the paths are irresolvable to the receiver, in which case, $\tau_k(t) \equiv \tau_0(t)$. In this case (also with the attenuations all being nearly equal), the $\alpha$’s, $\tau$’s and the persistence process $z(t)$ are no longer a function of multipath index $k$ and hence can be taken out of the summation. Hence for flat fading channels, (2.8) can be rewritten as
\begin{equation}
    h(\tau; t) = \alpha(t) \delta[\tau - \tau_0(t)] z(t) \sum_{k=0}^{L} \exp\{j[\omega_{d,k}(t - \tau_k(t)) - \omega_s(t) \tau_k(t)]\}
\end{equation}  \tag{2.11}

Again we note that all the echoes can be considered to have nearly the same delay when these delays fall within the minimum measurable delay value of the channel measurement. This minimum delay value is known as the measurement delay resolution, denoted \(\Delta\tau\). The delay resolution is approximately equal to the reciprocal of the transmission bandwidth \(B_{\tau}\). Hence with our channel sounder bandwidth of \(B_{\tau} = 50\) MHz, \(\Delta\tau \approx 20\) nanoseconds. In terms of distance resolution, this equates to approximately 6 meters; this means that two or more multipath echoes are not resolvable if they differ in propagation distance by approximately less than or equal to six meters.

As can be seen from (2.8) and (2.11), the channel impulse response can be modeled as a linear filter having a complex low pass response. This model can be represented by the well known “tapped delay line” (TDL) channel model [9]. Figure 2.3 shows a block diagram of the TDL model [9].

\begin{figure}[h]
    \centering
    \includegraphics[width=\textwidth]{tapped_delay_line.png}
    \caption{Tapped delay line channel model.}
    \end{figure}
This model is composed of various delay elements and multipliers (tap weights) that vary randomly. Here $x_k$ denotes the $k^{th}$ input symbol, $y_k$ the $k^{th}$ output symbol, with $k$ a time index. As before, $\tau$'s denote delays, with $\tau_a$ representing the group delay for the first symbol and all other delay blocks representing delay differences. For our 50 MHz signal bandwidth, the chip time is 20 nanoseconds, and delays are quantized to integer multiples of this [17]. For lower bandwidth signals, this delay interval between echoes will naturally increase. The time varying complex impulse response amplitudes are represented by the $h$'s in Figure 2.3. Specifically for the $k^{th}$ tap, $h_k(t)$ is given by

$$h_k(t) = z_k(t)\alpha_k(t)e^{j\phi_k(t)}$$  \hspace{1cm} (2.12)$$

where as defined earlier, the $z_k$'s, $\alpha_k$'s and $\phi_k$'s are the time varying tap persistence processes, fading amplitudes, and phases, respectively.

### 2.6 Fading Models

The use of deterministic channel models would require accurate knowledge of all the physical parameters in the local environment [9]. This means knowledge of the electrical and physical size parameters for all reflectors, scatterers, and diffractors at any given time. If the number of pertinent objects in the environment is large enough (as may be found in urban areas), estimating channel parameters can be a very difficult task. In mobile cases, the generation of deterministic channel models would require processing large sets of data, since the transmitter, receiver, and obstacles in the channel may be mobile and hence occupy a large number of positions over time. Hence employing deterministic models based upon solving electromagnetic field equations would be impractical in this situation. Thus as is common, the channel models that we employ are
statistical in nature. The path amplitudes are modeled as time varying random processes, which can be modeled using popular fading models. The models that are most commonly used to describe the tap amplitude statistics are given below.

2.6.1 Rayleigh

When there exists no dominant LOS path between the transmitter and receiver, with the channel composed of large number of reflectors and scatterers, and the received components are of approximately equal amplitude, the received envelope can be well modeled using the Rayleigh fading model [3]. This model represents something of a worst case condition. Usually the number of paths $L$ is considered to be fairly large. In this case, the composite received signal is made up of in-phase ($I$) and quadrature ($Q$) components which are well modeled as independent and identically distributed (i.i.d.) zero mean Gaussian random processes [18]. The received phases are uniformly distributed over $[0,2\pi]$ since each multipath component’s phase can change rapidly over $2\pi$ radians for small delay changes (the product of the carrier frequency times the delay varies substantially even though the delay is slowly varying in time). The Rayleigh probability density function (pdf) is given by [2]

$$p(r) = \begin{cases} \frac{2r}{\Omega} \exp\left(\frac{r^2}{\Omega}\right) & (r \geq 0) \\
0 & (r < 0) \end{cases}$$

(2.13)

where $r$ is the received envelope amplitude, given by $r(t) = \sqrt{I^2(t) + Q^2(t)}$ and $\Omega$ is the mean square value, given by $\Omega = E\{r^2\}$. The squared amplitude is exponentially distributed. This model is typically used in urban environments, where it may be difficult
to obtain a LOS path. Figure 2.4 shows the analytical Rayleigh pdf, with $\Omega$ being normalized to unity in this case.

![Rayleigh Probability Density Function](image)

**Figure. 2.4.** Rayleigh Probability Density Function.

### 2.6.2 Chi-Square

Another distribution commonly used in fading channels is the chi-square pdf. This pdf is the pdf of the square of the Rayleigh received envelope amplitude. For a chi-square pdf (with 2 degrees of freedom), the distribution is given by

$$p(g) = p(r^2) = \begin{cases} e^{-g/2} & (g \geq 0) \\ 0 & (g < 0) \end{cases}$$

where $g$ is the received envelope power, having $E(g) = 1$ and $Var(g) = 1$. Figure 2.5 shows the chi-square pdf, and as can be seen, the distribution has an exponential shape.
2.6.3 Rician

When the channel has a dominant LOS component in addition to a large number of weaker, nearly equal-strength multipath components, the fading amplitude can be characterized by the Rician distribution [2]. In this case, the composite received signal consists of $I$ and $Q$ components that are Rayleigh distributed, and an additional LOS signal component. The Rician pdf is given by [2]

$$p(r) = \begin{cases} 
\frac{2r}{\Omega} \exp \left( -\frac{r^2 + A^2}{\Omega} \right) I_0 \left( \frac{2Ar}{\Omega} \right) & (A \geq 0, r \geq 0) \\
0 & (r < 0) 
\end{cases}$$

(2.15)

where $A^2$ denotes power in the specular (LOS) component and $\Omega$ denotes power in the multipath components. The function $I_0(\cdot)$ is the modified Bessel function of the first
kind, zero order. The mean envelope power is given by \( E[r^2] = A^2 + \Omega \). The phase distribution is concentrated at the phase of the LOS component, with the phase of the scattered components uniform over \([0,2\pi]\) just as in the Rayleigh case. These quantities in \( E[r^2] \) are normally represented by the Rice factor, also known as the “K” factor. We have \( k = \frac{\text{LOS power}}{\text{Scattered power}} = \frac{A^2}{\Omega} \), and \( K \) is expressed in dB [2]:

\[
K = 10 \log \left( \frac{A^2}{\Omega} \right) \text{ dB} \quad (2.16)
\]

When \( k \) is 0 or from (2.15), when \( K = -\infty \), there is no LOS component and hence the pdf is Rayleigh distributed. When \( k \) takes on large values, the channel can be considered to be an essentially non-fading one. Hence the Rician pdf can serve as a consolidated model for the Rayleigh pdf and unfaded cases such as the additive white Gaussian noise (AWGN) channel. Figure 2.6 shows Rician pdfs for different \( K \) factors, obtained by setting \( E[r^2] = 1 \). It can be observed that when \( K = 0 \), the pdf is a Rayleigh pdf and as \( K \) increases, the pdf resembles an impulse.
2.6.4 Nakagami-m

Similar to the Rician pdf, the Nakagami pdf serves as a flexible model that can yield the Rayleigh pdf, approximate the Rician pdf, as well as approximate an unfaded AWGN channel. This pdf is known to provide a good fit for empirical data [3] and can model fading conditions fairly accurately in a mobile channel. The Nakagami pdf is given by [3]

\[
p(r,m) = \begin{cases}
2 \Gamma(m) \left( \frac{m}{\Omega} \right)^m r^{2m-1} \exp\left( \frac{mr^2}{\Omega} \right) & (r \geq 0, m \geq 0.5) \\
0 & (r < 0)
\end{cases}
\]  

(2.17)
Here the pdf is a function of two parameters, the received energy $\Omega$ and the “shape factor” or the “fading figure” $m$. Hence this model offers substantial flexibility in comparison to the Rayleigh. The function $\Gamma(\cdot)$ is the Gamma function and $\Omega = \mathbb{E} \{ r^2 \}$ again denotes the mean square value. The shape factor is given as:

$$m = \frac{E^2 \{ r^2 \}}{\text{var}(r^2)} = \frac{E^2 \{ r^2 \}}{E \{ r^4 \} - [E \{ r^2 \}]^2}$$  (2.18)

The shape factor indicates fading severity: the smaller the value of $m$, the more severe the fading becomes. For the special case of $m = 1$, the Nakagami pdf becomes the Rayleigh pdf. When $m \to \infty$, the pdf becomes an impulse, which specifies the AWGN case. Figure 2.7 shows the Nakagami pdf for different values of $m$. As can be seen when $m = 1$, the pdf is a Rayleigh pdf, with $m = 0.5$ indicating worse than Rayleigh fading.

Figure 2.7. Nakagami-m distribution.
2.6.5 Weibull

The last type of pdf that we consider in our discussion is the Weibull pdf. Similar to the Nakagami distribution, this model doesn’t rely upon propagation theory, but mainly on empirical data [9]. The Nakagami pdf can be thought of as a generalization of the Rayleigh distribution, with the received composite signal in this case obtained by transforming the Rayleigh distributed complex envelope [19]. This transformation is given by: 

\[ R_w = R_R^2 \]  

where \( R_R = \sqrt{I^2(t) + Q^2(t)} \) is the Rayleigh distributed envelope as described previously. The Weibull pdf is given by [17]

\[
p(r, b) = \begin{cases} 
\frac{b}{a} r^{b-1} \exp\left[-\left(\frac{r}{a}\right)^b\right] & (r \geq 0) \\
0 & (r < 0)
\end{cases}
\] (2.19)

where \( b \) is the shape factor that indicates fading severity. A factor of \( b = 2 \) indicates Rayleigh fading, with a value of \( b < 2 \) indicating worse than Rayleigh fading. The parameter \( a \) is the scale factor given by [17]

\[
a = \sqrt{\frac{\Omega}{\Gamma\left(\frac{2}{b}\right) + 1}}
\] (2.20)

where \( \Gamma(\cdot) \) is the Gamma function and \( \Omega = E\{r^2\} \) is the mean square value. Figure 2.8 shows the Weibull distribution for different shape factors. As can be observed, for \( b = 2 \), the pdf is a Rayleigh pdf, with \( b = 1.5 \), indicating fading that is worse than Rayleigh fading.
Figure 2.8. Weibull distribution with different shape factors.
Chapter 3

Measurement Technique

3.1 Introduction

In the previous chapter, we described the motivation for channel modeling, and some of the important channel parameters. Our next aim is to describe the way in which the channel measurements were carried out. This chapter deals mainly with this topic. We start with a brief description of the method used by our channel sounder to measure the channel’s impulse response. We then describe the features of our test equipment, which is followed by a description of the campus environment where we took our measurements. Test plans and example photographs are provided for the campus environment. Finally we provide measurement outputs, which depict sample plots of power delay profiles, channel transfer function, RMS delay spread distributions and frequency correlation estimates for different propagation conditions.

3.2 Overview of basic DS-SS Stepped Correlator Method

As stated, most mobile channels can be modeled using linear filters and hence can be completely characterized by their channel impulse response, denoted $h(\tau;t)$. The function $h(\tau;t)$ is the response of the channel at time $t$ to an impulse input applied at time $t - \tau$. It is seen from [9] that for a linear system, when white noise is applied at the input, and this delayed input is cross correlated with the output, the resulting expectation or cross correlation is proportional to the system impulse response which is a function of delay variable $\tau$. This can be written in equation form as [9]
\[ E[n^*(t - \tau)y(t)] = N_0 h(\tau), \] (3.1)

with \( n(t) \) the zero-mean white noise process with one-sided power spectral density \( N_0 \), and \( y(t) \) is the filter output. Hence for a linear system, we can use white noise to obtain the channel impulse response. But this is rarely implemented in practice since white noise has a flat psd (a fixed non-negative value of Power/Hz at any given frequency) and more importantly an infinite bandwidth, which makes this type of impulse response measurement impossible. An alternative to this would be to make the noise band limited, and then measure the impulse response using (3.1). However the indeterminate (random) nature of noise over time makes this approach unrealistic.

Considering these limitations of white noise, the most common deterministic signals used to estimate impulse responses are signals composed of pseudonoise (PN) sequences [9]. The PN sequences are also known as “maximal length,” and are generated using linear feedback shift registers. The “pseudonoise” or “pseudorandom” name comes from the fact that the autocorrelation of PN sequences resembles an impulse.

This type of signaling is also used by our channel sounder, specifically by using the PN sequence in a spread-spectrum manner. This spread spectrum signal is the direct-sequence (DS) type of spread spectrum (DSSS), and the PN signal is used to modulate a sinusoidal carrier in the desired band. This modulated signal is then transmitted onto the channel. The received signal is the convolution of the modulated signal and the channel impulse response. This output is then cross correlated with an exact replica of the transmitted PN sequence (at the receiver) to obtain the desired impulse response estimate (IRE). Figure 3.1 shows the process of impulse response estimation obtained using the DSSS technique.
In Figure 3.1, \( c(t) \) denotes the spreading code (PN sequence) and for simplicity, the channel impulse response is represented by the time-invariant version \( h(t) \). The PN code is transmitted repeatedly, to allow gathering of sufficient energy at the receiver, and also to allow data to assess channel time variations. The PN sequence elements, which are pulses known as chips, serve to emulate an impulse input. Since the chip sequence is convolved with the actual CIR, the delay resolution is limited to the chip duration \( T_c = \frac{1}{R_c} \). It is desirable to have the chip rate \( R_c \) sufficiently large such that any 2 multipath echoes can be discernible at the receiver. With a resolution of \( T_c \), the receiver estimates two multipath impulses as a single pulse if they are separated by less than \( T_c \) seconds. This is one of the limitations of using this method.

Possibly the earliest use of the spread spectrum technique to carry out impulse response measurements was by Cox [20] at 910 MHz in suburban New Jersey. At the transmitter a spreading code of 511 bits (clocked by a 10 MHz clock) phase reversal (binary phase-shift key) modulated a 70 MHz carrier wave. This modulated signal was then upconverted to 910 MHz by mixing it with a carrier wave of 840 MHz, and then transmitted onto the radio channel. At the receiver, the pseudonoise sequence used to
cross correlate the channel output is identical to the transmitted PN sequence in appearance but was clocked at a slightly slower rate (9.998 MHz). This “slow” sequence was correlated with the received sequence and was reset in time every 75 milliseconds to obtain a power delay profile (PDP) that has a delay span of 15 microseconds (μsec).

Hence this correlator can be considered a “sliding” correlator because of the difference in timing associated with the PN sequences at the transmitter and receiver.

As mentioned earlier, our channel sounder also makes use of PN sequences and the correlation procedure is similar to the one used by Cox, but for the following important differences. The clock signals used at the transmitter and the receiver have the exact same frequency. The receiver correlates at some delay \( \tau \) for a certain time and outputs this correlated value. It then shifts its delay by chip time \( T_c \) and correlates at this delay, then outputs this value, then shifts by \( T_c \) again, and so on until the last chip of the sequence is reached, thus covering the specified delay span. Thus, instead of a sliding correlator, we have a “stepped” correlator.

The correlated output value associated with a particular delay shift (also known as delay bin) is proportional to the square root of the power collected over that bin. Hence, for our channel sounder with a length \( N_c = 255 \) chip sequence, clocked by a 50 MHz signal \( (T_c = 20 \text{ nanoseconds}) \), the unambiguous delay span is given by the product of \( N_c \) and \( T_c \), which equates to 5.1 μsec. In terms of distance this corresponds to a delay difference of approximately 1.5 km. One of the limitations of the channel sounder is that if the distance difference between the first arriving impulse and a multipath echo is larger than 1.5 km, then it appears as if that echo is within the 5.1 μsec delay range, because of a “wrap-around” effect of the channel sounder. However during our measurement campaign, in
the fairly cluttered non-line-of-sight campus environment, it is unlikely that distances between significant reflectors could be this large. The receiver has \( I \) and \( Q \) channels, so that it measures the received amplitude and phase, with amplitude-squared proportional to the measured received signal strength information (RSSI).

### 3.3 Equipment Description

Our wireless channel sounder was manufactured by Berkeley Varitronics Systems (BVS) [21]. Their “Raptor” model spread spectrum stepped correlator was modified to operate in the 5 GHz spectrum (the commercial version otherwise operates in the 2.4 GHz band). In addition, our version also can employ a faster output rate as well as a higher output power. The term “sounder” pertains to the set of transmitter (which is used to transmit a test signal onto the unknown wireless channel) and receiver (which receives the channel output). The transmitter and receiver are both powered by 12V DC power supplies, with an option provided for the receiver to be powered by a battery pack. This battery pack was essential for our measurement settings, where as previously mentioned, the transmitter location is fixed and the receiver, mobile. The sounder features are summarized as follows:

**Transmitter**

1. adjustable center frequency ranging from 5090-5250 MHz, with frequency step size of 25 MHz. For our measurements we used 5120 MHz as the carrier frequency.

2. adjustable power level from 5 dBm (\( \approx 3.2 \) milli watts) to 33 dBm (2 watts) with power step size of 1 dB.
3. two chip rates, 25 Mcps or 50 Mcps. We used a chip rate of 50 Mcps.

4. The modulation bandwidth of the transmitted spread spectrum signal is 50 MHz, with the 99% power bandwidth of this signal measured to be approximately 53 MHz using a conventional spectrum analyzer. The modulation used is of type Binary Phase Shift Keying (BPSK).

5. GPS receiver.

Receiver

1. Similar to the transmitter, an adjustable frequency from 5090 to 5250 MHz in step size of 25 MHz.

2. dynamic range of 65 dB (dynamic range is the range of powers at which the receiver can gather PDP data without any problems or damage). The absolute power levels are from a maximum of -20 dBm to a minimum of -85 dBm.

3. minimum multipath delay resolution of 20 nanoseconds. This equates to a distance resolution of approximately 6 m.

4. Just like the Tx, either of the two chip rates: 25 Mcps or 50 Mcps.

5. The receiver uses QPSK demodulation.

6. GPS receiver, which can be used for navigation (GPS antennas have to be connected to the transmitter in order to do so) of the mobile receiver.

Since a mobile radio channel can introduce arbitrary phase variations on the transmitted signal for distance changes of the order of a quarter wavelength, under these circumstances if we used a BPSK demodulator, the correlated output can be zero (as it uses a single correlator), and we do not employ phase tracking at the receiver. Hence as
an alternative, we use QPSK demodulator that uses two correlators (I and Q channels), with either or both of them always giving a non-zero correlated output.

We also made use of an uninterruptible power supply (UPS) during the training period (described later) as well as for the time period during which the transmitter was moved from the training site to its transmitting site. Figure 3.2 shows the channel sounder in training mode, connected in a back to back configuration by an RF cable of low loss. This picture was taken in the Multiuser Mobile Communications Laboratory (MMCL), which serves as a research lab and also as a repository for most of our communications hardware and software. For actual measurements when Tx and Rx are separated, RF cables were used to connect the transmitter and receiver with their respective omnidirectional antennas. These antennas are monopoles above ground planes with an approximate gain of 1.5 dBi.

Figure 3.2. Raptor Channel Sounder with Rx (left) and Tx (right).
An RS-232 cable is used to connect the receiver unit to a laptop computer, which configures the receiver during training mode or measurement mode as well as collects the receiver output data. A wooden platform was used to mount the transmitter unit, its power supply, and the omnidirectional antenna, which can be down tilted to any desired angle in elevation. A similar smaller platform was used at the receiver to mount the omnidirectional antenna. The cart we used to convey all the Rx equipment was spacious enough to hold the antenna unit, receiver unit and its battery pack, and the laptop. Other items that we made use of during our measurements include RF attenuators, adaptors, a pair of walkie talkies for communication from Tx team to Rx team, and a digital camera. Figures 3.3 and 3.4 show photographs of the channel sounder transmitter and receiver, respectively, with their associated equipment and the people involved during the measurement campaign.

Figure 3.3. Raptor Channel Sounder Tx setup
3.4 Environment Description

Before conducting actual measurements, our first task was to select appropriate locations on campus where terrestrial communication is desirable or essential. This then led to selection of transmitter locations, which was done via a visual inspection around the campus to determine potentially suitable (tall) buildings. These locations also had to be near the areas where the receiver unit was to be moved.

Ohio University’s campus is segregated into various regions known as “Greens,” namely East Green (EG), West Green (WG), South Green (SG) and College Green (CG). These Greens have a lot of pedestrian traffic during day time (during which we made our measurements), as in any college campus. Apart from these dynamic scatterers that are prevalent in the campus environment, other potential dynamic scatterers include vehicles
of various sizes. In addition, trees and buildings of various shapes and sizes can cause static scattering.

The first set of campus measurements was conducted on WG for which Stocker Center was chosen to set up the transmitter. Not only is Stocker Center home for all engineering departments and thus is easily accessible, but also acts as a vantage point for other locations on the WG. Figure 3.5 shows an aerial “3 dimensional diagram” of West Green with the Tx location and our numbered (Rx) measurement locations approximately marked. These locations were selected to cover both LOS and non-LOS (NLOS) propagation conditions. For instance points 1-3, 18-20 are LOS regions, whereas points 9-11 cover NLOS regions with no direct path between the Tx and Rx. Some of the test points also covered regions frequently used by pedestrians and vehicles, which correspond to points 1-6 and 16-20, respectively.

![Figure 3.5. View of West Green with measurement test points.](image-url)
The WG Tx antenna height was approximately 15 meters from the ground, with an approximate maximum achievable link distance of 275 meters to point #12. Figure 3.6 shows the Tx location in the background, taken from point #7 in Figure 3.5. As noted, this is a LOS region. Figure 3.7 further illustrates the WG, depicting points 9-10 in Figure 3.5. These pictures were taken during the lunch time, and hence we don’t observe as much pedestrian or vehicular traffic as we do during other normal hours of the day. On West Green we conducted our measurements on 17 November 2005, with outside temperatures being approximately 17 degrees F. No precipitation was observed on that day and the number of PDPs collected was approximately 9000.

*Figure 3.6. View of West Green, with the Tx location in the background.*
In comparison to the West Green, the College Green has a larger number of buildings (as well as trees) and is located adjacent to the downtown Athens area. Hence CG is the busiest of all the Greens. Bromley Hall, with its higher elevation and larger number of stories (nine) is the tallest building on campus. Hence it served as an ideal Tx location to carry out our measurements on the CG. As for the WG measurements, the CG measurement locations were chosen to cover both LOS and NLOS regions, and the reception sites were often among a large number of mobile reflectors and scatterers. It can be noted that even though College Green is relatively an older part of the campus in comparison to the other Greens, the electrical properties of the buildings here don’t vary significantly compared to the other Greens. However the surrounding trees, which are substantially large (and tall), can contribute significantly to attenuation of multipath components. Figure 3.8 depicts the test points and the Tx location on College Green. As
an example, points 1-3 serve as LOS locations and points 11-13 denote the “worst-case” NLOS conditions.

Figure 3.8. View of College Green with measurement test points.

The Tx antenna height was approximately 35 meters, with a maximum achievable link distance of approximately 500 meters to point #12. Most of the pedestrian traffic is concentrated in the “infield” section of CG, denoted by points 4-10 in Figure 3.8. A further illustration of the College Green’s environment is given by Figures 3.9 and 3.10. As can be seen in Figure 3.9, the buildings, parked and moving vehicles, and pedestrians are all possible sources of reflection and scattering or absorption. Figure 3.10 shows a part of the infield section of the CG taken from point #9, with the Tx location in the background. Measurements at CG were carried out on 30 March 2006.
Figure 3.9. Photograph of College Green, showing pedestrian traffic on President Street.

Figure 3.10. View of College Green, showing Tx location (Bromley Hall).
Weather conditions were temperate, with temperatures close to 65 degrees F. Approximately 5800 PDPs were collected on this day.

The final part of our measurement campaign involved sampling the measurement locations on the East Green and South Green. These Greens combined together are fairly large and have a large number of dormitories and associated dining halls. Figure 3.11 shows the Tx location on Morton Hall for measurements on East Green, with Figure 3.12 illustrating the residential nature of South Green. Also worth noting in Figure 3.12 are the hills in the background, located opposite to the Tx site. As before, the measurement regions were segregated into LOS regions and NLOS regions and hence covered a range of propagation conditions.

![Figure 3.11. Picture showing Tx location (Morton Hall), East Green.](image)
Figure 3.12. View of one of the measurement locations on South Green.

Figure 3.13 shows the Tx location along with the associated measurement test points on East Green and South Green. The off campus locations (indicated by lightly shaded blocks in Figures 3.8 and 3.13) were not sampled, even though they were fairly close to the on campus locations where we conducted our measurements. The maximum Tx antenna height was about 20 meters, with maximum link distance about 600 meters to point #16. Measurements at these Greens were carried out on 27 April 2006 with no precipitation observed on this day. The number of PDPs collected was approximately 7700.
3.5 Measurement Description

Before actually carrying out channel measurements, we had to gain access to the roof tops of all buildings, which would be used to set up the transmitter unit. This involved contacting and if required, meeting the person(s) in charge of these buildings. Typically this process required us to mention the topic of the thesis research, why the particular building was selected as the Tx location, our need to access the roof top with a readily available access to AC power, the planned measurement duration and a brief description of the measurement procedure, the people involved in carrying out these measurements, and the frequency band of operation. After obtaining full support, the next step involved contacting the custodial staff, who oversaw the storage area (where the equipment was kept overnight in training mode), as well as the roof top. We normally did
not encounter any kind of public interference whatsoever when the receiver unit was moved on different locations on OU campus.

To ensure no interference with existing wireless services when channel sounding, we conducted a radio frequency interference analysis. This was done in co-ordination with OU’s Communication Network Services (CNS), which manages most of the university’s wireless services. The analysis involved testing an IEEE 802.11a WLAN link in proximity to our channel sounder and verifying their performance in the presence of each other. More details on this testing, its results, and the channel sounder’s out of band emissions are given in Appendix B.

Carrying out accurate wireless channel measurements required us to calibrate the channel sounder first. This calibration period is known as “training” time. The Tx and Rx units use rubidium oscillators, which have very good stability for short durations, but like any oscillators, their frequencies can fluctuate over time. The training time ensures frequency locking of the Tx-Rx pair; with a longer training time, the resulting testing time is also lengthened. In the training mode, the Tx and Rx units are connected in a back to back configuration via a low loss RF cable and an attenuator. The sounder units were trained overnight to conduct measurements for the next morning. This yielded a measurement time of close to two hours in most cases.

Once done with training, the sounder Tx, along with its battery and UPS, was disconnected from the Rx and placed on the wooden platform set up on the roof top. The wooden platform had an omni antenna mounted on top, as seen in Figure 3.3. This was done as quickly as possible, since the UPS has a back up time of about 9 minutes. The
omni antenna was downtilted by 30 degrees in most cases, pointing towards the receiver, based upon the known radiation pattern of the antenna.

The sounder Rx, along with its battery, omni antenna (connected to the Rx via an RF cable) and laptop, was placed onto a cart and carried from the storage area onto the ground, thereby made ready for data logging. A pair of walkie talkies was used for periodic contact between the Tx team and Rx team. The Rx team described its observations to the Tx team, indicating events such as a transition from a LOS region to a NLOS region. A member of the Rx team was also involved in taking notes on the surrounding environment to correlate them with observed PDPs recorded on the laptop. The measurements were made at various test points, as indicated on the test plan, with a new data file recorded for each segment (between two consecutive test points) of the measurement. At the end of each segment, the Rx cart was stopped for a while to close the data file and open a new one. Once the measurements were complete, the Rx team returned to the Tx site where everything was disconnected and packed. Table 3.1 and Table 3.2, from [22], describe the training phase and the channel sounding phase respectively, used during our experiments.

**Table 3.1.** Test procedure for initialization and training phase [22].

<table>
<thead>
<tr>
<th>#</th>
<th>Procedure Description</th>
<th>Estimated Duration (minutes)</th>
<th>Notes</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td><em>Initialization and Training</em></td>
<td>44-104</td>
<td>Training (step 13) MAY be able to be conducted overnight</td>
</tr>
<tr>
<td>1</td>
<td>Position Tx platform, Tx and its power supply, and connect</td>
<td>5</td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>Turn on Tx power (not RF), and warm up</td>
<td>3</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>Connect laptop to Rx, and power up laptop</td>
<td>0</td>
<td>During Tx warmup</td>
</tr>
<tr>
<td></td>
<td>Procedure Description</td>
<td>Estimated Duration (minutes)</td>
<td>Notes</td>
</tr>
<tr>
<td>---</td>
<td>--------------------------------------------------------------------------------------</td>
<td>------------------------------</td>
<td>----------------------------------------------------------------------</td>
</tr>
<tr>
<td>4</td>
<td>Connect power (battery) to Rx</td>
<td>0</td>
<td>During Tx warmup</td>
</tr>
<tr>
<td>5</td>
<td>Turn on Rx power</td>
<td>0.5</td>
<td></td>
</tr>
<tr>
<td>6</td>
<td>Connect RF cable from Tx to Rx, through attenuator</td>
<td>1</td>
<td>Use 40 dB attenuation, minimum, and ensure received power ≤ -10dBm</td>
</tr>
<tr>
<td>7</td>
<td>Set Tx RF $f_c$</td>
<td>0.25</td>
<td>$f_c$=5120 MHz</td>
</tr>
<tr>
<td>8</td>
<td>Set Tx output power $P_{Tx}$</td>
<td>0.25</td>
<td>Max $P_{Tx}$=33 dBm for sounding; ~ +5 dBm for training</td>
</tr>
<tr>
<td>9</td>
<td>Set Tx chip rate $R_c$</td>
<td>0.25</td>
<td>$R_c$=50 Mcps</td>
</tr>
<tr>
<td>10</td>
<td>Turn on Tx RF power output</td>
<td>0.25</td>
<td>Again: $P_{Tx}$~ +5 dBm, with ≥40 dB attenuator between Tx and Rx for training</td>
</tr>
<tr>
<td>11</td>
<td>Invoke Raptor SW on Rx, and make sure display indicates “Raptor Stable,” “Raptor Locked,” and received power $P_{R}$&gt; “-3 dB”</td>
<td>2</td>
<td></td>
</tr>
<tr>
<td>12</td>
<td>Configure Raptor Rx SW</td>
<td>1</td>
<td>During training, do NOT change any settings or display on laptop</td>
</tr>
<tr>
<td></td>
<td>a. set $f_c$ and $R_c$ to match those of Tx</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>b. initiate training</td>
<td></td>
<td></td>
</tr>
<tr>
<td>13</td>
<td>Train for desired duration, depending upon desired measurement duration</td>
<td>30-90+</td>
<td>Measurement duration displayed via max(Current Count, Last Count)</td>
</tr>
</tbody>
</table>

Table 3.2. Test procedure for channel sounding phase [22].
At Tx, set $P_{Tx}=33$ dBm, and turn RF output power ON (sounder transmitting).

At Rx, configure Raptor SW to measure a full delay span. Record location 1 profile, and capture reference RSSI.

At Rx, set up log file for initial course. Rx Team should communicate readiness to begin to Tx Team.

Begin moving from location point 1 to point 2, and observe Rx display. Log PDPs.

When location point 2 reached, stop cart, end PDP logging.

Create new Rx log file, and move from location point 2 to point 3, and log PDPs.

Repeat step 10 for the remaining measurement location points.

File names include Tx location (e.g., Stocker), antenna types, terminal points (e.g., from point 1 to point 2, label 1to2) and Tx antenna tilt angle.

Rx team should communicate to Tx team impending change of location from LOS to NLOS. Rx team should also take notes on environment characteristics.

Both teams should communicate status, and convey any descriptions of problems, anomalies, etc.

In this section we illustrate some of the measurement results. More detailed discussion on these and also on the developed channel models is given in Chapter 4. The illustrations provided below are intended to help the reader comprehend some of the channel parameters that were briefly discussed in Chapter 2.

Figures 3.14 and 3.15 show PDP plots for a NLOS region on the West Green and a LOS region on the College Green, respectively. Root Mean Square values of the Delay Spread (RMS-DS) are also shown for both of these. Details on how we calculate RMS-DS and why it is used in our channel models are given in the next chapter. The NLOS
PDP of Figure 3.14, collected between points 9 and 10 in Figure 3.5, indicates some of the most dispersive conditions found on the WG. As can be seen, the channel energy is spread in time, and as expected, this yields a relatively large value of delay spread. The LOS PDP of Figure 3.15 was obtained between points 2 and 3 in Figure 3.8 on the CG. As there is a direct LOS path between Tx and Rx, most of the channel energy lies in the first arriving component, with no multipath present after about 0.4 microseconds. These PDPs have low values of RMS-DS.

![Figure 3.14. Example PDP plot for a NLOS region at WG.](image)

\( \sigma = 1.84 \mu \text{sec} \)
Figure 3.15. Example PDP plot for a LOS region at CG.

Figure 3.16 shows a 3 dimensional plot for a series of NLOS PDPs that were collected over time. The plot provides an illustration of multipath fading in time. Similar to Figure 3.16, Figure 3.17 shows a 3-D plot of a channel transfer function over time, for a LOS case. The transfer function (TF) is obtained by taking the Fourier transform of each PDP in time. As expected, the channel TF obtained as a result is time variant.
Figure 3.16. Example PDPs vs. Time for a NLOS region, West Green.

Figure 3.17. Example Time Varying transfer function for a LOS region, College Green.
Figure 3.18 shows a histogram that illustrates the distribution of RMS-DS attained over all PDPs that were collected on West Green. This bi-modal distribution indicates the presence of two distinct propagation regions, namely LOS and NLOS regions. A more detailed discussion on how we segregate regions based on RMS-DS values is given in the next chapter.

![Histogram of RMS-DS](image)

**Figure 3.18.** Distribution of RMS-DS for all PDPs collected on West Green.

Finally in Figure 3.19 we show an example Frequency Correlation Estimate (FCE) for a LOS region. FCEs are related to the channel transfer function. Here for a correlation value of 0.7 and 0.3, the frequency separations are 8.2 MHz and 18.4 MHz, respectively. Chapter 4 provides a more detailed explanation on FCEs.
Figure 3.19. Example FCE for a LOS region on West Green.
Chapter 4

Channel Modeling

4.1 Introduction

In this chapter we discuss the channel modeling results developed from the campus area wireless channel measurements. The chapter begins with the data processing techniques that need to be carried out to obtain important channel impulse response features, such as the separation of valid multipath components from noise. We then provide statistical results for the channel parameters described earlier in Chapter 2. This is then followed by a description of the channel models developed for the campus environment in the 5 GHz band for a 50 MHz channel bandwidth, with additional channel models provided for smaller bandwidths.

4.2 Data Pre-Processing

This section describes the data pre-processing steps required to convert the “raw” data collected from the channel sounder, to an ASCII format that can be used for developing subsequent channel models. The pre-processing consists of three steps: Data Conversion, which involves translating the proprietary format sounder data into readable ASCII data; Noise Thresholding, to separate multipath echoes from noise impulses; and Multipath Thresholding, done to enable a tradeoff between implementation complexity and an accurate description of the mobile channel.
4.2.1 Data Conversion

As described earlier in Chapter 3, a laptop was used to collect the receiver output data during our measurement campaign, with a new data file recorded for each segment of the measurement. This log file is in a proprietary .rap format, which mainly has some header information followed by binary data. This data is in a format which is not interpretable by the primary computational software, MATLAB®, used in our channel model development. The “Chameleon” software package, provided by the sounder manufacturer BVS [21], converts the .rap file to a .out file, which is in ASCII file format. The generated output file can then be read by MATLAB using the `csvread` command. This file is made up of many rows (records) with each record indicating the channel information (magnitude, phase, I values, Q values and RSSI) and other relevant information (like RTC date and time, GPS date and time, position information, and so on) collected at any particular time instant. Figure 4.1 shows a screen shot of the Chameleon application software that we made use of for data conversion [17].
This software is able to convert only one log file at a time. One can choose the option of having a header record as well as choosing a delimiter for each field in a record. The fields we have used during data translation are: Magnitude (dBm), Phase (radians) and received signal strength information (RSSI) (dBm). Table 4.1 illustrates the output data file format for the selected fields of an arbitrary $n^{th}$ record [17].

**Table 4.1.** Output data file format for the $n^{th}$ record.

<table>
<thead>
<tr>
<th>Magnitude (dBm), 1$^{st}$ Sample, $n^{th}$ record</th>
<th>Phase (radians), 1$^{st}$ Sample, $n^{th}$ record</th>
<th>Magnitude (dBm), 2$^{nd}$ Sample, $n^{th}$ record</th>
<th>Phase (radians), 2$^{nd}$ Sample, $n^{th}$ record</th>
<th>…</th>
<th>Magnitude (dBm), 1020$^{th}$ Sample, $n^{th}$ record</th>
<th>Phase (radians), 1020$^{th}$ Sample, $n^{th}$ record</th>
<th>RSSI, $n^{th}$ record</th>
</tr>
</thead>
</table>
Each field selected represents a column, with magnitude and phase alternating for each output record. Here each sample in a record represents the correlated output value obtained at a particular delay for half chip intervals. Hence in order to obtain an unprocessed or a raw power record, the magnitude samples need to be combined (concatenated) for values ranging from 1 to 1020 for a given record. Similarly a phase record can be obtained by combining all phase samples ranging from 1 to 1020 in a particular record. RSSI is recorded once every record. Each record makes a single PDP.

4.2.2 Noise Thresholding

As mentioned, the magnitude samples in a particular record need to be combined to obtain a power record. Figure 4.2 shows an unprocessed power record collected at one of the test locations on College Green, for a 50 Mcps chip rate.

![Example power record for 50 Mcps after using the Chameleon Software.](image)

**Figure 4.2.** Example power record for 50 Mcps after using the Chameleon Software.
The sounder uses a 255-chip length PN sequence and a sampling rate of 100 MHz. Hence when we use a chip rate of 50 Mcps and a bandwidth of 50 MHz, we have 2 samples per chip. Thus for a chip length of 255 chips, this corresponds to 510 samples. But as can be seen in the figure above, we have a total of 1020 samples, with the last 510 samples being a replica of the first 510 samples. This can be verified by observing the values of samples 4 and 514 in Figure 4.2. This is also true for the phase record. The reason the sounder outputs 1020 samples is due to the fact that for a 25 Mcps chip rate, we have 4 samples per chip and thus a total of 1020 samples for the PN sequence. When using 50 Mcps, the sounder just buffers the additional 510 samples. Hence we actually use only the first 510 samples for both records (power and phase).

Once this is done, the next step involves applying a noise threshold algorithm, since, as is so often the case, the samples that make up a power record are affected by noise, mainly due to the receiver electronics. The algorithm we used is adopted from [23], which involves applying a noise threshold to achieve a constant false alarm rate (CFAR) that is not dependent on the signal to noise ratio. This thermal noise can be assumed to be Gaussian, with the noise amplitude having a Rayleigh distribution [23]. Hence the probability that this noise amplitude exceeds some value \( n_0 \) can be written as

\[
P(n_0) = \exp\left( -\frac{n_0^2}{2\sigma_n^2} \right),
\]

where \( \sigma_n^2 \) is the noise variance, which can be estimated from the median value \( \sigma_m \) of all noise samples. The median value \( \sigma_m \) can be obtained by letting \( n_0 = \sigma_m \) and equating the probability in (4.1) to 0.5. Hence \( \sigma_n \) is related to \( \sigma_m \) by

\[
\sigma_n \approx 0.849\sigma_m.
\]
With the CFAR algorithm, the probability of making a wrong decision, i.e., of choosing any noise sample to be a valid multipath echo, was chosen to be one out of all the samples in a given power record. This equates to \[ \frac{1}{510} \approx 2 \times 10^{-3} \] in our case. In equation form, this probability corresponds to

\[ P(\eta) = \exp\left(-\frac{\eta^2}{2}\right), \tag{4.3} \]

with \( \eta \) being a constant. Using the CFA probability, we have \( \eta \approx 3.526 \). Finally, the CFAR method yields a noise threshold, \( N_{ih} \) set to

\[ N_{ih} = \eta \cdot \sigma_n. \tag{4.4} \]

Hence in order to determine \( N_{ih} \), we first must estimate the noise median value and then find the noise standard deviation. This process first requires us to separate the actual noise samples from the multipath echoes. This can be readily achieved by selecting a threshold, below which all samples can be considered to be attributable to noise. We select a threshold of 25 dB; thus all samples that are below 25 dB of the maximum value in a power record are segregated as noise samples, from which we then compute \( \sigma_m \) and then \( N_{ih} \) using (4.2), (4.3) and (4.4). All samples that are above \( N_{ih} \) are considered valid multipath echoes and those below \( N_{ih} \) (which are assumed noise impulses) are set to a minimum value of -130 dBm. This procedure is repeated for all power records gathered on campus. Figure 4.3 shows an example power record, before and after the noise thresholding algorithm is applied. As can be seen, after applying a 25 dB noise threshold, the multipath echoes are separated from the noise impulses.
4.2.3 Multipath Thresholding

The noise thresholding algorithm eliminates inherent noise samples, thus outputting only the genuine multipath components. Before applying a multipath threshold, these samples must be combined vectorially (which is made possible by using a sample’s power and phase information), based on the channel bandwidth being used, to obtain a PDP. As mentioned in Chapter 2, for a 50 MHz channel bandwidth, the chip time corresponds to 20 nsec, with each of the delays being an integer multiple of the chip time. For our sounder sampling rate of 100 MHz, we have a multipath component arriving within a delay span of 10 nsec at the receiver. Hence for a 50 MHz bandwidth, we need to combine 2 adjacent samples to obtain each chip sample in the PDP. This process first involves converting the power samples in a power record to voltage samples,
and then creating $I$ and $Q$ samples using the voltage and phase information of the respective samples. We then add the respective $I$ and $Q$ samples that lie within a chip, to obtain resultant $I$ and $Q$ values for that chip. The power collected in that chip can then be obtained by taking the squares of resultant $I$ and $Q$ values and adding them. Hence for a 50 MHz channel bandwidth; we have a PDP length of 255 chips, after combining 2 samples per chip. As a result, when we develop channel models for different bandwidths (as seen later in Section 4.5), the number of samples that need to be combined for the PDP varies based on the bandwidth.

The above algorithm produces our PDPs, which show distribution of multipath components at different delays. These multipath components can be strong, medium or weak in intensity, based on the path taken by them to reach the receiver. The weaker multipath components normally do not contribute significantly towards cumulative energy (described later) of a wireless channel. As discussed earlier in Chapter 2, we model the channel impulse response as a tapped delay line. To keep the model’s implementation complexity reasonable (as well as accurate), we select a threshold to eliminate the weaker multipath impulses. This threshold is known as a multipath threshold $M_{Th}$. All multipath echoes in a PDP that are below $M_{Th}$ from the main (strongest) multipath component are discarded. Typical values of $M_{Th}$ used by researchers are 20 dB [12] and 25 dB [17]. We selected $M_{Th}$ to be 25 dB; hence in a given PDP, any multipath that lies below 25 dB of the main multipath component is eliminated and set to a minimum value of -130 dBm.
4.3 Channel Parameter Extraction

After applying the multipath thresholding algorithm, the processed PDPs are ready for extraction of important channel parameters. The parameters that are extracted are classified into two types: a) parameters obtained directly from PDPs; and b) parameters obtained after applying a Fourier transform to the PDPs.

4.3.1 Parameters Extracted from PDPs

As seen in Chapter 2, multipath delay spread quantifies the spreading of a received signal in time. The parameters related to delay and other parameters obtained from PDPs are described below.

1. *Mean Excess Delay (MED) (μτ):* The mean excess delay gives the mean value of the energy delay of a PDP. It can be obtained from the normalized first moment of a PDP. In equation form, this can be written as [3]

\[
\mu_\tau = \frac{\int_0^\infty \tau \phi_g(\tau) d\tau}{\int_0^\infty \phi_g(\tau) d\tau} \approx \frac{\sum k \alpha_k^2}{\sum k \alpha_k^2},
\]

(4.5)

with \( \phi_g(\tau) \) being the PDP and \( \int_0^\infty \phi_g(\tau) d\tau \) the normalization factor. The approximation here pertains to multipath echoes at discrete delay values.

2. *Root Mean Square Delay Spread (RMS-DS) (στ):* The root mean square delay spread can be defined as the square root of the normalized second central moment of a PDP and thus gives the RMS value of a PDP’s energy spread in delay. It is given by [3]:
\[ \sigma_\tau = \sqrt{\int_0^\infty \frac{(\tau - \mu_\tau)^2 \phi_\tau(\tau) d\tau}{\int_0^\infty \phi_\tau(\tau) d\tau}} \approx \sqrt{\frac{\sum_k \tau_k^2 \alpha_k^2}{\sum_k \alpha_k^2} - (\mu_\tau)^2} \quad (4.6) \]

As can be seen from the above equation, the RMS-DS is dependent on the mean excess delay. The RMS-DS is often used to characterize the dispersive nature of a wireless channel. Due to this fact, we make use of RMS-DS values (as described later), in developing channel models. Figure 4.4 shows the distribution of RMS-DS values, obtained over all PDPs that were gathered on campus. This bi-modal distribution of RMS-DS is similar to the one seen in Figure 3.18 for all PDPs collected on the West Green, indicating the presence of 2 distinct propagation conditions viz. LOS and NLOS regions. We select an RMS-DS threshold \( \sigma_{\tau_{th}} \) of 950 nsec to separate PDPs into these regions, thus enabling us to build separate models for the LOS and NLOS regions. All PDPs that have RMS-DS less than \( \sigma_{\tau_{th}} \) are classified as LOS PDPs, and those PDPs with RMS-DS above \( \sigma_{\tau_{th}} \) are classified as NLOS PDPs.
Figure 4.4. RMS-DS distribution for measurements on Ohio University campus.

Table 4.2 lists the RMS-DS statistics for a 50 MHz channel bandwidth for the OU campus area, with the maximum, minimum and mean values of RMS-DS listed for LOS and NLOS regions. Approximate values of coherence bandwidth of the channel $BW_c$ (obtained by taking the reciprocal of RMS-DS) are also given. More accurate values of $BW_c$ can also be obtained, as will be shown in Section 4.3.2.

<table>
<thead>
<tr>
<th>Statistic</th>
<th>LOS $\sigma_t$ (nsec)</th>
<th>$BW_c$ (MHz)</th>
<th>NLOS $\sigma_t$ (nsec)</th>
<th>$BW_c$ (MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Max</td>
<td>949.9</td>
<td>1.052</td>
<td>2427.6</td>
<td>0.412</td>
</tr>
<tr>
<td>Min</td>
<td>5.1</td>
<td>196</td>
<td>951.3</td>
<td>1.05</td>
</tr>
<tr>
<td>Mean</td>
<td>341.1</td>
<td>2.93</td>
<td>1503.8</td>
<td>0.665</td>
</tr>
</tbody>
</table>
3. **Maximum Excess Delay** ($W_M$): The maximum excess delay of a PDP can be defined as the time delay over which the multipath energy falls to some value $X$ dB below the maximum multipath component energy [2]. In other words, it is the delay difference between the last and first multipath components ($W_{M2} - W_{M1}$) that are above the energy threshold.

4. **Delay Window** ($W_D$): The delay window is defined as the width of the middle portion of the PDP that contains $x\%$ of the total energy in that PDP [3]. Figure 4.5 illustrates a PDP with its MED, RMS-DS, $W_M$ ($X=20$ dB) and $W_D$ ($x=95$) values listed as shown.

![Figure 4.5. Example PDP showing MED, RMS-DS, $W_M$ ($X=20$ dB) and $W_D$ ($x=95$).](image)
5. Phase ($\phi_k$): In addition to obtaining the power information on multipath components, we also gather the phase variations, yielding a phase record. The received phase is composed of all the terms discussed before in Section 2.4, given by (2.10). The phase varies rapidly from $-\pi$ to $\pi$ radians for distance changes of the order of a wavelength. This phase for all components can be assumed to have a uniform distribution.

4.3.2 Parameters Extracted from Fourier Transform of PDPs

1. Channel Transfer Function (CTF): As seen in Chapter 2, we obtain the channel transfer function by taking the Fourier transform of an Impulse Response Estimate (IRE). The IRE can be obtained by taking the amplitude (square root) of a PDP on which the multipath threshold has already been applied, and combining that with the phase record. In equation form, the CTF can be expressed as:

$$H(f; t) = F\{h(\tau; t)\} = \int_{-\infty}^{\infty} h(\tau; t)e^{-j2\pi f\tau} d\tau$$  \hspace{1cm} (4.7)

Since the IREs are discrete in time (delay), we make use of the Fast Fourier Transform (FFT) algorithm to compute the channel transfer function. Consequently the transfer function obtained is discrete in the frequency domain, having “complex amplitudes” at distinct frequency points.

2. Frequency Correlation Estimate (FCE): The coherence bandwidth can be computed using frequency correlation functions (FCFs) that are obtained by taking the expectation of the channel transfer function and its conjugate. In this method, the Wide Sense Stationary Uncorrelated Scattering (WSSUS) approximation is used, since the
coherence bandwidth is roughly the width of the spaced-frequency, spaced-time correlation function, with time separation set to zero. This approximation assumes that the multipath variations at two separate delays are uncorrelated and the channel impulse response is wide sense stationary (WSS) in time [24]. However, when correlated scattering is observed between neighboring multipath components (such as is often the case in our campus measurements), WSSUS can no longer be assumed. Hence as an alternative, an accurate estimate of the coherence bandwidth can be obtained using the frequency correlation estimate (FCE). The FCE can be obtained using the formulas [24]:

\[
FCE = \frac{\gamma_H(a_{\text{ref}}, a_i)}{\sqrt{\gamma_H(a_{\text{ref}}, a_{\text{ref}})\gamma_H(a_i, a_i)}}
\]

(4.8)

where \( \gamma_H(a_{\text{ref}}, a_i) \) is the cross correlation between complex amplitudes at different discrete frequencies within the transmission bandwidth with the time variations of the complex amplitude at the reference frequency. The quantity \( a_i \) is the complex amplitude at frequency index \( i \), \( a_{\text{ref}} \) is the complex amplitude at the reference frequency, and \( j \) indexes time. Figure 4.6 shows an example FCE for the LOS and NLOS regions with transmission bandwidth of 50 MHz. As expected, the LOS FCE is wider than that of the NLOS region. For a correlation coefficient of 0.5 and 0.37, the correlation bandwidth is approximately 13 MHz and 15.6 MHz respectively for the LOS region. For the NLOS region, for a frequency separation of approximately 8.6 MHz, the correlation falls to 0.37. In this case, due to the lobed nature of the FCE, it is difficult to determine bandwidths for correlation values that are above 0.4.
3. **Doppler Spread** ($f_D$): As defined earlier in Chapter 2, the Doppler spread specifies the amount of spectral spreading that is acquired by a transmitted tone when it traverses the channel. Rewriting the equation 2.2, $f_D$ is given by [2]:

$$f_D = \frac{\nu \cos(\theta)}{\lambda}$$  \hspace{1cm} (4.9)

For a carrier frequency of 5120 MHz, $\lambda$ corresponds to approximately 6 cm. The maximum velocity of the moving cart was approximately 1.5 m/s. This gives a maximum Doppler shift of 26 Hz. In our case, since the transmitted signal bandwidth (50 MHz) is much greater than $f_D$, the campus channel can be assumed to be slowly fading. An approach to obtain $f_D$ is to measure the width of the channel transfer function $H(f;t)$ for each multipath component or tap. In this case, $H(f;t)$ is obtained by taking the

![Figure 4.6. Frequency Correlation Estimates for LOS and NLOS regions.](image-url)
Fourier transform of $h(\tau;t)$ with respect to the time variable $t$, rather than the delay variable $\tau$ [17]. However, since the update rates of the sounder are slow in comparison to $f_D$, we could not use this approach in measuring the Doppler spread.

4.4 Development of Channel Impulse Response Model

In this section we describe the channel models developed for the Ohio University campus in the 5 GHz band. We begin by providing a brief description of the tapped delay line channel model and the method used to calculate the number of channel taps. We then provide channel models in the form of probability of tap existence, channel tap energies and finally the tap amplitude distributions for 50 MHz transmission bandwidth channel.

4.4.1 Tapped Delay Line Channel Model

As seen commonly, the channel models that we have developed are statistical in nature. The channel can be represented in the form of a tapped delay line model, which is composed of different delay elements and random tap weights. For convenience, this model (seen before in Chapter 2) is shown again in Figure 4.7, where $x_k$ is the $k^{th}$ input symbol, $y_k$ is the $k^{th}$ output symbol, the $\tau$ 's denote delays, and the $h$ 's are the time variant impulse response complex amplitudes [25].
4.4.2 Calculation of the Number of Taps

We calculate the number of channel taps (L) based on the mean value of RMS delay spread. Since the RMS-DS varies for LOS and NLOS regions, we calculate the number of taps separately for both of these regions. As seen in Table 4.2, we have computed the maximum, minimum and mean values of RMS-DS over all PDPs on the Ohio University campus. Using the maximum or minimum values of RMS-DS would mean that the channel model is based on a small set of PDPs. The mean value of RMS-DS is a common way of setting L [17]. The number of taps is computed using the formula: [17]

\[ L = \left\lceil \frac{\text{mean}(\text{RMS-DS})}{T_c} \right\rceil + 1 \]  \hspace{1cm} (4.10)

where \( T_c \) is the chip duration and \( \left\lceil x \right\rceil \) denotes the ceiling function, the smallest integer greater than or equal to \( x \). For the LOS region, \( L_{\text{LOS}} \) is 19 and for the NLOS region, \( L_{\text{NLOS}} \) is 77!
4.4.3 Probability of Tap Existence Estimation

As seen previously in equation (2.8), each multipath echo has a tap persistence process associated with it, which takes values 1 or 0 for the presence or absence of multipath components, respectively. A Markov chain is used to model this tap persistence process [17]. It is composed of 2 matrices viz., the transition matrix (TS) and the steady state matrix (SS). For our 2-state chain, TS specifies the probability of a tap going from one state to itself or to the other state. For our CIR form, we have 2 states: the “on” state \( (z_k(t) = 1) \) and the “off” state \( (z_k(t) = 0) \). Hence for any channel tap, TS is a \( 2 \times 2 \) matrix. The SS matrix (here a vector) specifies the percentage of time any multipath echo is present \( (z_k(t) = 1) \) or absent \( (z_k(t) = 0) \). Thus for any multipath echo, SS is a \( 2 \times 1 \) matrix. To determine the probability of tap existence for a given tap, we employ a threshold of 25 dB below the main tap. Figure 4.8 illustrates the probability of occurrence for a given tap versus the tap index. The probability of tap existence captures the “medium scale” time variant behavior of the channel, whereas the random amplitude fading \( (\alpha’s) \) constitute the small scale fading variations. As can be seen, the probability of tap existence decreases as tap index increases. This happens because the higher indexed taps persist only for a short duration. Also, for the NLOS region, for higher indexed taps, this probability is higher than in the LOS region.
4.4.4 Tap Energies

The energy associated with each multipath echo is the aggregate from a sum of different “rays” that arrive at the receiver within the delay bin of a given tap. Once again we make use of our empirical data to compute the channel tap energies. We first compute the average energy associated with each tap. For this we normalize the total energy in a given PDP to 1 and then consider only the energy of a given tap in that PDP which is a valid multipath component \( z_k(t) = 1 \). The procedure is repeated for all PDPs (LOS or NLOS). The average energy for any given tap is then computed by dividing the sum total of energy gathered over all the PDPs for that particular tap, by the sum of number of PDPs that the tap was a valid multipath component. The above procedure is repeated for all the taps. The cumulative tap energy is then obtained by multiplying the average tap
energy by the tap steady state probability of being “on.” Figure 4.9 illustrates the cumulative tap energy versus the tap index for both the LOS and NLOS regions. As can be seen for the LOS region, most of the channel energy (95%) is accumulated within the first 9 taps, whereas for the NLOS region, as expected, it takes larger number of taps (62) to accumulate the same amount of energy.

![Figure 4.9. Cumulative tap energy vs. tap index for LOS and NLOS cases.](image)

### 4.4.5 Tap Amplitude Distributions

In this section we provide the channel model parameters for both the LOS and NLOS regions for a 50 MHz bandwidth. These amplitude parameters were introduced in Chapter 2. Tables 4.3 and 4.4 show the fading tap amplitude parameters for LOS and NLOS regions, respectively. The parameters considered here are the Weibull shape factor ($b$) and the channel tap energy, which together can be used to describe a Weibull fading
amplitude model. An alternative distribution parameter in the form of Nakagami $m$ factor is also provided. In addition, the tap steady state probability of being “on,” $P(z=1)$ and the transition probabilities, $P(1 \rightarrow 0)$ and $P(0 \rightarrow 0)$ are also shown.

### Table 4.3. Amplitude statistics and persistence process parameters for LOS regions, 50 MHz bandwidth.

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<tr>
<th>Tap Index</th>
<th>Weibull Shape Factor ($b$)</th>
<th>Tap Energy</th>
<th>Alternative Distribution Parameter (m) (Nakagami)</th>
<th>$P(z=1)$</th>
<th>$P(1 \rightarrow 0)$</th>
<th>$P(0 \rightarrow 0)$</th>
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### Table 4.4. Amplitude statistics and persistence process parameters for NLOS regions, 50 MHz bandwidth.

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<th>Weibull Shape Factor ($b$)</th>
<th>Tap Energy</th>
<th>Alternative Distribution Parameter (m) (Nakagami)</th>
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</tr>
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</table>

Although Rayleigh and Rician models are fairly popular, the fading tap amplitude distributions we obtained were found to best fit the Weibull and Nakagami distributions [26]. As mentioned in Chapter 2, the Weibull shape factor denotes fading severity, with a value of $b = 2$ indicating Rayleigh fading and a value of $b < 2$ indicating worse than Rayleigh fading. Figures 4.10 and 4.11 show the probability density functions and the measured data histograms for LOS case, 1st tap and NLOS case, 5th tap. As can be seen, the Weibull, Nakagami and Rician distributions provide good fits for these tap amplitude distributions. All these fits are maximum likelihood fits.
Figure 4.10. Fit to fading amplitude data for LOS case, 1st tap.

Figure 4.11. Fit to fading amplitude data for NLOS case, 5th tap.
4.5 Channel Models for Different Bandwidths

As seen in Sections 4.2.2 and 4.2.3, for a chip rate of 50 Mcps and bandwidth of 50 MHz, we have 255 chips, when the channel sounder uses a 255 length PN sequence and sampling frequency of 100 MHz. Thus for a 50 MHz bandwidth, 2 adjacent samples need to be combined in order to convert a power record to a PDP. Table 4.5 lists the number of samples that need to be combined to convert power records to PDPs as well as the number of taps (for LOS and NLOS conditions) obtained for different bandwidths. Since the effective chip duration increases as the channel bandwidth decreases, consequently $L$ decreases as a result.

<table>
<thead>
<tr>
<th>Bandwidth (MHz)</th>
<th>Samples to be combined</th>
<th>Number of taps ($L$)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>LOS</td>
<td>NLOS</td>
</tr>
<tr>
<td>50</td>
<td>2</td>
<td>19</td>
</tr>
<tr>
<td>20</td>
<td>5</td>
<td>8</td>
</tr>
<tr>
<td>10</td>
<td>10</td>
<td>5</td>
</tr>
<tr>
<td>5</td>
<td>20</td>
<td>3</td>
</tr>
</tbody>
</table>

Figures 4.12 and 4.13 illustrate the cumulative tap energy for a 20 MHz bandwidth, and the probability of tap occurrence for a 10 MHz bandwidth channel, respectively. As can be seen, the rate of energy accumulation is faster for 20 MHz case in comparison to 50 MHz, as it takes just 4 LOS taps to gather 95% of channel energy, whereas for the NLOS condition it requires 25 taps to do so. For the 10 MHz bandwidth, as can be surmised, the LOS taps persist for only a short time duration in comparison to the NLOS taps.
Figure 4.12. Cumulative tap energy vs. tap index for LOS and NLOS cases [20 MHz].

Figure 4.13. Probability of tap occurrence vs. tap index for LOS and NLOS cases [10 MHz].
Finally, Tables 4.6 and 4.7 list the fading tap amplitude parameters, along with the tap steady state probability for State 1 and transition probabilities for LOS and NLOS regions when the channel bandwidth is 5 MHz. The number of taps $L_{\text{LOS}}$ in this case is 3, whereas $L_{\text{NLOS}}$ is only 8! For the LOS and NLOS regions, all but the first tap show worse than Rayleigh fading. Also, it can be noted that for the LOS region, the first 2 taps gather most of the channel energy.

**Table 4.6.** Amplitude statistics and persistence process parameters for LOS regions [5 MHz].

<table>
<thead>
<tr>
<th>Tap Index</th>
<th>Weibull Shape Factor $(b)$</th>
<th>Tap Energy</th>
<th>Alternative Distribution Parameter $(m)$ (Nakagami)</th>
<th>$P(z=1)$</th>
<th>$P(1\rightarrow 0)$</th>
<th>$P(0\rightarrow 0)$</th>
</tr>
</thead>
<tbody>
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<td>0.9178</td>
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<td>0.0298</td>
<td>0.7644</td>
<td>0.6421</td>
<td>0.2611</td>
<td>0.5319</td>
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</table>

**Table 4.7.** Amplitude statistics and persistence process parameters for NLOS regions [5 MHz].

<table>
<thead>
<tr>
<th>Tap Index</th>
<th>Weibull Shape Factor $(b)$</th>
<th>Tap Energy</th>
<th>Alternative Distribution Parameter $(m)$ (Nakagami)</th>
<th>$P(z=1)$</th>
<th>$P(1\rightarrow 0)$</th>
<th>$P(0\rightarrow 0)$</th>
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4.6 Model Validation

In this section we provide results from a simulated channel model for a LOS region, for a 10 MHz channel bandwidth. As noted in [27], there are numerous ways to select the number of channel taps for the model, and the mean RMS-DS we used previously is a popular one. Nonetheless, it is clear that using the mean RMS-DS to set the number of taps is limiting. For example, for a 10 MHz signal bandwidth, each delay bin is of 100 nsec duration. Thus, from Table 4.5 we see that for the LOS case, with 5 taps, the maximum possible delay spread attainable from this model is 400 nanoseconds. With the energy per tap decreasing as tap index increases, even this value of 400 nanoseconds is unlikely to be obtained for RMS-DS, and the model is most likely to generate CIRs with smaller delay spreads. From Table 4.2, the maximum RMS-DS found for the LOS case is listed as over 900 nanoseconds. Clearly this can not be reproduced by the 5 tap model, so for our validation, we employ a more complex, but more accurate model. Note that the models based upon RMS-DS are of use for their simplicity, and for their ability to capture some of the “average” channel conditions, but for a more accurate, or “high fidelity” model, the approach we describe in this section is preferred.

To account for the total delay spread of 5000 nsec that can be encompassed by our empirical data, for the 10 MHz channel bandwidth we would require a total of 50 taps. We have thus created a model of this size using the same methods as previously described. Table 4.8 lists the channel parameters for the LOS regions when the entire IRE length is used to develop the channel model. As can be seen, it takes just 3 taps to
gather 90% of the channel energy, and the first 4 taps persist for at least 68% of the time (which is similar to that seen in Figure 4.13).

<table>
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<tr>
<th>Tap Index</th>
<th>Weibull Shape Factor ((b))</th>
<th>Tap Energy</th>
<th>Alternative Distribution Parameter ((m)) ((\text{Nakagami}))</th>
<th>(P(z=1))</th>
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</table>
We make use of the following information from the empirical data to generate simulated CIRs (and thereby compute the RMS-DS distribution) for LOS regions:

1. The number of LOS taps is $L=50$.
2. The shape factor $b$, and the energy associated with each of the taps, $\Omega$, is used to compute the scale factor $a$, using equation (2.20).
3. The worst case correlation matrix is used to obtain the Rayleigh correlation coefficient matrix from the desired synthesized channel’s Weibull correlation coefficient matrix [27]. This Weibull matrix has each of its elements equal to the largest value found among all campus area measurements, with each element indicating the correlation coefficient between the corresponding two taps. For each region, the matrix is of size $L \times L$.

The algorithm proposed in [28] was then used to generate correlated Weibull random variables, used in generating the simulated CIRs. Once these CIRs are generated, equation (4.6) was used to compute the RMS-DS. Figure 4.14 shows the RMS-DS distribution for the empirical data and from the simulations, for LOS regions. The simulations use two models: one employs 5 taps from the mean RMS-DS, and the other higher fidelity model uses 50 taps.

<p>| | | | | |</p>
<table>
<thead>
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<td>1.95661</td>
<td>0.0431</td>
</tr>
<tr>
<td>48</td>
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<td>0.0002</td>
<td>1.66485</td>
<td>0.0383</td>
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<tr>
<td>49</td>
<td>2.5143</td>
<td>0.0001</td>
<td>1.6712</td>
<td>0.0254</td>
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<tr>
<td>50</td>
<td>2.3479</td>
<td>0.0001</td>
<td>1.49709</td>
<td>0.0106</td>
</tr>
</tbody>
</table>
When the mean RMS-DS model is used to generate the simulated CIRs, the maximum RMS-DS attained is around 170 nsec. As noted above, this model generates CIRs with smaller values of delay spreads. In contrast, when the high fidelity model is used to generate the simulated CIRs, with most (95%) of the channel energy lying in the first 4 taps, most of the simulated CIRs’ RMS-DS values lie below 400 nsec. The mode of the distribution generated using the 5-tap model better agrees with the mode of the data than does the mode of the 50-tap model, but the variance of the generated RMS-DS values in the 5-tap case is far too small. We claim that there is a reasonable agreement between the high fidelity model and the data. One reason for the discrepancy between the higher fidelity model and the data is that for the simulated channel model, for simplicity, the tap persistence process was not used in generating the correlated Weibull random variables. Hence all the taps persist for the entire time duration, and this increases the maximum value of delay spread generated by the 50-tap model. The simulated channel model we have provided can be considered a first order approximation for the campus wireless channel. Given the large number of model parameters (two amplitude parameters for each of 50 taps), errors in parameter estimates are likely part of the reason for discrepancy between simulations and analysis. Modeling of this channel is complicated, and clearly, more accurate simulated channel models can be generated when all of the channel information (including tap persistence process) is used, as seen in [27]. We refer readers to [27] for a more comprehensive treatment of this issue, and leave development of more accurate simulations as future work.
Figure 4.14. RMS-DS distribution for empirical data and from simulations (using $L$ based on mean RMS-DS and the entire IRE length), for LOS region, 10 MHz channel bandwidth.
Chapter 5

Summary, Conclusions and Future Work

5.1 Summary

In this thesis we have provided a characterization of the wireless channel in the 5 GHz band for a campus environment. The measurement techniques as well as the way in which the campus regions were segregated into LOS and NLOS regions were also described. We provided a statistical characterization of the channel in terms of the RMS-DS distribution, and its maximum, minimum, and mean values for both regions were given. For frequency domain characterization, we provided example transfer functions and frequency correlation estimates. We defined channel models in terms of the number of taps, their probability of existence, the energy associated with these taps, and their fading amplitude distributions. Finally channel models were also obtained for different signal bandwidths and a channel model validation for our empirical models was provided.

5.2 Conclusions

The RMS-DS distribution obtained is bi-modal which indicates the presence of two distinct regions, viz. LOS and NLOS regions. As expected, the mean value for the LOS case was significantly smaller than the mean value for the NLOS regions. The channel model tap “probability of occurrence” yields interesting insights for modeling the channel accurately; as noted, we model this process by a Markov chain that effectively produces the “on/off” behavior seen in the measured data. The presence of some severe
(“worse than Rayleigh”) fading is not unique to this environment (e.g., [29]), but is atypical, as is the correlated scattering we observed. Some of this is due to the relatively small delay resolution (20 ns), and the presence of large, immobile reflectors (large buildings).

5.3 Recommendations for Future Work

Future work involves gathering additional data from other measurement locations on various Greens that was not covered during this work. Gathering more data would result in increased accuracy of the existing models. As stated previously, these measurements were carried out in late fall and spring seasons, from which the channel models were subsequently developed. Thus we did not consider seasonal effects on our channel models. We could have separate channel models for various seasons, which would then account for different seasonal effects like snow effects during winter and foliage effects in mid spring and summer seasons.

In addition to having mobile links, the campus channel is also likely to have non-mobile or point-to-point links. Hence additional measurements could be carried out using directional antennas that would eliminate some significant multipath and thus help characterize less dispersive channels. Finally, channel measurements can also be carried out into the buildings with the transmitter and receiver in two separate buildings and within the building wherein the transmitter and receiver are located in the same building [30].
References


% -------------- This program generates the various Analytical PDFs for the fading models mentioned in Section 2.6 -------------- %

clear all;
close all;

% Generating an Analytical Rayleigh pdf
a = 0:0.1:3;
p_Rayleigh = 2.*a.*exp(-a.^2);
figure(1);
plot(a,p_Rayleigh,'Color','magenta','Linewidth',2);grid;
xlabel('r');
ylabel('p(r)');
title('Rayleigh Probability Density Function');

% Generating an Analytical Chi-Square pdf
a = 0:0.1:3;
p_Chi = exp(-a);
figure(2);
plot(a,p_Chi,'Color','red','Linewidth',2);grid;
xlabel('g');
ylabel('p(g)');
title('Chi-Square pdf');

% Generating an Analytical Rician pdf
r = 0:0.05:2.5;
K=[-10 10 5 0];
p_Rician=zeros(length(K),length(r));
for i=1:length(K)
    Z = 2.*r.*(10^(K(i)/20))*sqrt(1+(10^(K(i)/10)));
    _0 = besseli(0,Z);
    p_Rician(i,:)=2.*r.*(1+(10^(K(i)/10))).*exp((-r.^2)*(1+(10^(K(i)/10))))-
    (10^(K(i)/10))).*_0;
end
figure(3);
plot(r,p_Rician(1,:),':',r,p_Rician(2,:),'-.kd',r,p_Rician(3,:),'--
rs',r,p_Rician(4,:),'g','Linewidth',2);grid;
xlabel('r');
ylabel('p(r,K)');
title('Rician pdfs with various K factors');
legend('p(r,-10)','p(r,10)','p(r,5)','p(r,0)');

% Generating an Analytical Nakagami pdf
r = 0:0.1:3;
m=[3 1 0.5];
p_Nakagami= zeros(length(m),length(r));
for i= 1:length(m)
  p_Nakagami(i,:) = (2/gamma(m(i)))*(m(i)^m(i)).*(r.^((2*m(i))-1)).*exp(-
                         (m(i).*r.^2));
end
figure(4);
plot(r,p_Nakagami(1,:),'-.kd',r,p_Nakagami(2,:),'--b',r,p_Nakagami(3,:),'g','Linewidth',2);grid;
xlabel('r');
ylabel('p(r,m)');
title('Nakagami-m distribution');
legend('p(r,3)','p(r,1)','p(r,0.5)');

% Generating an Analytical Weibull pdf
r = 0:0.03:2.5;
beta=[6 2 1.5];
p_Weibull= zeros(length(beta),length(r));
for i= 1:length(beta)
  alpha=sqrt(1/gamma((2/beta(i))+1));
  p_Weibull(i,:) = (beta(i)/(alpha^beta(i))).*(r.^(beta(i)-1)).*exp(-(r./alpha).^beta(i));
end
figure(5);
plot(r,p_Weibull(1,:),'-.kd',r,p_Weibull(2,:),'--b',r,p_Weibull(3,:),'g','Linewidth',2);grid;
xlabel('r');
ylabel('p(r,b)');
title('Weibull distribution with different shape factors');
legend('p(r,6)','p(r,2)','p(r,1.5)');
% ------------------ Function to compute RMS Delay spread ------------------ %
% This function computes the RMS Delay spread (RMS-DS) and the Mean excess delay (MED) with its inputs as Power Delay Profile Values and Delay values

% The formula to compute MED and RMS-DS is given by equations 4.5 and 4.6 respectively in Section 4.3.1

% Inputs: tauk: Delay vector of size 1 X k, with k being the number of Taps
%        alphak: Impulse response estimates of size L X k, with L being the number of IREs

% Outputs: rmsDelaySpread: RMS-DS of size 1 X L
%          MED: MED of size 1 X L

function [rmsDelaySpread,MED] = getRMSDelay(alphak,tauk)
    arrDim = size( alphak );
    numRows = arrDim( 1 );
    numColumns = arrDim(2);
    for index = 1:numRows
        num1 = 0; num2 = 0; denom = 0;
        for index1 = 1:numColumns
            num1 = num1 + (alphak(index,index1)^2)*(tauk(index1)^2);
            num2 = num2 + (alphak(index,index1)^2)*tauk(index1);
            denom = denom + (alphak(index,index1)^2);
        end
        t1 = num1/denom;        % First term in RMS-DS equation
        MED = num2/denom;       % Compute the MED
        rmsDelaySpread(index) = sqrt(t1 - (MED^2)); % Compute the RMS-DS
    end

% ------------------ Function to compute Wm ------------------ %
% This function computes the Maximum excess delay (Wm) with its inputs as PDPs, delay values and energy level below max multipath energy

% Wm is defined in Section 4.3.1

% Inputs: AlphadBm: Impulse response estimates of size L X k, with L being the number of IREs and k are the number of echo paths
%        tau: Delay vector of size 1 X k
%        X: Indicates energy (in dB) below the maximum multipath component energy. It is of size 1 X 1
% Output: Wm: Maximum excess delay vector of size 1 X L
function [Wm] = MaxExcessDelay(AlphadBm,tau,X)

LenIRE = size(AlphadBm);
for jj = 1:LenIRE(1)
    maxdBm = max(AlphadBm(jj,:));
    threshold = maxdBm - X; % Set the threshold
    fg=0; % Set the flag value to 0
    for kk = 1: LenIRE(2)
        % To find the first multipath component greater than threshold
        if((AlphadBm(jj,kk)> threshold) && (fg==0))
            fg = 1;
            FirstDelay(jj)=tau(1,kk);
        end
    end
    % To find the last multipath component greater than threshold
    indices = find (AlphadBm(jj,:)> threshold);
    lastElem = indices(end);
    LastDelay(jj)=tau(1,lastElem);
end
Wm=LastDelay-FirstDelay;

% ------------------ Function to compute magnitude, phase and RSSI -------------------- %
% This function can be used to separate the magnitude, phase and RSSI from
% raw data, with its input as raw PDPs generated over a single log file.

% The output file format is explained in detail in Section 4.2.1
function [ RSSI, magdBm, phaseDeg ] = getMagnitudeAndPhase( arr )

    arrDim = size( arr ); % get the dimensions for the entire PDF File
    numRows = arrDim( 1 ); % Total valid rows in PDP
    numColumns = arrDim( 2 ); % Total valid columns in PDP
    for index = 1 : numRows
        RSSI( index ) = arr( index, numColumns );
        tempPDP = arr( index,: );
        magdBm( index,:) = tempPDP( 1:2:( numColumns-1 ) ); % get the Magnitude Vector
        phaseDeg( index,:) = tempPDP( 2:2:( numColumns - 1 ) );% get the Phase Vector
    end

    % Since we use full Span, consider only the first set of PDP samples
    magdBm = magdBm( :,1:510 );
    phaseDeg = phaseDeg( :,1:510 );
% This program plots a time Varying transfer function for a LOS region, College Green.

clear all;
close all;
clc;
load LOS3-D.mat; % This data file contains multipath threshold
% applied on PDPs and RMS-DS values for these PDPs.

[a1,a2]=max(RMSDelSpr);
upRate=4;
numSecs = 1.5;
numPDPs = 0:1/upRate:numSecs;
PDPcnt=length(numPDPs)-1;
Freqs=-25:50/256:(25)-(50/256); % 50 MHz bandwidth
PDPs = PowerdBMN(a2:a2+PDPcnt,:);
sizePDPs = size(PDPs);
for i = 1:sizePDPs(1)
    magnitudeSq = 10.^((PDPs(i,:) - 30)./10);
    IRE=sqrt(magnitudeSq);
    IRE_Freq=fftshift(fft(IRE,256));
    IRE_Freq_Abs=abs(IRE_Freq).^2;
    IRE_Mag_dB(i,:)=10*log10(IRE_Freq_Abs./max(IRE_Freq_Abs));
end

for idel = 1:length(numPDPs)
    for jdel=1:length(Freqs)
        x(idel,jdel)=Freqs(jdel);
    end
end

for jelem=1:length(Freqs)
    for ielem = 1:length(numPDPs)
        y(ielem,jelem)=numPDPs(ielem);
    end
end
z=IRE_Mag_dB;

figure; grid;
surf(x,y,z);
title('Example Time Varying transfer function for a LOS region, College Green');
xlabel('Frequency in MHz');ylabel('Update time in secs'); zlabel('Relative Power in dB');
This program computes the RMS-DS for simulated LOS IREs and compares them with RMS-DS for LOS PDPs obtained from data by plotting their distribution.

```matlab
clear all;
close all;
clc;

load('LOS_10MHz_PDP.mat'); % This data file contains simulated LOS PDPs
load('10MHzRMS-DS_LOSData_Sun.mat'); % This data file contains RMS-DS for LOS PDPs from data
load('LOS_10MHz_PDP_MuRMS_DS.mat'); % This data file contains simulated LOS PDPs based on mean RMS-DS model

losPDPs = LOSPDPs;
sizeLOSPDPs = size(LOSPDPs);
BW = 10*10^6;                   % signal bandwidth
delayBin = 1/BW;                % delay resolution
tauLOS = 0:delayBin:sizeLOSPDPs(2)*delayBin;
[rms_dsLOSPDPs,MEDLOSPDPs] = getRMSDelay(losPDPs,tauLOS);
rms_dsLOSPDPs = rms_dsLOSPDPs./(10^-9);
RMSSimLOS = rms_dsLOSPDPs;
RMSDataLOS = LOS2RMS;

mulosPDPs = muRMS_LOSPDPs;
sizemuloLOSPDPs = size(mulosPDPs);
taumuLOS = 0:delayBin:sizemuloLOSPDPs(2)*delayBin;
[mu_rms_dsLOSPDPs,MEDLOSPDPs] = getRMSDelay(mulosPDPs,taumuLOS);
muRMSSimLOS = mu_rms_dsLOSPDPs./(10^-9);

[yhmuSimLOS xhmSimLOS] = hist(muRMSSimLOS,30);
yhSimLOS xhSimLOS] = hist(RMSSimLOS,30);
yhDataLOS xhDataLOS] = hist(RMSDataLOS,30);

figure(1)
plot(xhmSimLOS,yhmuSimLOS,'-.kd',xhSimLOS,yhSimLOS,'-r*',xhDataLOS,yhDataLOS,'--mo','Linewidth',2);
xlabel('RMS-DS in nsec');
ylabel('Number of Profiles');
legend('From Simulation (using mean RMS-DS)','$\text{From Simulation (using entire IRE)}$','From Empirical Data');
grid on;
```
Appendix B

Radio Frequency Interference Analysis: WLAN Band

Communication Network Services (CNS) handles most of OU’s wireless services. Before training the channel sounder prior to measurements, the campus spectrum was surveyed in the 5 GHz band to make sure that we were not interfering with the campus’ existing wireless network. This also ensured that no external signal was present in this band in which we planned to test. The campus wireless service is though in a band adjacent to our measurement band, the UNII band. Hence we conducted a brief radio frequency interference analysis between these bands.

The maximum power output of the channel sounder is 2 watts. This value accounts for the actual RF output at the transmitter, the small amount of RF cable loss, and the antenna gain of the omnidirectional antenna. With the transmitted signal chip rate of 50 Mcps and a center frequency of 5.12 GHz, the 99% power bandwidth of the signal is approximately 53 MHz. Hence the Power Spectral Density (PSD) of the in band signal is given by the following:

\[
PSD_i = \frac{2 \text{ watts}}{53 \text{ MHz}}
\]

\[
= 3.774 \times 10^{-8} \text{ W/Hz} = -44.23 \text{ dBm/Hz}
\]

For our sounder, the out of band emissions are specified to be greater than 70 dB below the power in-band. Hence, the PSD outside the signal band is

\[
PSD_o = \frac{PSD_i}{10^7}
\]

\[
= 3.774 \times 10^{-15} \text{ W/Hz} = -114.23 \text{ dBm/Hz}
\]
The maximum single sided thermal noise PSD $N_0$ is given by $N_0=kT$, where $k$ is Boltzmann’s constant, equal to $1.38 \times 10^{-23}$ J/K, and $T$ is the temperature in degrees Kelvin. This thermal noise is present in any receiver that is not purposely cooled (“supercooled”) to minimize this effect. At room temperature, $T \approx 290^\circ$K, hence, $N_0 \approx -174$ dBm/Hz. So for this case, the thermal noise floor is approximately 60 dB below the signal level. Worth noting is that this is pessimistic for the out-of-band received signal, in that the 70 dB value of out-of-band attenuation pertains to the signal band edge—that is, for frequencies farther away, the attenuation continues to increase beyond this 70 dB value.

Also deserving consideration is that the value of PSD at the output, $PSD_o$, computed in (B2), is that directly adjacent to the antenna. For receivers distant from the antenna, the signal power level is substantially decreased. From a worst-case perspective in terms of interference to out-of-band receivers, we can use free-space path loss to estimate the signal power level. If any obstructions are present between the sounder transmitter and the out-of-band receiver, the sounder signal level will be even lower.

The free space path loss is given by [16]

$$L_{fs}(d, f) = 20 \log(4\pi df / c)$$

(B3)

where $c$ is the speed of light, $3 \times 10^8$ m/s, $f$ is the frequency of the transmitted signal, and $d$ is the distance in meters. As an example, for $d = 10$ m and $f = 5.12$ GHz, $L_{fs} = 66.63$ dB. This shows that if the distance $d$ between the sounder transmitter and any receiver is greater than or equal to 10 m, the sounder PSD will be more than 6 dB below that of the ambient thermal noise. This in turn means that the out-of-band receiver’s performance will be essentially unaffected by the sounder transmissions.
This benign result is attributable also to the sounder signal format—since it is a spread spectrum signal, the power of the transmitted signal is distributed across a wide band of frequencies, which also makes the signal itself appear like “white” noise to unintended receivers. Finally, for commercial systems, it is unlikely that the receiver quality is such that the actual receiver thermal noise floor can be considered to be equal to -174 dBm/Hz. As a good estimate, for a receiver with a noise figure of $F$ dB, the noise floor is increased by this value, hence if $F=4$ dB for example, the thermal noise floor is increased to -170 dBm/Hz. For commercial wireless local area network (WLAN) equipment that operates in the 5.2 GHz band, this value of $F=4$ dB is within the typical range of 2.4 to 7.4 dB, even though the IEEE specification allows the noise figure to be as large as 10 dB [31].

The IEEE 802.11a standard has 12 non-overlapping channels (5180 MHz-5805 MHz), with a 20 MHz separation between adjacent channels. Since we use a carrier frequency of 5120 MHz and a transmission bandwidth of approximately 53 MHz, the upper band edge frequency of the channel sounder, $f_{c,\text{sounder}}=5146$ MHz. Applying the limiting condition of $d=10$ m, as mentioned above, the free space path loss, $L_{fs}(d,f)\approx 66dB$. The transmitted power used by the 802.11a WLANs at Communication Network Services (CNS), $P_{LAN,Tx}$ was 30 mW $\approx 15$ dBm. Hence the received power at ten meters distance is

$$P_{LAN,Rx} \approx P_{LAN,Tx} - L_{fs}(d,f) \approx -51 \text{ dBm.}$$

Hence the PSD of the received signal
Comparing (B5) with (B2), it can be seen that the WLAN received PSD value lies below the PSD of the sounder for out of band emissions. Due to the low signal levels of the WLANs, we couldn’t use a conventional spectrum analyzer to reliably detect any out of band emissions from our sounder, or from the 802.11a WLAN device in the proximity. Hence as a substitute, we tested our sounder and an actual 802.11a WLAN card and confirmed their operation in the presence of one another.

The WLAN card was connected to a laptop that had a software interface that showed the different available channels in the UNII band, their signal-to-noise ratio (SNR) in dB, and their relative signal qualities (strong, good, fair and poor) periodically. This software enabled qualitative evaluation of WLAN performance via the signal quality descriptors. The accuracy of the quantitative evaluation in terms of SNR is not known; details of the SNR measurement used by the software were unavailable, and in addition, since the WLAN uses an adaptive modulation and coding scheme, a given SNR can correspond to several resulting system performance levels in terms of, for example, error probability. Thus our interference analysis here is not as quantitative as desired. Yet, with both the SNR and qualitative performance measure, we were able to determine relative distances and power levels such that our sounder transmitter did not, to the satisfaction of the OU CNS personnel, interfere with the WLAN performance.

To begin the testing, the WLAN access point was tuned to operate at its lowest frequency, $f_{c,1} = 5.18$ GHz. The sounder’s transmitted frequency was set to our desired testing frequency of 5.12 GHz, and the sounder transmitter’s power level was increased.
in steps from its minimum value until it reached full power (2 watts). Following recommendations by OU CNS personnel, we set the WLAN link distance to approximately 5 m. Although this distance is smaller than the maximum WLAN operating distance, it was suggested as a typical value. The sounder Tx distance from the WLAN equipment was also varied, with a maximum distance separation of 10 m. This procedure was repeated for other WLAN channels with larger center frequencies.

Figure B.1 shows a block diagram of the radio frequency interference testing set up. The measurement results are summarized in Table B.1 below.

![Block Diagram](image)

**Figure B.1.** Block diagram of the radio frequency interference testing set up.

<table>
<thead>
<tr>
<th>WLAN carrier frequency, $f_c$</th>
<th>Sounder State (on/off) and transmitted power, $P_{Tx}$</th>
<th>Distance between WLAN Rx and Sounder Tx, $d_{LAN-Sounder}$</th>
<th>WLAN Performance Comments</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_{c,1}=5.18$ GHz</td>
<td>Off</td>
<td>5 m</td>
<td>Strong SNR of 33 dB</td>
</tr>
<tr>
<td>On; 5 dBm to 20 dBm</td>
<td>5 m</td>
<td>No interference</td>
<td></td>
</tr>
</tbody>
</table>
Hence from Table B.1 we can see that interference is detectable when the WLAN is set to operate on the lower two channels in the UNII band. For these channels, the sounder was operated at full power and the separation distance between the WLAN Rx and the sounder Tx was less than or equal to 10 m. As can be seen, the WLAN SNR is fair to poor for these frequencies and distance separations. There was no detectable interference on the other remaining channels, when the sounder was operated at full power and the separation distance between the WLAN Rx and the sounder Tx was set to 10 m. Hence in order to ensure there was no interference during channel sounding, we coordinated with CNS so that no WLAN access points were set-up to operate in the two lowest-frequency channels. Also it was seen that there was negligible interference on

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Power (dBm)</th>
<th>Distance (m)</th>
<th>SNR (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>5.2 GHz</td>
<td>On; 25 dBm</td>
<td>5 m</td>
<td>Noticeable Interference effect</td>
</tr>
<tr>
<td></td>
<td>On; 25 dBm</td>
<td>5 m-10 m</td>
<td>Good SNR observed</td>
</tr>
<tr>
<td></td>
<td>On; 30 dBm</td>
<td>10 m</td>
<td>Good SNR of 25 dB</td>
</tr>
<tr>
<td>5.22 GHz</td>
<td>On; 33 dBm</td>
<td>5 m</td>
<td>Poor LAN SNR of 8 dB</td>
</tr>
<tr>
<td></td>
<td>On; 33 dBm</td>
<td>10 m</td>
<td>Fair SNR of ~ 15 dB – 23 dB observed</td>
</tr>
<tr>
<td></td>
<td>On; 30 dBm</td>
<td>10 m</td>
<td>Good SNR of ~ 35 dB</td>
</tr>
<tr>
<td></td>
<td>On; 33 dBm</td>
<td>10 m</td>
<td>Fair LAN SNR of ~ 20-25 dB</td>
</tr>
<tr>
<td></td>
<td>On; 33 dBm</td>
<td>10 m</td>
<td>Good SNR of ~ 28 dB</td>
</tr>
</tbody>
</table>
these channels, when distances between sounder and WLAN Rx was much larger than 10 m (e.g., 50 m or more).