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Broadband Point-to-Point Wireless Channels Over Flat Land with low Antennas

by

Masoud Mashhouri

A thesis submitted to the
Faculty of Graduate Studies and Research
in partial fulfillment of the requirements for the degree of
**Master of Applied Science in
Electrical Engineering**

Ottawa-Carleton Institute for Electrical and Computer Engineering
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Abstract

The recent commercial success of WLAN, especially the IEEE802.11 family triggered a lot of interest in using this technology in Internet wireless access everywhere. In metropolitan areas, 802.11 will be mostly used in what is known as “hot spots”. These are public areas with broadband wireless access over relatively short distances (about 100 meter maximum). In rural areas, however, the same technology can be used for point-to-point or mesh networking among small communities. This second application is the subject of this thesis. The objective of this thesis is to study the performance of point-to-point 802.11a commercial cards. Our study show very promising results in terms of high throughputs achieved using hardware available to everyone.

The existing solutions to fixed wireless access require sizeable investments which, makes them less desirable to the communities in rural communities, while using commercial WLAN hardware connected to high-antennas with relatively low-heights provides a more economic solution. The study of this thesis presents a general model of a wireless point-to-point access using the 802.11a technology as the building block of wireless mesh network.

The existing path-loss models have certain limitations that make them limited in terms of applications. In this research different physical models are examined and by developing a detailed form of the two-ray model to include a set of different parameters such as the specific antenna pattern, roughness of the terrain, misalignment of the transmitter and receiver antennas, a propagation model for low antenna-heights is introduced. Finally this model is examined for validity.

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List of Acronyms

Acronym	Equivalent Phrase		
ACI	Adjacent Channel Interference	EIFS	Extended Inter-frame Space
ACK	Acknowledgement	EIRP	Effective Isotropic Radiated Power
AM	Active Mode	ESS	Extended Service Set
AP	Access Point	FHSS	Frequency-Hopped Spread Spectrum
ARF	Auto-Rate Fallback		
ARQ	Automatic Repeat Request	HPBW	Half-Power Beam-Width
ATIM	Ad-Hoc Traffic Indication Message	IBSS	Independent BSS
AWGN	Additive White Gaussian Noise	IFS	Inter-frame Space
BER	Bit Error Rate	IP	Internet Protocol
BSS	Basic Service Set	IR	Infra-Red
CFP	Contention-free Period	LAN	Local Area Network
CP	Contention Period	LLC	Logical Link Control
CSMA/CA	Carrier Sense Multiple Access with Collision Avoidance	MAC	Media Access Control
CSMA/CD	Carrier Sense Multiple Access with Collision Detection	MPDU	MAC Protocol Data Unit
CTS	Clear-to-Send	MSDU	MAC Service Data Unit
DCF	Distributed Coordination Function	NAV	Network Allocation Vector
DIFS	DCF Inter-frame Space	OFDM	Orthogonal Frequency Division Multiplex
DS	Distribution System	OSI	Open System Interconnect
DSSS	Direct Sequence Spread Spectrum	PCF	Point Coordination Function
		PER	Packet Error Rate
		PLCP	Physical Layer Convergence Procedure
		PHY	Physical Layer
		PPDU	PLCP Data Unit

PSDU	PHY Service Data Unit		Time
PIFS	PCF Inter-frame Space	TCP	Transfer Control Protocol
QAM	Quadrature Amplitude Modulation	TSF	Time Synchronization Function
RABR	Receiver-based Auto-Rate	UDP	User Datagram Protocol
SIFS	Short Inter-frame Space	UNII	Unlicensed National Information Infrastructure
SNIR	Signal-to-Noise-and- Interference Ratio	WLAN	Wireless Local Area Network
SNR	Signal-to-Noise Ratio		
SSA	Service Set Area		
TBTT	Target Beacon Transfer		

List of Symbols

Chapter 2		P_{Rmin}	Minimum (acceptable) received power
Symbol	Significance	d	Distance
L_{DATA}	Packet length (Bytes)	f	frequency
T_{DATA}	Time length of data packet	h_b	Transmitter antenna height
T_{SLOT}	Slot time	h_m	Receiver antenna height
CW_{min}	Minimum length of contention window	α	Index of reflection
T_{DIFS}	DIFS time length	α_v	Index of reflection for vertical polarization
T_{SIFS}	SIFS time length	α_h	Index of reflection for horizontal polarization
τ	Radio Propagation time	ϵ	Electric permittivity
τ_i	Processing time for repeater i	ϵ_o	Free space electric permittivity
τ_{tot}	Total processing time	ϵ_r	Relative electric permittivity
P_M	Symbol Error rate (M-QAM)	μ	magnetic permeability
$Q(.)$	Q-function	σ	Electric conductivity
W_i	Bandwidth of carrier i	E_r	Reflected electric field
R_b	Overall bit rate	E_i	Incident electric field
E	Energy	θ_i	Incident angle
N_o	White Noise Spectral Density	θ_r	Reflection angle
Chapter 3		θ_t	Traversed angle
E_b	Bit energy	h_c	Critical height
k	Number of bits / symbol	λ	Wavelength
T_b	Bit time duration	ρ_s	Roughness correction factor
R	Bit rate	σ_h	Terrain roughness standard deviation
L_p	Pathloss	G_{max}	Maximum antenna gain
P_T	Transmitted power	$G(\theta)$	Parabolic antenna gain function
P_R	Received power	$G_L(\theta)$	Normalized parabolic antenna gain function
G_T	Transmitter antenna gain		
G_R	Receiver antenna gain		

Δd	Path difference		height
$\Delta\varphi$	Phase difference	θ	Hill slope
K_f	Space factor at a fixed distance	Δh	Height difference
u_o	Initial arc length	β	Slope of the line connecting the two antennas
u	Arc length	F_t	Focus point (transmitter site)
r	distance	F_r	Focus point (receiver site)
v	v-parameter	A_{tot}	Total loss
$C(v)$	Real part of Fresnel integral	$A_{freespace}$	Free space loss
$S(v)$	imaginary part of Fresnel integral	$A_{reflection}$	Reflection loss
$R_{first-Fresnel}$	Radius of the first fresnel zone	$A_{diffraction}$	Diffraction loss
D_f	Distance where the first fresnel zone is tangent to ground.	Chapter 4	
H_t	Actual transmitter antenna height	R	Bit rate
h_t	Projected transmitter antenna height	N_{BPSC}	Number of coded bits per symbol
H_r	Actual receiver antenna height	T_{FFT}	FFT period
h_r	Projected receiver antenna height	N_{FFT}	Number of FFT points
		T_{SYMBOL}	Symbol duration

Chapter 1

Introduction

1.1 General

The recent commercial success of WLAN, especially the IEEE802.11 family triggered a lot of interest in using this technology. In metropolitan areas, 802.11 will be mostly used in what is known as “hot spots”. These are public areas with broadband wireless access over relatively short distances (about 100 meter maximum). In rural areas, however, the same technology can be used for point-to-point or mesh networking among small communities. This second application is the subject of this research where there are scattered communities around a relatively large city. The purpose of this research is to introduce and study a low-cost point-to-point link using 802.11a technology operating in an Ad-Hoc mode to provide the wireless link. Our field tests have shown promising results in terms of high throughputs achieved using off-the-shelf hardware. An example of such networks is shown in Figure 1.1[next page]. In this network, the broadband links provide wireless connection network to the scattered communities around the city. In this example, the wireless link use 802.11a technology [at 5 GHz frequency band] and the distribution network 802.11g [which support the same data rate [at 2.4 GHz].

Elements of the network: **Wireless Nodes and links**

Figure 1.1 A Wireless Ring network

Previous works related to 802.11 radio coverage such as [1-3] use simulation to model the propagation behavior of the channel and also limit the discussion to the indoors environment. Another study [4] investigates the 5.8 GHz frequency band in the outdoor environment including building penetration losses. The present study attempts to include most factors that limit the transmission speed including the MAC behavior of commercial 802.11 cards.

Promising results from our field tests motivated a thorough study to model the propagation of radio waves at the UNII band. The resultant model provides a tool to predict the performance of the radio links based on the topologic properties of the “coverage” area.

A feasible approach to implement such a service is a mesh network in which the communities are connected by wireless links comprising the backbone. Each of the links connecting the nodes in the mesh must provide a significant capacity either in single hop

or multi-hop. The systems providing the single-hop links are available in the market at high costs. Two major parts of the cost are the real estate and the antenna tower. These systems provide fixed wireless access for fairly long distances, by using higher transmitted power and higher antenna elevations.

A more economic approach to provide such access is using 802.11a commercial cards at low antenna elevations in multi-hops. Using such links brings about economic incentive and relatively high capacities, however there are certain restrictions to this approach. The rest of the chapter discusses these issues and how they are related to the present research work.

1.2 Issues Related to 802.11a

The 802.11 technology, in general, provides access within a LAN environment (about 100 meter radius). The 802.11 uses event-based time schedules to provide access to the physical environment. The assumption of a WLAN for the technology plays a significant role in defining the time slots and transactions such as time-outs, and etc. For example, if a user sends a packet from a distance beyond the WLAN defined limit, even if the radio signal is strong, the packet will be discarded due to excessive propagation delay. Chapter two provides the 802.11 media access operations in some details to the extent that is related to this research.

The following summarizes a set of issues of using 802.11a technology to provide wireless links using directional antennas at low heights. The issues are classified in two categories:

1.2.1 Issues Related to the physical environment

A) Range

A typical access point (AP) uses omni-directional antennas for the obvious necessity of providing access to users randomly distributed within the service area. Providing service in small cells does not require substantial amount of transmitted power, and we have experimentally shown that the power provided by commercial cards is sufficient to provide longer distances, but the range is normally predefined and restricted. The access range may be extended by using directional antennas.

Using directional antennas will change the scenario significantly, because it restricts the communications to the link only (to serve our purpose).

B) Propagation issues

As mentioned earlier, the solutions using high towers for wireless links have the advantage of height gain and Fresnel zone clearance at the expense of higher costs (tower, real-estate, etc). In the proposed link using directional antennas at low-heights, the Fresnel zone clearance quickly disappears as distance grows. As may be easily found in the literature, this will result in rate of change in attenuation with distance, from 20dB/decade to 40dB/decade. Severe attenuation at lower heights, limits the length of the link for a fixed target data rate. Therefore, in order to provide higher data rate, a long distance link may have to be broken down into several shorter hops..

C) Multi-hop link issues

Two modes of operation may be conceived for the multi-hop link:

(a) **Repeater mode** and

(b) **Smart mode:**

a) In the repeater mode, each node receives data from one hop and delivers it to next hop.

b) In the smart mode, the connection node may provide access to other links(/hops). Therefore the station must have a supervisory scheme to organize the transactions from all connecting links. Scheduling of these transactions is beyond the scope of this research.

In both of the above modes, a certain processing delay in the reception and transmission will be added to other delays. The delay added by each mode will, however, be different. As we will see in chapter 2, these delays reduce the throughput of the systems significantly.

D) Frequency allocation

For proper operation of multi-hop connections, the links must operate at different frequencies, so the leakage from the front to back does not hamper the communication (note that the front-to-back ratio is not infinite in a practical situation). Also, the wireless hops that comprise the link are not necessarily on a straight line. Along the link, it is possible that any of the directive antennas might receive significant interference from interfering antennas, since no matter how sharp an antenna pattern, the half-power-beam-width of the antenna (for a directional antenna such as a parabolic reflector) mapped into a spatial angle will have a footprint which only grows with distance.

Please note that the definition of “channel frequency” is defined in the 802.11a physical layer specifications; if in the 5 GHz frequency band the lower- and middle sub-band are used, only 8 channels would be available, but if the upper sub-band is also used, 4 more channels become available.

Considering the above issues, a procedure for frequency mapping is necessary. This will be covered in later chapters.

1.2.2 Protocol issues

Two-point Ad-hoc

The Ad-hoc mode assumes that a number of independent terminals communicate with one another without a central supervisory system (e.g. AP). In this case the scenario is simpler; there are only two nodes involved in the link. In fact, the general mesh network, as proposed here, consists of a number of nodes which use two-node connected links (in Ad-hoc mode). The existing protocol may be simplified to accommodate the simpler configuration in order to reduce the overheads, and in turn, increase the throughput.

Parallel operation of links over multiple channels

In order to use the available bandwidth for higher capacities, it is theoretically possible to combine parallel links (defined above) to multiply the transmission rate by up to 8 times (or 12 times if all three 802.11a sub-bands are used). While parallel operation of these channels has no problem in actual networks consisting of a distributed set of nodes, combining the channels at RF stage (to use the same directional antenna) results in serious problems due to RF leakage from channels and non-linear properties of the input low-noise amplifiers, as verified by our experiments [Appendix A]. As a solution to this problem, modification to the MAC protocol may prove useful. This topic has been discussed in future chapters, although it is not the main focus of this research.

1.3 Contributions of this research

In this thesis, the analysis and modeling of wireless links using low antennas have been carried out. The thesis presents a comprehensive analysis of the link in terms of both protocol issues and radio wave propagation phenomena. As a result of this work, a model has been developed to predict the performance of a wireless link using of-the-shelf equipment. Along with the provided model, a model for parabolic antennas is also provided. Using the said models, phenomena such as antenna misalignments and wind vibrations have been provided. As another part of the model, the diffraction loss for a sloped terrain has been formulated and integrated into the model.

1.3.1 Publication based on the current research

As one of the results of this research a paper under the title of “Measure performance of point-to-point 802.11a” has been published and presented at the WOCN 2005 conference (March 6-8, 2005).

1.4 Organization of the thesis

In chapter 2, the reader is familiarized with the general backgrounds of the wireless LAN technology, and more specifically, 802.11a networks and also different techniques used in the network.

In chapter 3, we present detailed analysis of the radio wave propagation phenomena that affects the operation of the radio link. We also present a model of directional antennas. Using the results of this chapter, different factors of radio wave attenuation are predicted and the result is used in the link model provided in chapter 4.

.

In chapter 5, different aspects of the model are examined and a summary of the field test results are presented. [Samples of the field tests are presented in Appendix A]

Chapter 2

Review of 802.11 Networks

2.1 Introduction

In this chapter, a brief description of 802.11 MAC layer and 802.11a physical layer are presented. This chapter focuses on providing the protocol details to the extent required by the implementation of the proposed point-to-point link. Due to the change in the physical environment, certain restrictions will affect the operation of the link and its performance. Also, the use of repeaters and their effect on the link performance will be studied.

2.2 General

The IEEE 802.11 working group has standardized the WLAN technology into an affordable and ubiquitous solution for office networking over a number of years and achieved considerably higher data rates at each step. The 802.11 networks have been designed in full compatibility with TCP/IP network structure. Figure 2.1 shows the TCP/IP network as compared to OSI open system model.

Figure 2.1 TCP/IP and OSI network models

2.3 802.11 standards [6]

The draft edition of 802.11 came to existence using 2.4 GHz frequency band. It included a Physical Layer (PHY) which interfaces with the chosen physical environment and a Media Access Control (MAC) layer. Table 2.1 shows the structure of draft 802.11 network:

Table 2.1 802.11 Data Link Layer and Different PHY Layers

DATA LINK LAYER	LLC				
	MAC				
PHYSICAL LAYER	FHSS	DSSS	IR	802.11b IR/DSSS	802.11a/g OFDM

The MAC layer is the same for all forms of 802.11(with small variations), but it is the PHY that really defines the letter assignments.

2.4 Description of the MAC layer [6]

Media Access Control (MAC) layer, together with LLC, constitutes the 802.11 Data Link Layer. This layer interfaces the LLC to the physical layer. The main functions of the MAC layer are as follows:

1. Accessing the physical media (addressing, access coordination and frame check sequence generation).
2. Joining a network
3. Providing authentication and privacy

Some other functions included in MAC layer are *fragmentation* of the upper layer packets, *flow control* and *error correction*.

2.5 802.11 modes of operation [7]

Two modes of operation have been defined in 802.11: Infrastructure mode and Ad-hoc mode. MAC layer operation of the 802.11 applies to all a/b/g versions with little variation.

In each of the above modes service sets are defined as follows. The 802.11 network has three service sets: Basic Service Set (BSS), Independent Basic Service Set (IBSS) and another service set sometimes called extended service set (ESS). The basic service set comprises of an access point (AP) and the terminals within its coverage area among which the AP manages the interactions. The communication is only between a terminal

and the AP and the terminals do not communicate with one another. The IBSS, on the other hand, is a system in which there are only terminals which work independently and since there is no AP, the terminals manage their own communications within the network (Figure 2.2).

Figure 2.2 BSS, IBSS and ESS networks

Two (or more) BSS's can be connected via a distribution system (DS). A DS may be a wired network such as 802.3 or any other standard. In this case the service set is called an extended service set (ESS) [Figure 2.2]. It should be noted that a DS poses no specific restriction on the BSS's. The IBSS is more often called the Ad-Hoc network and as the name implies it is more suitable for ad-hoc situations. Therefore two modes of operations for 802.11 networks are defined: Infrastructure Mode (using BSS) and Ad-Hoc Mode. Since, unlike a wired network environment in which CSMA/CD protocol is used, collision detection is not convenient [note that the WLAN card either sends data or "listens to" the radio channels]. In Ad-Hoc mode a similar protocol with a more conservative approach is utilized: CSMA/CA (CA stands for Collision Avoidance) in which, the stations always "listen" to the medium for a combination of a predefined and

random time period and start transmission. In the Infrastructure mode, since every transaction is coordinated by the AP collisions do not happen.

2.6 Medium Access [7]

The existing 802.11 technology provides two modes of operation: Ad-Hoc mode and Infrastructure mode. In the 802.11 two functions are defined: The Point Coordination Function (PCF) and Distributed Coordination Function (DCF), for infrastructure mode and the Ad-Hoc mode, respectively.

2.6.1 Definitions of Access methods

Point Coordination Function (PCF) which is used in the BSS/ESS service set and operates based on polling (by AP). [This function is an optional feature.]

Distributed Coordination Function (DCF) in which stations can make decisions independently based on CSMA/CA algorithm.

In our study we are only interested in DCF on a single-hop, that is, a peer-to-peer communication between two stations.

2.6.2 Operation of the PCF [7]

This mode can only be used in the infrastructure network configuration. In this mode the AP polls the associated terminals in the BSS. All stations within the coverage area which have the same SSID (service set ID) are assigned specific times. The system alternates between two periods: a contention period (CP) and a contention-free period (CFP). During the CFP the medium is controlled by the access point. The AP coordinates all the

transactions with the service set area (SSA), by polling the terminals. The standard defines three types of frames: Management frames, Control frames and Data frames.

The purpose of the management frames are listed in Table 2.2 [11]:

Table 2.2 Types of frames and the respective functions

Type of frame	Function of the frame
Management frame	<i>Station Association, Re-Association and De-Association with the AP, Authentication and De-authentication</i>
Control frame	<i>Handshaking during CP, to end the CFP and Acknowledgement during the CP</i>
Data frame	<i>Transmission of Data both for CP and CFP (may be combined with ACK during CFP)</i>

Polling is done sequentially even if there is no data to be transmitted. When the communication process begins, as a first step, the AP waits for a PIF period and if the medium is clear (no one is sending), it starts transmission. At this time, as seen in the Figure 2.3, a beacon is sent that includes a variety of information, including the SSID of the network, the Timestamp and Beacon interval [Re: Sec. 7.2.3.1, IEEE1999]. Then a “Data/Poll (*terminal(i)*)” is sent after an SIF period, which can only be responded to by station “*i*”, with an acknowledgement packet (ACK). Then, following another SIF period, the sequence continues to the next station. If a station is polled while no transaction is ready, then a *null* packet will be exchanged, instead of the data content. If for any reasons, the packet is not acknowledged by the receiver (e.g. receiver not receiving the packet or the ACK is lost) the AP waits a PIF before passing to the next station (Figure 2.3).

Figure 2.3 DCF transmission timing

- The receiver sends an acknowledgement to the transmitter after predefined time period (SIFS). If the transmitter receives the (positive) acknowledgement from the receiving end, the next cycle begins. If not, the transmitting station goes into an extended randomly chosen back-off period (EIFS) in order avoid simultaneous transmission with other stations. The time periods DIFS (DCF Inter-frame Space), SIFS (Short IFS) and EIFS (Extended IFS) are shown in Figure 2.4.

2) RTS/CTS method

This method is used to avoid or minimize the hidden terminal problem [10]. In this method the transmitter, instead of sending data packets right away, it first sends a shorter frame called Request-to-Send (RTS), and the receiver responds with Clear-to-Send (CTS) frame. The RTS frame has information about the source, destination and duration of the frame. Therefore, any receiving station will realize if the transmission belongs to them or not. Also, if a collision occurs, since the frame is much shorter than the data-containing frame, a shorter amount of time is wasted. For longer data frames, this overhead will be

negligible (Since the collision on the data frames is avoided, this method is, sometimes, called virtual collision detection-Figure 2.5).

Any other station that senses any transmission goes into a predefined back-off state called network allocation vector (NAV). It starts, from the beginning of the sensed transmission to the end of the data frame acknowledgement. Any activity before the expiry of NAV may extend its duration.

Figure 2.4 Exchange of frames in DCF

Despite the fact that the hidden terminal problem may be avoided by RTS/CTS method, this method includes an overhead which has a significant effect on the link throughput, especially when the frame sizes get smaller.

2.6.4 Use of Ad-Hoc Mode in a Point-to-Point Link

As discussed above, the Ad-Hoc mode is designed for networks in which each station is independent and there is no central control station to supervise the interactions among the stations. If a wireless link is implemented by using only two stations in Ad-Hoc mode, the configuration will be simplified. However, it must be noted that the communication on a single channel, based on the protocol itself, is always between two computers at a time and if there are other transactions waiting for access, they have to wait until the channel is clear. In other words, the point-to-point link using the Ad-Hoc protocol is a special case for the Ad-Hoc network in which only two stations exist and no other stations other than the comprising nodes share the channel. Therefore the timing diagrams depicted above directly applies to the point-to-point link.

2.6.5 Transmission of frames [11]

Time synchronization function (TSF) mechanism

In both Basic Access and RTS/CTS methods, the transmission is done in beacon interval time. It start/repeats at times, called Target Beacon Transfer Times (TBTT). At these times, a beacon is sent. According to the 802.11 standard, only one station can send the TBTT. The beacon packet includes the information such as Service Set Identification

(*SSID*) to identify the “name” of the link/network, the clock value for synchronization purposes, and the beacon period (*aBeaconPeriod*). To avoid collision of beacon packets, every station initially backs off for a random value between 0 and twice the contention window value number of slots. Any other station sensing the beacon, will cancel the beacon transmission.

Synchronized by the beacon, the stations go through two sub-intervals in each TBTT: In the Ad-Hoc Traffic Indication Message (ATIM) window, every station “wakes up” into Active Mode (AM). In the rest of the TBTT, all except the transmitting and receiving stations go to power-save mode. The (ATIM) sends out the required parameters to all members of the IBSS based on CSMA/CA. In order to avoid unwanted simultaneous transmission of ATIM packets, a random back-off procedure is followed. ATIM packets are acknowledged by the targeted receiving station.

The 802.11x interfaces with the TCP/IP protocol in the same way as other 802.* protocols; the standard only defines the MAC and a specific PHY (normally with a letter designation (except for the original 802.11!).

As shown in Table 2.1, in the original edition of 802.11, three different physical environments have been considered for operation. In fact, only the DS type has been implemented and it has also survived up to the 802.11b version (although with higher data rates).

Figure 2.5 Fragmentation

2.6.6 Fragmentation [12-13]

Figure 2.6 shows an example of fragmentation of the upper layer frames to accommodate shorter transmissions when the channel is noisy and the data frame is more error-prone. In order to avoid the loss of long frames in case of error/collision, smaller packets are preferred, however this will result in additional overhead since every fragmented frame has to follow the same protocol overheads

The criteria for fragmentation are the channel conditions in terms of available SNR and the mode of transmission at the PHY. According to the standard, in the beginning of transmission of a frame, a fragmentation threshold (normally 800 Octets as used in literature) is chosen and the frame will be divided into fragments of this size which will stay fixed until the second last fragment of the frame is transmitted (the last fragment is obviously less than or equal to the fragment threshold in size). Further discussion on the subject is beyond the scope of this research. However, it should be pointed out, that when the channel is occupied by a terminal to send a fragmented packet, it is not released until all the fragments of the packet are transmitted, or the connection is lost for some reason.

Figure 2.6 802.11a PHY and the frequency subbands spectrum

2.7 Description of the Physical Layer (PHY) of 802.11a [14]

The draft edition of the 802.11 standard and its off-spring 802.11b, together with 802.11g all address the physical layer at the 2.4 GHz frequency band; however 802.11a operates at the 5 GHz unlicensed frequency band known as Unlicensed national Information Infrastructure (UNII). Figure 2.7 shows the RF channeling scheme in the 5 GHz frequency band and the allowed transmitted power [EIRP]. The 802.11a channel definition assigns 20 MHz to each channel. Using OFDM as the modulation technique to transmit data, the space between subcarriers (within the OFDM signal) is 312.5 KHz which allows space for 64 subcarriers, but in order to reduce the adjacent channel interference, only 52 subcarriers are used, of which 48 are for data and 4 are for pilot signals.

Figure 2.7 Channel division and OFDM spectrum

As Figure 2.8 shows, the frequency band is divided into 8 non-overlapping channels. A mask is defined by the standard as in Figure 2.9, such that the adjacent channel interference (ACI) is minimal for normal operation.

Figure 2.8 Estimates of interference based on a typical power mask

In later sections, the formulation of ACI for two immediately adjacent channels using a typical spectral power mask is presented.

2.8 Performance Analysis of 802.11 MAC in Ad-Hoc mode

2.8.1 Throughput of an 802.11 network [15-17]

The nominal 54 Mbps data rate stated by the 802.11a standard is, only available at the physical layer (if the channel conditions allow). The actual *throughput* [i.e. average data transfer rate] that the user observes, say when one transmits a file over such a link is significantly lower than the nominal value. The main reason for this is the overheads necessary for the protocol operation in each of the stacked layers, including the access functions defined in the MAC layer. There is another metric of performance which we do not discuss here and it is: the ratio of the expected delivered payload to the expected transmission time is called *expected effective good-put*.

In the basic mode of transmission, the maximum throughput is given by [15]:

$$Throughput_{\max} = \frac{8L_{DATA}}{\frac{T_{slot}CW_{\min}}{2} + T_{DIFS} + T_{DATA} + T_{SIFS} + T_{ACK} + 2\tau}, \quad (2.1)$$

where the denominator is the total time taken for a packet of L_{DATA} bytes to be transmitted and acknowledged.; T_{DIFS} is the *DIFS* interval time, T_{SIFS} is the *SIFS* interval time, T_{DATA} is the actual time of the payload, T_{ACK} the acknowledge time and τ is the overall propagation delay. T_{slot} and $CW_{\min}$ are values of the time-slots and length of the minimum contention window, defined in the standard. This is a maximum, because back-to-back

transmission of the packets is assumed from the MAC layer viewpoint. A summary of these values quoted from the standard is given in the Appendix A.

2.8.2 Effect of repeaters on the link throughput

In a multihop wireless link (assumed using repeaters), the processing time at each node is significant. This is important since, the data packets at each repeater will be pushed up to the application layer and then down again to the physical layer. This added overhead may affect the overall performance of the link. In Figure 2.10, the wireless link consists of $(N-1)$ hops. At each node i , a certain processing delay τ_i , is introduced.

Figure 2.9 Multi-hop implementation of the wireless link

If the link uses N repeater with a processing delay τ' , the user sees $2N\tau'$ delay. This will affect the throughput of the link drastically. This effect may be illustrated by the following analysis.

In the wireless link [Figure 2.10] the total processing delay due to the repeaters (and end nodes) is:

$$\tau_{tot} = \sum_{i=1}^N \tau_i.$$

Taking this delay into account for the throughput equation, we will have:

$$Throughput(Mbps) = \frac{8.L}{\frac{T_{slot}CW_{min}}{2} + T_{DIFS} + T_{DATA} + 2\tau + T_{SIFS} + T_{ACK} + 2\sum_{i=1}^N \tau_i} \cdot (2.3)$$

The delay caused by a repeater which may process the data from the physical layer up to the application layer, is transparent to the link it is seen just as a net delay [the variables of equation (2.3) are the same as those for (2.1), except whenever they are different].

The sum of all delays along, including the protocol delays and the delay introduced by the repeaters must not exceed the time-out, so the link can be established; the protocol delay, depending on the implementation, includes the delays due to all the layers the data packet passes through. The worst case is when the packets are passed all the way through to the application layer (similar to our field tests in which a FTP software handled the packet transmission).

Using a set of parameter listed below, Figure 2.11 shows the effect of N repeaters each adding 50 μ Sec to the link. This test assumes no retransmission has happened which is not far from reality in a point-to-point scenario. Table 2.4 summarizes the OFDM frame parameters[next page].

Table 2.3 Parameters used for the throughput estimation

Type	Frame Length Parameters	Value
<i>MPDU Frame Fields</i>	FrameControl	2 byte
	Duration_ID	2 bytes
	Address	6 bytes
	SequenceControl	2 bytes
	Lpacket	1500 Byte
	FCS	4 bytes
<i>Interframe Spacing</i>	aSIFSTime	16 μ Sec
	aSlotTime	9 μ Sec
	SIFS = aSIFSTime	16 μ Sec
	DIFS=aSIFSTime+(2*aSlotTime)	34 μ Sec
<i>Data and Acknowledge Frame lengths</i>	Ldata=FrameControl +Duration_ID+4*Address+ SequenceControl+ Lpacket+FCS;	1534 Bytes
	Lack=FrameControl +Duration_ID+4*Address+ SequenceControl +FCS	14 Bytes

Figure 2.10 Throughput estimates over distance using different number of repeaters

Table 2.4 OFDM frame parameters

Type	Frame Length Parameters	Value
<i>OFDM Frame Timing</i>	Preamble Duration	16μSec
	PLCP Header Duration	4μSec
	aPreambleLength	20μSec
	aCW	<i>Minimum(2ⁿ - 1, 1023), n ≥ 5th Re transmission</i>
	Data Rate (Mbps)	6,9,12,18,24,36,48,54

2.9 Analysis of error probability

For an M-QAM signal, the symbol error rate is given by $P_M = 1 - (1 - P_{\sqrt{M}})^2$ [31],

where:

$$P_{\sqrt{M}} = 2 \left(1 - \frac{1}{\sqrt{M}} \right) Q \left(\sqrt{\frac{3E_{avg}}{(M-1)N_0}} \right)$$

This is the probability of a $\sqrt{M}$ -ary PAM modulation. If Gray code is used, the BER of an M-QAM signal can be approximated to:

$$P_b^{(M)} \approx \frac{1}{\log_2 M} \cdot P_M.$$

For $M = 4$, the M-QAM becomes QPSK. For BPSK, the BER and SER are the same:

$$P_b^{(2)} = P_2 = Q\left(\sqrt{\frac{2E_{avg}}{N_0}}\right).$$

where $Q(x)$ is defined as: $Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-y^2/2} dy$.

2.10 A look at OFDM and channel bit per symbol allocation [18-19]

An OFDM system is a multi-tone system as shown in Figure 2.12.

A multi-tone QAM system

for each block: P_i (power), f_i (carrier frequency),
 n_i (bits/symbol) and W_i (bandwidth)

Figure 2.11 Multi-tone implementation of an OFDM modulator

A bounded total power P is assumed for the system. The overall bit-rate of the channel may be written as: $R_b = \sum_{i=1}^N n_i W_i$, assuming relatively flat bandwidth W_i around the carrier frequency. The authors of [18] have analyzed the system for $N=2$. In this case: $R_b = (n_1 + n_2)W$, and the total power is $P = P_1 + P_2$. For even values of n , the probability of error is:

$$\Pr(\varepsilon) = 4 \left(1 - \frac{1}{\sqrt{2^{n_i}}} \right) Q \left[a \sqrt{\frac{2E_b}{N_0}} \right], \quad (2.4)$$

From the constellation for a M-QAM system in which $M = 2^{n_i}$, the average power of the symbol is acquired as: $a^2 = \frac{3E}{2(2^n - 1)}$. For odd values of n , the same relations provide only

approximations. From $\Pr(\varepsilon)$ we may find the number of bits per channels as follows:

$$n_1 = \log_2 \left[1 + \frac{3kP / N_0 W}{\left[Q^{-1} \left(\frac{\Pr(\varepsilon)}{K_{n_1}} \right) \right]^2} \right] \text{ and } n_2 = \log_2 \left[1 + \frac{3kP / N_0 W}{\left[Q^{-1} \left(\frac{\Pr(\varepsilon)}{K_{n_2}} \right) \right]^2} \right]. \quad (2.5)$$

in which K_{ni} 's are functions of n_i 's and $2 \leq K_i < 4$. Also k is the fraction of power in each channel. As it is seen from these equations, for a relatively flat channel W_i , the number of bits per symbol depends on the total power P in presence of noise power $N_0 W$. This relationship may be easily expanded to the general case.

2.11 Link Adaptation [20-21]

Wireless environment is very dynamic in nature. Various factors such as physical configuration of the surrounding objects, type of surrounding objects, and etc. affect the transmission media drastically. With the presence of noise, and path-loss increased over a distance, signal-to-noise ratio is reduced. As a result the BER (also PER) of the transmission channel, given a fixed transmitted power increases. Likewise, interference from other sources operating in the same frequency band or leakage from adjacent strong sources (adjacent channel transmitters) together with the non-linearity of the low-noise amplifiers have similar effects on the probability of error.

As seen from the above review, the capacity of the multi-tone OFDM channel heavily depends on the E_b/N_0 . In order to maximizing the channel capacity, it is important that the link transmission parameters be adopted to E_b/N_0 . The evaluation normally takes place in the training phase, however there different techniques which use more adaptive approaches. Normally the interactions established at the establishment of the link or transmission of packages such as RTC/CTS are carried out at the lowest mode (lowest capacity) to ensure the reception of data for the worst case. Then in later phases, the data transmission rate is increased to maximize the capacity. The process of adapting the data rate to the volatile wireless channel condition is link adaptation. More specifically, the procedure to adapt the wireless link speed and coding rate to the channel conditions is called *link adaptation*.

In 802.11a (and also 802.11g due to identical structures), several modes of operations have been defined which adapts the transmission speed and coding rate based on the available

SNR [Re. Appendix A]. The standard, however, does not specify any algorithms for the link adaptation. The procedure to adapt the wireless link speed and coding rate to the channel conditions is called link adaptation in the literature and different manufacturers have their own proprietary schemes for implementation. In this chapter a review of link adaptation is presented to better understand the operation of the link, however a detailed work on this issue is not the subject of our study.

2.11.1 A review of Link Adaptation strategies

There are several link adaptation schemes available in the market such as *Auto-Rate Fallback (ARF)* protocol which is introduced by Lucent Technologies. Another method is *Receiver-Based Auto-Rate (RBAR)* which is based on RTS/CTS mechanism (this mechanism is discussed in detail in the MAC layer). There are a few of other methods found in the literature. We will discuss some of these protocols in the following sections.

2.11.2 Auto-Rate Fallback (ARF) protocol

In this protocol the WLAN card switches between two PHY of 1 and 2 Mbps modes. The card checks if the Acknowledge (ACK) of the next data packet, it changes to the lower rate and re-transmits the packets under this lower mode (only two modes are available in the original 802.11 standard. Accuracy of this algorithm may not satisfactory for higher modes/rates and the speed of reaction to the changes in the channel may pose problems. This method which may be intuitively suggested is used in industry [Re. 22].

2.11.3 Receiver-Based Auto-Rate (RBAR) protocol

This protocol uses a by-product of RTS/CTS mechanism. This protocol follows this algorithm: (1) the receiver estimates the channel quality by sampling the end of the RTS signal, and selects an appropriate transmission rate and sends it back to the transmitter using the CTS and (2) The transmitter sends back the data packet using the estimated rate. The problem with this proposed protocol is the required changes in the RTS/CTS and Data frame formats which makes it less attractive to the industry.

2.11.4 Other Link-adaptation views and schemes

It seems that using only SNR for the optimizing the bit-rate and coding rate may not be an optimal way of optimizing the link. All of the proposed algorithms start from an estimate/measured of the SNR as the single input to the adaptation procedure, and may imply, to some extent, knowledge about the environment, or some by-products of the existing system (such as, the built-in viterbi decoder operation). The major deficiency of all of the proposed algorithms is that they may not respond most efficiently, to the dynamic nature of the channel. In [21] two more functions are defined for the link adaptation: power control, dynamic fragmentation which existing standard does not include them. However, the authors shed some new light on the link adaptation:

Authors of [21] classify the link adaptation algorithms for the systems using Automatic Repeat Request (ARQ), in two major categories: (1) Delay-oriented and (2) Throughput-oriented.

Delay-oriented algorithms try to minimize the retransmissions of the packets, while the *Throughput-oriented* algorithms try to maximize the “good-put” (actually delivered correctly throughput), irrespective of retransmissions.

Schemes such MPDU-based Link Adaptation scheme have been also introduced. The higher the mode of PHY, the faster the link, and the shorter the required transmission time. On the other hand, within (the now-reduced) frame retry limit (allowed number of retrials before giving up), it will be less likely that the delivery will be successful. Therefore, an optimal set of PHY modes may be found for which the data frame delivery is best optimized for a certain channel condition; the so-called “the best transmission strategy”[21].

2.11.5 Industrial method example [22]

The WLAN card uses a table such as this to associate the received power with the appropriate data rates for 802.11a mode:

Table 2.5 Example from D-Link DWL-520AB WLAN card

Appropriate Data Rate (Mbps)	54	48	36	24	18	12	9	6
Min. Power required (dBm)	-73	-76	-82	-85	-88	-89	-90	-91

In our discussions, Table 2.5 has been used as a practical reference [22].

2.12 Conclusion

In this chapter, a general description of the 802.11 MAC and the physical layer of the 802.11a, details of the DCF operation and different factors of media access have been presented. This included the basic access method and the RTS/CTS methods, fragmentation

and link adaptation. Also, the multi-hop implementation of the link and its effect on the average throughput were studied.

Chapter 3

Study of Radio-wave propagation effects

3.1 Introduction

The most important factor that affects the operation of the 802.11a point-to-point link is the physical environment. In this chapter, we introduce a detailed “two-ray” model, one that accounts for the effects of the terrain roughness, wind vibration, different types of antenna misalignments, the parabolic antennas and application of classical theory of diffraction on the attenuation rate of the signal with distance. Also, we develop an equation describing the ACI based on the spectral power masks of the 802.11a.

3.2 General

As it was observed in the previous chapter, the capacity of the wireless link is heavily dependant on the available signal power in presence of noise, or in other word the SNR. For a fixed SNR, in order to meet a specific error rate target, certain energy-per-bit of the received symbols must be maintained. If S is the signal power and R is the transmission bit rate, the energy-per-bit is calculated as follows.

$$E_b/N_0 = \frac{E_{symbol}}{k} \cdot \frac{1}{N_0} = \frac{S \cdot kT_b}{kN_0} = \frac{S}{R \cdot N_0}. \quad (3.1)$$

If this equation is expressed in dB:

$$E_b/N_0 \text{ (dB)} = 10 \log \left(\frac{S}{R \cdot N_0} \right) = 10 \log(S/N_0) - 10 \log R. \quad (3.2)$$

It is seen that for a fixed SNR, the higher the bit rate, the lower the energy-per-bit.

In order to compensate the attenuation of the signal, it may be transmitted at a higher power, or using an antenna with higher gain. This sort of analysis is known as link budget, in which the relationship between, the transmitted power, the received power and the pathloss is studied. The basic elements of link budget are presented schematically in Figure 3.1.

Figure 3.1 Elements of the Link budget

$$P_R \text{ (dB)} = P_T \text{ (dB)} + G_T + G_R - L_p \text{ (dB)}, \quad (3.3)$$

where P_T , P_R , G_T , G_R and L_p are the transmitted power, the received power, the transmitter antenna gain, the receiver antenna gain and the pathloss respectively. In addition to these, factors such as cable attenuation and different losses may also be included in general. If for a

certain data rate, a minimum received power P_{Rmin} is required, then the condition of successful transmission becomes:

$$P_R(dB) = P_T(dB) + G_T + G_R - L_p(dB) \geq P_{Rmin}. \quad (3.4)$$

It should be noted that, this is only a minimum threshold to establish the link at the target data rate.

In a link budget analysis, normally the significant part which is forced by physical environment, is the attenuation of the signal over a distance i.e. pathloss. In literature, there is a large diversity of pathloss models, which differ in terms of variables such as frequency band of validity, accuracy, and also if they are experimental or theoretical. The following section briefly discusses a few popular models.

3.3 Examples of path loss models [23-24]

1) The free-space model

This model assumes an isotropic antenna radiating in free space. Due to the unrealistic assumptions in this model, large errors are expected in predicted values of the path loss. However, this model may provide limited accuracy on high elevations and over a long distance:

$$L_{p-freespace} = (4\pi d / \lambda)^2. \quad (3.5)$$

2) The simplified two-ray model

This model is an improved version of free space model in a sense that it considers a direct radio ray from the transmitter to the receiver in addition to a second one reflected/refracted

ray, and that it considers the effects of the terrain to some extent (the relative heights from a reference level). Although the accuracy of this model is improved compared to the free-space model, but due to still unrealistic assumptions, it also, suffers from inaccuracies:

$$L_{p\text{-}two\text{-}ray} = \frac{h_1^2 h_2^2}{d^4}. \quad (3.6)$$

A major problem with this model is that it does not depend on frequency.

3) Egli's Model

$$L_{p\text{-}Egli} = \left(\frac{f(\text{MHz})}{40} \right) \frac{h_1^2 h_2^2}{d^4}. \quad (3.7)$$

This model is an obvious attempt to include frequency dependency into the model. It resembles in behavior the simplified two-ray model for a fixed frequency; only a shift-down will be observed if plotted in dB.

4) Hata's model for open environment:

$$PL_{Hata} = A + B \log_{10}(d) - C, \text{ in which: } A = 69.55 + 26.16 \log_{10}(f) - 13.82 h_t - \alpha h_r,$$

$$B = 44.9 - 6.55 h_t, C = 40.94 + 4.78 \log_{10}(f)]^2 - 18.33 \log_{10}(f) \text{ and}$$

$$\alpha = [1.1 \log_{10}(f) - 0.7] h_r - [1.56 \log_{10}(f) - 0.8], [f = \text{frequency in MHz, } d = \text{distance in km, } h_t = \text{transmitter height, } h_r = \text{receiver height}]. \quad (3.8)$$

5) Cost231 Model:

$$PL_{Cost} = A + B \log_{10}(d) + C, \quad (3.9)$$

$$A = 46.3 + 33.9 \log_{10}(f) - 13.28h_t - \alpha h_r,$$

$$B = 44.9 - 6.55h_t,$$

$C = 0$ for medium-sized urban or suburban areas.

6) Model proposed in [23]:

$$L_p = 40 \log_{10} d + 20 \log_{10} f - 20 \log_{10}(h_t h_r). \quad (3.10)$$

This model has the advantage that it takes into consideration the effects of the Fresnel zones, but it shows inaccuracies in practical cases. Fresnel zones will be discussed in the diffraction analysis section. Despite the significant inclusion of frequency, based on our experimental data, the predictions of the model do not prove to be accurate enough. Therefore, due to inaccuracies and/or frequency restrictions of the above models, we resort to ray-tracing technique in order to estimate the path-loss appropriately.

If a wireless link uses highly directional antennas, it is safe to assume that analysis using two rays of radio is sufficient to estimate the pathloss since the intersection of the spatial volume defined by the antenna's half-power beamwidth (HPBW) angle at the actual location of the receiver antenna is very small area. HPBW is defined as the angle between two lines on the major cross-section of the pattern of the antenna which pass through the points where the (received/transmitted) power is half of the maximum (received/transmitted). Therefore using the classical two-ray model may provide enough accuracy. However, in the classical model, factors such as reflection from ground,...etc have been approximated. To obtain a more accurate model, the following sections investigate different propagation effects in some details in order to study the extent of their effects. Also factors such as roughness of the terrain and diffraction have been discussed.

3.4 Effects of Reflection [25]

3.4.1 Reflection of Electromagnetic waves

Consider two different cases of an incident wave from an environment with electrical parameters: (ϵ_1 , μ_1 and σ_1) and (ϵ_2 , μ_2 and σ_2). In one case the incident wave is vertically polarized and in the other one, it is polarized horizontally (Figure 3.2). Due to boundary conditions at the interface of two environments with different parameters, always the normal components of electric fields and parallel components of the magnetic fields across the interface are equal and the corresponding parallel magnetic and vertical electric components, respectively, add up to zero.

Figure 3.2 Planar electromagnetic wave boundary conditions at the interface of two different media

Therefore after equating/vanishing the appropriate components, the reflection coefficient of the electromagnetic wave at the interface of the two environments is:

$$\alpha_h = \frac{E_r}{E_i} = \frac{\eta_2 \sin \theta_t - \eta_1 \sin \theta_i}{\eta_2 \sin \theta_t + \eta_1 \sin \theta_i} \quad (3.11)$$

for a horizontally polarized wave, and for a vertically polarized wave:

$$\alpha_v = \frac{E_r}{E_i} = \frac{\eta_2 \sin \theta_t - \eta_1 \sin \theta_i}{\eta_2 \sin \theta_t + \eta_1 \sin \theta_i}. \quad (3.12)$$

Note that the incident and reflected angles are “usually” taken as similar to those used in optics, but as long as the derivations are consistent, this is not of major importance.

If the appropriate boundary conditions and the comprising components of the wave impedances (η) are inserted, the reflection coefficients for the vertical and horizontal polarizations are obtained as:

$$\alpha_h = \frac{-\varepsilon_r \sin \theta_i + \sqrt{\varepsilon_r - \cos^2 \theta_i}}{\varepsilon_r \sin \theta_i + \sqrt{\varepsilon_r - \cos^2 \theta_i}} \text{ and } \alpha_v = \frac{\sin \theta_i - \sqrt{\varepsilon_r - \cos^2 \theta_i}}{\sin \theta_i + \sqrt{\varepsilon_r - \cos^2 \theta_i}}. \quad (3.13)$$

In all of the analysis above, it was assumed that ε_r is constant. However, it is more accurate to assume a lossy environment. In this case we may always take the permittivity as:

$$\varepsilon = \varepsilon_0 \varepsilon_r - j\varepsilon' \text{ where: } \varepsilon' = \frac{\sigma}{2\pi f}. \text{ However in the 5 GHz frequency band operation, } \varepsilon' \text{ is}$$

negligible and the above formula is approximated as: $\varepsilon = \varepsilon_0 \varepsilon_r$.

Table 3.1 provides some materials electric parameters [next page].

Table 3.1 Typical material electric parameters

Material	Relative Permittivity ϵ_r	Conductivity σ (s/m)	Frequency (MHz)
Poor Ground	4	0.001	100
Typical Ground	15	0.005	100
Good Ground	25	0.02	100
Fresh Water	81	0.001	100
Brick	4.44	0.001	4000
Limestone	7.51	0.028	4000

Figure 3.3 change of reflection index over distance at different frequencies and vertical/horizontal polarizations[the arrows within the curves starts from 1GHz and ends at 10GHz].

For values of $\varepsilon_0 = 8.8419 \times 10^{-12}$, $\mu_0 = 4\pi \times 10^{-7}$ and in the frequency range of 100MHz-10 GHz, plots of Reflection index are shown in Figure 3.3.

It can be observed from Figure 3.3 that approximation of $|\alpha| = 1$, for grazing angle (i.e. angle of incidence approaches zero) is safe; however this approximation may not be valid when the incident angle at the reflecting point is significantly different from zero. As a rough measure angles less than 5 degrees may be considered as a good approximation for grazing angle. This may happen on a sloped terrain or at short distances (when the height of the antennas is not negligible). In all of these cases, use of a directional antenna, reduces the effect of reflection because of the sharp attenuation of the rays received in an angle different from the axis of the antenna.

3.5 Effect of roughness of the ground

The ground in between the two antennas is most of the times very rough. The roughness of a reflecting surface is measured by comparison to a reference critical height h_c which is

defined by [25]:

$$h_c = \frac{\lambda}{8 \sin \theta_i}. \quad (3.14)$$

Figure 3.4 Critical height over distance at different frequencies

If the difference between the highest point of a rough surface and its lowest point is less than h_c , the surface is considered as smooth. This is called the “Rayleigh’s Criterion” for surface roughness [25]. Figure 3.4 illustrates the change of h_c over a distance of 1000 meter [both antenna heights are assumed less than 2m]. It is clearly seen that for higher frequencies, the critical height becomes smaller. Therefore, the effect of roughness becomes more significant. For a fixed wave length, if the distance between two directional antennas increases, the incident angle decrease to grazing angle and h_c is increased. In other words the rough features become less “visible”.

3.6 A Quantitative measure for roughness

In [25] the reflection coefficient is modified by a roughness factor defined as:

$$\rho_s = \exp \left[-8 \left(\frac{\pi \sigma_h \sin \theta_i}{\lambda} \right)^2 \right]. \quad (3.15)$$

This factor has been further modified by Landron et al [25] as:

$$\rho_s = \exp \left[-8 \left(\frac{\pi \sigma_h \sin \theta_i}{\lambda} \right)^2 \right] J_0 \left[8 \left(\frac{\pi \sigma_h \sin \theta_i}{\lambda} \right)^2 \right], \quad (3.16)$$

in which in which J_0 is the Bessel function of the first kind and zero order. Therefore, the reflection factor is modified by the roughness factor as: $\alpha_{rough} = \rho_s \alpha$. (3.17)

Figures 3.5 through 3.10 show the effect of roughness for different values of standard deviations of the surface. As it is seen the clear distinction of the reflection factor becomes ambiguous and diminishes in value for angles above 25 degrees. This may be contributed by presence of more scattering as roughness increases. However for angles less then 10 degrees, the effect is negligible. In analyzing the roughness for a surface, one has to note for a link on a sloped surface incident angle increases and this is no longer negligible.

Figure 3.5 Index of reflection for sigma height =5cm; Horizontal polarization

Figure 3.6 Index of reflection for sigma height = 10cm ; Horizontal polarization

Figure 3.7 Index of reflection for sigma height = 30cm ; Horizontal polarization

Figure 3.8 Index of reflection for sigma height = 5cm ; Vertical polarization

Figure 3.9 Index of reflection for sigma height = 10cm ; Vertical polarization

Figure 3.10 Index of reflection for sigma height = 30cm ; Vertical polarization

3.7 A simple model for parabolic antennas

Maximum gain of an antenna in general may be approximated by [26]:

$$G_{\max} = \frac{4\pi}{\theta_{HPBW} \phi_{HPBW}}, \quad (3.18)$$

in which θ_{HPBW} and ϕ_{HPBW} are half-power beam-widths (in radians) in vertical and horizontal planes, respectively. In this model, it is assumed the E-plane and H-plane patterns are identical (such as in a symmetric parabolic reflector antennas). On the other hand, pattern shape of a parabolic antenna may be modeled as:

$$G(\theta) = G_{\max} \cos^N(\theta), \quad (3.19)$$

in which θ is the azimuth angle and N is a factor which determines the sharpness of the pattern of the antenna. The above relationships may be useful in finding an optimal sharpness of an antenna pattern, which directly translates into the expenses for a radio link.

The high gain antenna reduces the probability of interference and also reduces the power of the ground reflected ray in the classical two-ray model. However, when the antenna beam gets sharper, the link gain becomes sensitive to misalignment errors and vibration due to wind or other possible factors. Figure 3.11 shows an example of antenna radiation pattern simulated by computer.

Figure 3.11 Pattern of antenna with HPBW = 6 degrees

Therefore from the maximum gain formula, for an antenna with E-plane and H-plane pattern (such as a parabolic dish reflector), HPBW can be calculated as:

$$\theta_{HPBW} = \sqrt{\frac{4\pi}{G_{\max}}}. \quad (3.20)$$

For the simulation purposes, a value of N which corresponds to this angle may be found from the equations (3.19) and (3.20):

$$N = \left\lceil \frac{\ln\left(\frac{1}{2}\right)}{2 \ln(\cos(\theta_{HPBW}))} \right\rceil. \quad (3.21)$$

The upper limit means to the nearest rounded-up integer. Such calculations are illustrated in Figure 3.12.

Figure 3.12 N-parameter for the antenna shown in Figure 3.11

3.8 Review of the classical two-ray model [25]

If a transmitter is at point A and another receiver at point B (at distance d from the transmitter), the configuration will look like Figure 3.13:

Figure 3.13 Two-ray model scenario

The received power at point B is the sum of the direct ray and the reflected ray which has a phase delay $\Delta\phi$. Therefore,

$$P_r = P_0 \left(\frac{1}{4\pi\lambda d} \right)^2 \left| G_{\max 1} G_{\max 2} + \alpha G_1(\theta_1) G_2(\theta_2) e^{j\Delta\phi} \right|^2 \quad (3.22)$$

in which $G(\theta)$ is the antenna gain as defined in the previous section. Also, α is the reflection coefficient. This value is of significant importance depending on what type of polarization is used (section 3.4).

If the $G_{\max}(\theta)$ values are factored out, we will have:

$$P_r = P_0 \left(\frac{1}{4\pi\lambda d} \right)^2 G_{\max 1} G_{\max 2} \left| 1 + \alpha G_{L1}(\theta_1) G_{L2}(\theta_2) e^{j\Delta\phi} \right|^2 \quad (3.23)$$

From the geometry of Figure 3.13, the direct path length is: $d'' = \sqrt{d^2 + (h_1 - h_2)^2}$, and the path length for the reflected wave is: $d' = \sqrt{d^2 + (h_1 + h_2)^2}$. Therefore the path length difference between the rays is: $\Delta d = \sqrt{d^2 + (h_1 + h_2)^2} - \sqrt{d^2 + (h_1 - h_2)^2}$. (3.24)

The terms containing square roots may be approximated using Taylor series and the equation can be simplified as follows:

$$\Delta d \approx d \left[1 + \frac{1}{2} \left(\frac{h_1 + h_2}{d} \right)^2 - 1 - \left(\frac{h_1 - h_2}{d} \right)^2 \right] = \frac{d}{2} \left[\frac{4h_1 h_2}{d^2} \right] = \frac{2h_1 h_2}{d}. \quad (3.25)$$

Figure 3.14 (on the next page) shows the absolute error due to this approximation. It is observed that the error is less than a wavelength (6 cm at 5GHz) for distances above 387 meters, when the heights of the transmitter and the receiver are both 15 meters.

Figure 3.14 Estimate error on the distance difference

Therefore the phase difference becomes: $\Delta\varphi = \frac{2\pi h_1 h_2}{\lambda d}$.

At grazing angles, i.e. for long distance, the reflection coefficient becomes: $\alpha_v \approx 1$, for vertical polarization and $\alpha_h \approx -1$, for horizontal polarization at grazing angles. As mentioned before, at sloped ground, these approximations may sometimes be inaccurate. Therefore, in our calculations, we consider the reflection index which is a function of distance and applies to both types of polarization. Whenever necessary, it is shown by subscripts v or h for clarity.

Therefore we may conclude that path-loss equation becomes:

$$\frac{P_r}{P_0} = \left(\frac{G_{\max 1} G_{\max 2}}{4\pi\lambda d} \right)^2 \left| 1 + \alpha G_{L1}(\theta_1) G_{L2}(\theta_2) e^{j\Delta\varphi} \right|^2. \quad (3.27)$$

This is the general equation that may be applied to different antennas with different gain equations.

Figure 3.15 shows the received power using different antenna gains:

Figure 3.15 Received power versus distance for different antenna gains

Compare this to the free-space model using GR and GT of both 30 dBi and the same 13dBm transmitted power:

$$\frac{P_r}{P_0} = \left(\frac{1}{4\pi\lambda d}\right)^2 \cdot G_T G_R \quad (3.28)$$

As seen from Figure 3.16 (next page), this model produces very unrealistic results.

Figure 3.16 Received power predicted by the Free-space model

Assuming vertical polarization ($\alpha = 1$), Figures 3.17-3.20 show the received power for different antenna gains of 10,20,30 and 50 dBi parabolic antennas using a 13 dBm transmitted power.

Figure 3.17 Received power versus distance for a 10 dBi antenna

Figure 3.18 Received power versus distance for a 20 dBi antenna

Figure 3.19 Received power versus distance for a 30 dBi antenna

Figure 3.20 Received power versus distance for a 50 dBi antenna

It is observed from Figures 3.17-3.20 that if the antenna beam becomes sharper, the reflected ray (as compared to the direct ray) becomes less effective. This property may also be used to reduce the effect of interference.

3.8.1 Further approximations

If in the pathloss equation, the exponential term is expanded by the Euler's rule, we will

have:
$$\frac{P_r}{P_0} = K_f^2 \left| 1 + \alpha G_{L1}(\theta_1) G_{L2}(\theta_2) (\cos \Delta \varphi + j \sin \Delta \varphi) \right|^2. \quad (3.29)$$

In which: $K_f = \left(\frac{G_{\max 1} G_{\max 2}}{4\pi \lambda d} \right)$

In case of parabolic antenna with the proposed model, the pathloss equation becomes as follows:

$$\frac{P_r}{P_0} = K_f^2 \left| 1 + \alpha \cos^N(\theta_1) \cos^N(\theta_2) (\cos \Delta \varphi + j \sin \Delta \varphi) \right|^2. \quad (3.30)$$

As distance d grows far greater than the antenna heights, the phase difference between the direct ray and the reflected ray becomes very small and a further approximation may be done as follows:

$$\frac{P_r}{P_0} = K_f^2 \left| 1 + \alpha \cos^N(\theta_1) \cos^N(\theta_2) \left(1 + j \frac{4\pi h_1 h_2}{\lambda d} \right) \right|^2. \quad (3.31)$$

The effect of this approximation is shown in Figures 3.21-3.22:

Figure 3.21 Received power estimated after the approximation (3.31) for a 20 dBi antenna

Figure 3.22 Received power estimated after the approximation (3.31) for a 30 dBi antenna

It is observed that this approximation, is valid only on the large-scale, however, it eliminates the short-range effects; compare Figures 3.21 and 3.22 with Figures 3.18 and 3.19 (which do not include such approximations).

3.9 Effects of misalignment of the antennas

1) Misalignment in vertical plane perpendicular to the line of propagation (Type I)

Definition: The Two antennas are aligned with an error angle δ in the vertical plane of the antenna.

Figure 3.23 Effect of misalignment (Type I) on polarization

In case of misalignment of Type I, the total waves received will have two differently polarized components; A horizontally-polarized and vertically-polarized component. If A is the amplitude of the signal: $\vec{A} = \vec{A}_h + \vec{A}_v$, where $|\vec{A}_v| = |\vec{A}| \cos \delta$, and $|\vec{A}_h| = |\vec{A}| \sin \delta$.

In this case, the pathloss equation will be modified as:

$$\frac{P_r}{P_0} = K_f^2 \left[\sin \delta \left(1 + \alpha_h \cos^N(\theta_1) \cos^N(\theta_2) (\cos \Delta \varphi + j \sin \Delta \varphi) \right) \hat{i} + \dots \right. \\ \left. \dots + \cos \delta \left(1 + \alpha_v \cos^N(\theta_1) \cos^N(\theta_2) (\cos \Delta \varphi + j \sin \Delta \varphi) \right) \hat{j} \right]^2 . \quad (3.32)$$

However, only the vertical component of the wave can be receive (assuming a very small cross polarization of the receiving antenna). Therefore, the actual pathloss equation becomes:

$$\frac{P_r}{P_0} = K_f^2 \left| \cos \delta \left(1 + \alpha_h \cos^N(\theta_1) \cos^N(\theta_2) (\cos \Delta \varphi + j \sin \Delta \varphi) \right) \right|^2. \quad (3.33)$$

Figure 3.24 shows the effect of this type of misalignment, at a distance d and different gains.

Figure 3.24 Effect of misalignment (type I) on (pathloss)

2) Misalignment in vertical plane including the line of propagation (Type II)

Definition: The misalignment is in the horizontal plane which contains the line of propagation. In this case the form of the two-ray model equation will remain the same, but the $\delta = 0$ and the θ 's will be replaced by $\theta + \epsilon$. Therefore:

$$\frac{P_r}{P_0} = K_f^2 |1 + \alpha \cos^N(\theta_1 + \varepsilon) \cos^N(\theta_2 + \varepsilon) (\cos \Delta \varphi + j \sin \Delta \varphi)|^2, \quad (3.34)$$

in which ε is the total alignment error on the same plane. This can be generalized to any plane that includes the line of propagation. The effect of this type of misalignment may be seen in Figures 3.25-27.

Figure 3.25 Effect of misalignment(Type II) on pathloss [Gain = 20 dBi]

Figure 3.26 Effect of misalignment (Type II) on pathloss [Gain = 30 dBi]

Figure 3.27 Effect of misalignment (Type II) on pathloss [Gain = 40 dBi]

It is observed that the misalignment type II is significantly more effective than misalignment of type I for antennas with gains of 30 dBi or more. The second type of misalignment, due to its sensitive nature, must be avoided with careful alignment of the antennas.

3.9.1 Effects of vibration of the antenna due to wind

For simplicity, the vibration of the antenna (possibly by factors such as winds) is assumed only in the horizontal plane. If the standard deviation (rms value) of the vibrations is assumed to be θ_ε (as considered from the LOS) then the path-loss equation may be modified in a similar way to the misalignment Type II, since the vibration is assumed to be in the horizontal plane. For simplicity and since it is the relative angle between the two antennas that is important, the error angle may be added only to one of the gain factor angles [equivalently: one side may be windy]. Therefore the equation for the path-loss is modified as:

$$\frac{P_r}{P_0} = K_f^2 \left| 1 + \alpha \cos^N(\theta_1 + \theta_\varepsilon) \cos^N(\theta_2) (\cos \Delta\varphi + j \sin \Delta\varphi) \right|^2. \quad (3.35)$$

The cosine function may be expanded as follows:

$$f(\theta + \theta_\varepsilon) = \cos(\theta + \theta_\varepsilon) = \cos \theta \cos \theta_\varepsilon - \sin \theta \sin \theta_\varepsilon \approx \cos \theta \left[1 + \left(\frac{\theta_\varepsilon}{2!} \right)^2 \right] - \sin \theta \sin \theta_\varepsilon, \quad (3.36)$$

If θ_ε is small, this can be further approximated by: $f(\theta + \theta_\varepsilon) \approx \cos \theta - \sin \theta \sin \theta_\varepsilon$.

Therefore,

$$f^N(\theta + \theta_\varepsilon) = (\cos \theta - \sin \theta \sin \theta_\varepsilon)^N = \cos^N \theta (1 - \theta_\varepsilon \tan \theta)^N \approx \cos^N \theta (1 - N \theta_\varepsilon \tan \theta). \quad (3.37)$$

Using this approximation, the pathloss equation becomes:

$$\frac{P_r}{P_0} = K_f^2 \left| 1 + \alpha \cos^N \theta (1 - N \theta_\epsilon \cdot \text{tg} \theta) \text{Cos}^N(\theta_2) (\text{Cos} \Delta \varphi + j \text{Sin} \Delta \varphi) \right|^2. \quad (3.38)$$

Effect of a vibration due to wind may be seen in Figure 3.28 for a 40 dBi antenna.

Figure 3.28 Effect of vibration due to wind

It is again observed that similar to misalignment Type II, the received is very sensitive to wind vibration, if the antenna gain is 30dBi or more. Also it is seen that the effects of misalignment (Type II) and wind-generated vibration deteriorates with distance.

3.9.2 Effect of terrain roughness on pathloss [25]

As discussed in the analysis of reflection, the roughness of the ground affects the reflection index by multiplying a factor ρ_s [25]. In the pathloss equation, we consider the reflection index as a function of roughness [according the previous discussion on roughness]:

$$\frac{P_r}{P_0} = \left(\frac{G_{\max 1} G_{\max 2}}{4\pi\lambda d} \right)^2 \left| 1 + \alpha(\rho_s) G_{L1}(\theta_1) G_{L2}(\theta_2) e^{j\Delta\varphi} \right|^2. \quad (3.39)$$

Figures 3.27 and 3.28 show how ρ_s changes over a distance using equations 3.15 and 3.16, using 1m antennas.

Figure 3.29 Reflection coefficient factor using (3.15)

[The inner arrow shows the change from 10cm to 1m.]

Figure 3.30 Reflection coefficient factor using (3.16)

It is observed that the roughness factor is small for distances under 600m. It means that the second ray (the reflected ray in the two-ray model) is more attenuated for shorter distances. Also the factor approaches 1 for longer distances.

Figures (3.29) and (3.30) show the received power at fixed distances for different degrees of the terrain roughness (in terms of the standard deviation of the heights from a flat surface).

Figure 3.31 Received power versus the terrain roughness (Gain = 20 dBi)

Figure 3.32 Received power versus the terrain roughness (Gain = 30 dBi)

It is evident that for shorter distances, the effect of the roughness of the terrain is more significant, however, in general, with the increased roughness, the received power is decreased; compare Figure (3.33) with Figures (3.17-20).

Figure 3.33 Received power in case of a rough surface

It can be seen that for a rough ground the pathloss increases with the rate of 20dB/decade throughout the interval. This is mainly due to the decreased value of roughness factor for shorter distances (larger incident angles). This may also happen when the transmitter and the receiver are located on a sloped terrain. Therefore in any estimate of the pathloss, the

In Figure 2.24, the α is indicated at the

$$dE = \frac{kdu e^{-j\beta r}}{f(x)}, \quad (3.40)$$

in which k is a constant, β is the wave figure and $f(r)$ is a function of r such as a polynomial.

From the geometry we have:

$$(QR)^2 = r^2 = (d_1 + d_2)^2 + d_1^2 - 2d_1(d_1 + d_2)\cos\frac{u}{d_1} \approx (d_1 + d_2)^2 + d_1^2 - 2d_1(d_1 + d_2)\left(1 - \frac{u^2}{2d_1^2}\right),$$

and

$$r^2 = (d_2 + \delta)^2 \approx d_2^2 + u^2 \frac{d_1 + d_2}{d_1}. \quad (3.41)$$

And if δ^2 is negligible:

$$E = \frac{k}{f(r)} \int_{u_0}^{u_1} e^{-j\beta r} du = \frac{ke^{-j\beta d_2}}{f(r)} \int_{u_0}^{u_1} e^{-j\beta \delta} du = \frac{ke^{-j\beta d_2}}{f(r)} \left(\int_{u_0}^{u_1} \cos \beta \delta du - j \int_{u_0}^{u_1} \sin \beta \delta du \right), \quad (3.42)$$

in which: $\beta \delta = \frac{\pi}{\lambda} \left(\frac{d_1 + d_2}{d_1 d_2} \right) u^2.$

Therefore:

$$|E|^2 = \frac{k^2}{f^2(d_2)} \left[\left(\int_{u_0}^{u_1} \cos \beta \delta du \right)^2 + \left(\int_{u_0}^{u_1} \sin \beta \delta du \right)^2 \right]. \quad (3.43)$$

Now consider this equation using the functions $C(v)$ and $S(v)$:

$$C(v) - jS(v) = \int_0^v e^{-j(\pi/2)v^2} dv \quad (3.44)$$

The integral is in the form of a *Fresnel Integral*. Plotting these the integral on the $C(v)$ - $S(v)$ plane will produce *Cornu Spirals*[Figure 3.35].

Figure 3.35 Cornu Spirals

For our problem at hand:

$$C = \int \cos \frac{\pi v^2}{2} dv, \quad S = \int \sin \frac{\pi v^2}{2} dv, \quad (3.45)$$

$$\delta s = \sqrt{(\delta C)^2 + (\delta S)^2} = \delta v, \quad v = s, \quad \tan \phi = \frac{\delta S}{\delta C} = \tan \frac{\pi v^2}{2}, \quad \phi = \frac{\pi v^2}{2} = \frac{\pi s^2}{2}, \quad \frac{d\phi}{ds} = \pi s;$$

Radius of Curvature = $\frac{ds}{d\phi} = \frac{1}{\pi s}$, $C(\pm\infty) = \pm \frac{1}{2}$, $S(\pm\infty) = \pm \frac{1}{2}$. These curve has the following

properties:

- 1) A vector from the origin to any point on the C-S plane is equivalent to the integral above at that point.
- 2) As $|v|$ grows, the vector (the integral) will rotate around point $(1/2, 1/2)$ for positive values and $(-1/2, -1/2)$ for negative values of v .
- 3) Magnitude of the integral $\sqrt{C^2 + S^2}$ has a maximum at $\Phi = 3\pi/4$ (or $v = 1.225$).

The Magnitude of the integral is of major significance to us.

The Electric field may be written in form of the Fresnel integral form as follows:

$$E = K \left(\left[\frac{1}{2} - C(v_0) \right] - j \left[\frac{1}{2} - S(v_0) \right] \right), K = \frac{E_0}{2} (1 + j). \quad (3.46)$$

From this equation, the electric field is acquired as: $E = \frac{E_0}{2} (1 + j) \int_{v_0}^{\infty} e^{-j(\pi/2)v^2} dv$ where:

$$v_0 = u_0 \sqrt{\frac{2(d_1 + d_2)}{\lambda d_1 d_2}}. \text{ The value of } u_0 \text{ is chosen to be } h \text{ [Re. 25]. Plot of this integral is shown}$$

in Figure 3.36.

Figure 3.36 Diffraction gain versus v-parameter

E_0 is the free-space field strength. This approximation is valid when d_1 and $d_2 \gg u_0$ and $d_1 d_2 \gg \lambda$. Theoretically, it is seen that when the obstacle prevents the radiation right at middle of the “field-of-sight”, the wave is attenuated by 6 dB.

In addition to the above observation, one has to note that the attenuation of the radio waves follows a 40dB/decade after the Fresnel distance (instead of the free-space 20 dB/decade).

Therefore the concept of Fresnel zone clearance becomes significant.

3.7.2 Definition of Fresnel zones [25]

In Figure 3.37, from geometry, if $F1$ and $F2$ are at distance $2d$, the loci of the points that the sum of their distances from $F1$ and $F2$, that is d' , is greater than d , only by a fixed value, is an ellipse with parameter: d . If two antennas are located at the same height h , at a distance d_0 at which the ellipse is tangent to the ground and the parameter d is equal to $\lambda/2$, the ellipsoid which is formed by the rotation of the ellipse around the line connecting the two antennas is called the first Fresnel zone. In general, an ellipsoids associated with n -times $\lambda/2$ is called the n -th Fresnel zone.

Figure 3.37 Definition of Fresnel zones

3.10.2 Fresnel zone clearance [25]

This is particularly important when the line of sight is close to the ground. In fact, when the height of the line-of-sight gets less than 0.55-0.60 of the first Fresnel zone, path-loss will

start to increase drastically which conforms with the above model. The first Fresnel zone is

given by a circle of radius: $R_{First-Fresnel} = \frac{\sqrt{\lambda d}}{2}$

The distance for which this radius touches the ground is given by: $D_f = 4h_1h_2 / \lambda$,

in which h_1 and h_2 are the antenna heights and λ is the wave-length of the operation.

If $h = \sqrt{h_1h_2}$ is the geometric mean of the two antenna heights, the Fresnel zone distance $D_f = 4h^2 / \lambda$, and it can be presented in Figure 3.38 .

Figure 3.38 Fresnel zone distance versus geometric mean of antenna heights

As it is seen from Figure 3.37, the Fresnel distance at the 5 GHz wavelength and antenna height of 15m (a typical country house) provides Fresnel zone clearance close to 15 km.

3.11 Application of Diffraction theory to a sloped terrain

Classic theory of diffraction using Fresnel zones in most of the literature is restricted to flat-earth scenario. Using the flat-earth for non-flat situation may not provide enough accuracy in many situations, especially if the angles θ and β (Figure 3.38) are significant. The following section includes an analysis of diffraction on a sloped terrain.

In Figure 3.39, H_t and H_r are the antenna heights from the respective bases on the slope hill. The hill has a slope of β and the angle of the hill to the flat ground is θ . In order to simplify the ellipse equation, the line connecting the top of the antennas has been taken as x-axis. The ellipse touches the hill at point $T(x_0, y_0)$.

Since the line describing the sloped hill passes through the antennas bases, the equation of the tangent line becomes:

Figure 3.39 Application of the classic theory of diffraction to a sloped terrain scenario

$y_t = \frac{x+c}{y+h_t} = \frac{c+c}{-h_r+h_t} = \frac{2c}{\Delta h}$, in which $\Delta h = h_t - h_r$. Therefore:

$y = \frac{\Delta h}{2c}x + \left(\frac{\Delta h}{2} - h_t\right) = \frac{\Delta h}{2c}x + \left(\frac{\Delta h}{2} - h_t\right) = \frac{\Delta h}{2c}x - \frac{h_t + h_r}{2} = Ax + B$, in which:

$$A = \frac{\Delta h}{2c}, B = -\frac{h_t + h_r}{2}, c = \frac{d}{2}, \quad (3.48)$$

and d is the distance between the transmitter and the receiver antenna top points.

By intersecting the tangent line equation and the ellipse equation, $\frac{x^2}{a^2} + \frac{y^2}{b^2} = 1$ and the tangent line this quadratic equation is acquired:

$$x^2(b^2 + a^2 A^2) + 2a^2 ABx + a^2(B^2 - b^2) = 0. \quad (3.49)$$

In order that this equation will only have one (double) solution, the discriminant must be equal to zero. Therefore, considering the ellipse parametric equation: $c^2 = a^2 - b^2$, we may

solve the discriminant equation to obtain: $a^2 = \frac{B^2 + c^2}{A^2 + 1}$, and $b^2 = \frac{B^2 - c^2 A^2}{A^2 + 1}$. (3.50)

In order to have a solution for the quadratic equation (equivalently, not having Fresnel clearance), the discriminant should be greater or equal to zero. From this condition, the restriction is acquired for clearance: $B^2 \geq c^2 A^2$.

Coordinates of the point at which the tangent line “touches” the ellipse, is therefore acquired as:

$$x_0 = -\frac{ABa^2}{b^2 + A^2 a^2}, y_0 = \frac{Bb^2}{b^2 + A^2 a^2}. \quad (3.51)$$

The relationship of the actual antenna heights H_t and H_r and the respective projections h_t and h_r are acquired from the geometry of Figure 3.35 as:

$$\frac{H_t}{h_t} = \frac{H_r}{h_r} = \frac{\cot \beta}{\tan \gamma - \cot \beta} \text{ in which: } \gamma = \theta - \beta. \quad (3.52)$$

So far the ellipse with foci at the transmitter and receiver antennas and touching a sloped hill has been defined. If the total distance of the ray diffracted at the point T, is a multiple of $\lambda/2$. The first of such ellipse (and in fact ellipsoid, since the actual setting is 3-D) comprises the first Fresnel Zone which is the most important for classical diffraction theory. The value of y_0 is the “negative h ” used in the calculation of the diffraction loss.

The path difference of the two rays (direct and diffracted) is calculated as:

$$\Delta d = \sqrt{(x_0 + c)^2 + y_0^2} + \sqrt{(x_0 - c)^2 + y_0^2} - d, \text{ in which } c = d/2. \quad (3.53)$$

The values for y_0 and Δd should be used with the classical diffraction formula. The total attenuation of the radio wave must include all of those factors. In the following section, three different approaches are discussed.

3.12 Combinations of different losses[28]

In an ideal picture, the path-loss between the transmitter and the receiver over a distance may be seen as a loss over a “horizontal” axis while the diffraction loss due to lack of Fresnel clearance zone may be seen as a loss over a “vertical” axis. There are three ways of combining these losses(in dB) proposed in the literature:

1) Bullington:

$$A_{tot} = A_{freespace} + A_{Reflection} + A_{Diffraction}, \quad (3.54)$$

which is readily suggested in ideal conditions.

2) Blomquist:

$$(3.55)$$

$$A_{tot} = A_{freespace} + \max(A_{Reflection} + A_{Diffraction})$$

and:

3) Blomquist empirical formula:

$$A_{tot} = A_{freespace} + \sqrt{A_{Reflection}^2 + A_{Diffraction}^2} \quad (3.53)$$

Note that in formula (1), $A_{freespace} + A_{Reflection}$ is the same as two-ray model. Therefore it is appropriate to simply add the attenuation due to diffraction to the pathloss. It has to be mentioned for the third formula, no mathematical proof has been offered by Blomquist, although it has been claimed to provide more accurate results. Therefore use of the first formula is justified. If the diffractive conditions allow such, one may also use the second combination method as a safety measure.

3.13 Sources of interference

There are two types of interference in the wireless link: 1) In-band (co-channel) interference, 2) Adjacent channel interference. The source of in-band interference may be a another network which is working on the same channel, or from devices such as cordless phones which recently emerged in the market. This issue is analyzed in more details later.

3.13.1 Maximum Adjacent Channel Interference [29]

The 802.11 channel is 20 MHz wide. In the 802.11a version, Orthogonal Frequency Division Multiplexing (OFDM), where the information is loaded on tones called sub-carriers. The space between sub-carriers is 312.5 KHz. Within 20 MHz, there is enough space for 64 sub-carriers, but in order to reduce the adjacent channel interference, only 52 sub-carriers are used, of which 48 sub-carriers are for data and 4 are for pilot signals.

In [1] the authors have studied the effect of the in-band and ACI. This study concluded that, linear amplifiers would significantly reduce the ACI but it won't help reduce the in-band interference.

Figure 3.40 Adjacent channels in a typical PHY

In order to model the interference, we consider the power spectrum mask as defined by the standard. Using a typical power spectrum mask, the following analysis may be performed: If γ (dBm) is the power level at each point and f is the frequency (in MHz), for the desired channel, we can express the mask in terms of the following for intervals: $fc-34 < f < fc-11$, $fc-11 < f < fc-9$, $fc-9 < f < fc+9$, $fc+9 < f < fc+11$ and $fc < f < fc+34$, respectively:

$$\begin{aligned}
 \gamma &= f - 16, \\
 \gamma &= 13.5(f + 9), \\
 \gamma &= 0, \\
 \gamma &= -13.5(f - 9), \\
 \gamma &= -f - 16.
 \end{aligned} \tag{3.57}$$

Figure 3.41 Maximum interference estimate with spectral power masks

The interference may be simulated by designing a filter with the above characteristic which filters a source of noise (for simplicity we consider an AWGN source). In order to find the ACI characteristics for the upper and lower channel ACI, the same equations are shifted up

(and down): *Up – Shifted*

$$\begin{aligned}
 \gamma &= f - 36, \\
 \gamma &= 13.5(f - 11), \\
 \gamma &= 0, \\
 \gamma &= -13.5(f - 29), \\
 \gamma &= -f + 4.
 \end{aligned}$$

Down – Shifted

$$\begin{aligned}
 \gamma &= f + 4, \\
 \gamma &= 13.5(f + 29), \\
 \gamma &= 0, \\
 \gamma &= -13.5(f + 11), \\
 \gamma &= -f - 36.
 \end{aligned} \tag{3.58}$$

The overlap areas for both cases are as follows:

<i>Up – Shifted</i>	<i>Range</i>	<i>Down – Shifted</i>	<i>Range</i>
$\gamma = 2f - 52,$	$-14 \leq f < -11,$	$\gamma = 2f - 12,$	$-34 \leq f < -30,$
$\gamma = f - 36,$	$-11 \leq f < 9,$	$\gamma = f - 16,$	$-30 \leq f < -11,$
$\gamma = -27,$	$9 \leq f < 11,$	$\gamma = -27,$	$-11 \leq f < -9,$
$\gamma = -f - 16,$	$11 \leq f < 30,$	$\gamma = -f - 36,$	$-9 \leq f < 11,$
$\gamma = -2f - 12.$	$30 \leq f \leq 34.$	$\gamma = -2f - 52.$	$11 \leq f \leq 14.$

These equations may prove useful in estimation of maximum ACI for parallel operation of the channel. In this case:

$$ACI(dBm) = P_{Adjacent_Channel}(dBm) - L_{combiner}(dB) - \alpha_{ACI} \quad (3.60)$$

In this equation, the first term is the operating transmitted power of the adjacent channel transmitter, $L_{combiner}$ is the isolation of the ports of a combiner/splitter and α is the value calculated from the above equations.

3.13.2 A practical scenario in parallel operation of 802.11a links [29]

Figure 3.42 shows a scenario in which RF outputs of two WLAN cards were connected to a single antenna through a splitter with maximum isolation of 20 dB. Assuming a transmitted power of 13 dBm, and a minimum sensitivity of -73 dBm for a maximum rate of 54 Mbps, it can be clearly seen that when two WLAN cards work together in an uncorrelated way using such an splitter/combiner, the ACI can easily shut down the receiving link. The phenomenon has been confirmed in our experiments.

Figure 3.42 ACI due to use of combiner/splitter

3.14 Conclusion

In this chapter, different factors which affect the propagation of radio waves in the frequency of interest (5 GHz) have been studied. These factors included reflection, diffraction, roughness, different types of antenna misalignments and vibration due to wind. Each factor has a certain degree of importance which has been covered with graphic presentation. Finally, the formulation for ACI and the method of combination of different losses have been discussed. Our models for a parabolic antenna, its maximum gain and effects of vibrations due to wind, different types of misalignments, application of the classic theory of diffraction to a sloped terrain and effect of terrain roughness have been modeled and their effects have been studied using simulations. The presented models were then combined to provide the aggregate effect on the operation of the proposed point-to-point link.

Chapter 4

Link model

4.1 Introduction

In this chapter, we use the models discussed and use them to define a point-to-point link model to estimate the received power (or equivalently) the pathloss, given the antenna elevations, and slope of the terrain. We also discuss the effects of pathloss, as well as interference and introduce a procedure for implementing the point-to-point link.

4.2 Link implementation

Figure 4.1 shows a single-hop model for the point-to-point link consisting of a terminal at point X and another one at point Y to form a wireless link at distance d .

Figure 4.1 A single-hop link

An alternative way to realize the link, is to use a multi-hop implementation as shown in Figure 4.2 [for clarity named as A and D with midpoints B and C]. The direct distance between the two end-points is the same (Figure 4.1). However, in practice for the multi-hop implementation (Figure 4.2), it is more reasonable to assume that hops are not aligned with the direct line XY in the single-hop scenario.

Figure 4.2 A multi-hop link

For simplicity, it is assumed that all of the stations use the same type of antennas and terminals B and C operate as repeaters. Also it is assumed that the front-to-back ratio of the antennas used in the link is finite. Obviously, this makes it hard or even impossible for the segments of link to work at the same frequency, since even for a very high front-to-back ratio (e.g. 60 dB), the direct reception from A to C by or B to D could be quite significant.

4.3 Estimate of losses for a single hop

In order to estimate the losses (pathloss and diffraction loss) we use the equations developed in chapter 3. For pathloss we use the complete version of the two-ray model:

$$\frac{P_r}{P_0} = K_f^2 \left| 1 + \alpha \cos^N(\theta_1) \cos^N(\theta_2) (\cos \Delta \varphi + j \sin \Delta \varphi) \right|^2. \quad (4.1)$$

in which $\Delta\varphi = \frac{2\pi h_1 h_2}{\lambda d}$. In this equation, the index of reflection uses one of equations (3.13)

depending on the type of polarization used:

$$\alpha_h = \frac{-\varepsilon_r \sin \theta_i + \sqrt{\varepsilon_r - \cos^2 \theta_i}}{\varepsilon_r \sin \theta_i + \sqrt{\varepsilon_r - \cos^2 \theta_i}} \text{ and } \alpha_v = \frac{\sin \theta_i - \sqrt{\varepsilon_r - \cos^2 \theta_i}}{\sin \theta_i + \sqrt{\varepsilon_r - \cos^2 \theta_i}}. \quad (4.2)$$

It is assumed that ε_r is constant.

In equation (4.1) the model used for the antenna is described by the pattern:

$$G(\theta) = G_{\max} \cos^N(\theta). \quad (4.3)$$

Given a specific maximum gain, the HPBW of the antenna is calculated as:

$$\theta_{HPBW} = \sqrt{\frac{4\pi}{G_{\max}}}, \quad (4.4)$$

The antenna parameter N is calculated as:

$$N = \left\lceil \frac{\ln\left(\frac{1}{2}\right)}{2 \ln(\cos(\theta_{HPBW}))} \right\rceil, \quad (4.5)$$

which defines the antenna model. Note that N is an integer (next greater integer).

4.4 Estimate of the diffraction loss

Using the model developed in chapter 3, the diffraction loss is estimated using the Fresnel's diffraction loss integral:

$$L_{Diffraction} = \frac{E}{E_0} = \frac{(1+j)}{2} \int_v^\infty e^{-j(\pi/2)u^2} du, \quad (4.6)$$

in which[25]:

$$v = u_0 \sqrt{\frac{2(d_1 + d_2)}{\lambda d_1 d_2}} = h \sqrt{\frac{2c}{\lambda(c^2 - x_0^2)}}. \quad (4.7)$$

For calculation of the parameter v , referring to Figure 3.35 as follows:

$$A = \frac{\Delta h}{2c}, B = -\frac{h_t + h_r}{2}, c = \frac{d}{2}, \Delta h = h_t - h_r, x_0 = -\frac{ABa^2}{b^2 + A^2a^2}, h = y_0 = \frac{Bb^2}{b^2 + A^2a^2}.$$

Also:

$$a^2 = \frac{B^2 + c^2}{A^2 + 1}, \text{ and } b^2 = \frac{B^2 - c^2 A^2}{A^2 + 1}. \quad (4.8)$$

Therefore the overall loss is estimated as follows:

$$L_{total} = L_p (dB) + L_{Diffraction} (dB). \quad (4.9)$$

4.5 Implementation of the link

Using this total loss, the Friis link budget equation is formed as follows. Assuming the minimum received power P_{rmin} for a specific data rate is known, this is the condition of link establishment at the specified data rate: $P_r (dB) = P_T (dB) - L_{total} (dB) \geq P_{rmin}$. (4.10)

Since the commercial cards have fixed transmitted power, the maximum gain of the antennas (included in the pathloss equation) must be chosen so that the above condition is met.

4.6 Estimate of the required SNR for a wireless link using OFDM

If the P_{rmin} of the receiver is not given, it can be estimated using the relation between SNR and E_b/N_0 which was presented at chapter 3:

$$E_b/N_0 (dB) = 10 \log \left(\frac{S}{R \cdot N_0} \right) = 10 \log(S/N_0) - 10 \log R. \quad (4.11)$$

This relationship may be modified for OFDM signal as follows:

$$SNR(dB) = \frac{E_b}{N_0} (dB) + 10 \log_{10} \left(R \cdot N_{BPSC} \cdot \frac{N_{ST}}{N_{FFT}} \cdot \frac{T_{FFT}}{T_{SYMBOL}} \right), \quad (4.12)$$

where R is the coding rate, N_{BPSC} is the coded bits per symbol, N_{ST} is the number of the subcarriers, N_{FFT} is the number of FFT points, T_{FFT} is the FFT period and T_{SYMBOL} is the symbol interval. Possible values of these parameters have been summarized in Table 4.1.

Table 4.1 OFDM symbol paramaters

Paremeter	Possible Values
Coding Rate R	1/2, for 6, ,12 , 24; 2/3, for 48, and 3/4 for 9.18,36 and 54 Mbps
N_{BPSC}	1 for BPSK,2 for QPSK,4 for 16QAM and 6 for 64QAM
N_{ST}	48 data + 4 pilots
N_{FFT}	64
T_{FFT}	3.2μSec
T_{SYMBOL}	4μSec

Using a desirable, BER an estimate of E_b/N_0 may be obtained for an M-QAM signal [31]:

$$P_{\sqrt{M}} = 2 \cdot \left(1 - \frac{1}{\sqrt{M}} \right) \cdot Q \left(\sqrt{\frac{3E_{avg}}{(M-1) \cdot N_0}} \right). \quad (4.13)$$

4.7 Study of interference within the field-of-sight of the wireless link

Consider Figure 4.3 in which there is an interferer which is within the field of vision of the transmitter A (i.e. within the HPBW spatial. There is another transmitter (interferer) at point I

which works on the same frequency and is located not too far away from the *HPBW* angle of the Rx antenna at a distance d_i .

For simplicity, we assume that the transmitter, the receiver and interferer, and also the main cross sections of the antenna patterns are all on the same plane. In reality, this is only a special case and the plane containing the terminals and the main cross sections might not be the same, but since this is an extreme case, it helps in outlining the criteria for the interference planning.

Figure 4.3 Interferer at a close angle to the LOS

Assume the pattern of the antenna at terminal A is defined by the equation (4.3) and the received EIRP of the interference is P_{int} (as if received by an isotropic antenna) the h_t , h_r , are transmitter and receiver antennas heights, and d is the LOS distance between the transmitter and the receiver and the receiver. Also, assume that the LOS and the line connecting the receiver to the interferer are at an angle ϕ (Figure 4.3). Then the received power from the transmitter, at the receiver is in this case:

$$P_R = P_0 G_{T \max}^2 G_{R \max}^2 |1 + \alpha \cos^{N_1}(\theta_1) \cos^{N_2}(\theta_2) e^{j\Delta\varphi}|^2. \quad (4.20)$$

in which:

$$\Delta\varphi = \frac{2\pi h_t h_r}{\lambda d}.$$

However, the interference power is:

$$P_{\text{int}} = P_{\text{int}} G_{R \max}^2 \cos^{2N_1}(\theta_2 + \varphi). \quad (4.21)$$

Since the line of interference is not at the direction of the maximum gain of the antenna, it would be attenuated by the pattern of the receiving antenna, since it is reasonable to assume that the two antennas are not aligned for the interference reception.

4.8 Interference suppression

Unlike cellular telephony in which, the goal of the base station is to cover the cell area which is modeled as a circle or a hexagon, in a point-to-point the propagation is supposed to be restricted to the LOS, however, the practical directional antennas are not perfectly directional (zero HPBW). Therefore, if there is an interferer (a transmitter working on the same frequency) within the “field of vision” of the antenna, the received signal will be corrupted. In this situation, it is desired to find out the power of interference and its angle of arrival. We may assume that the interferer is Gaussian random signal and we may characterize it by the angle of direction and its EIRP. If interference avoidance is not possible (e.g. another channel is not available) the sum of the transmitter power and the antenna gain must be adjusted for a guaranteed SNIR at the receiver. Based on this, the criteria for frequency selection and reuse on a specific topology are proposed.

4.10 Interference avoidance using higher antenna gains

As discussed in chapter 3, the antenna gain (which is closely related to its *HPBW*) may be adjusted to a minimum so that the γ required for the target data transfer rate is available. However, if the pattern of the antenna becomes too wide and there is an interferer within the field of vision of the antenna, the received interferer becomes larger. Therefore in selection of an antenna, a trade-off must be made to minimize the interference, while keeping the SNIR above the minimum level to sustain the data rate. This process is graphically illustrated in Figure 4.4. As it is seen, an interfering source is avoided or minimized by using an antenna with a higher gain. [Note the scale of distance and a fixed transmitted power.]

Figure 4.4 Graphic presentation of interference suppression (20 dBi)

Figure 4.5 Graphic presentation of interference suppression (25 dBi)

Figure 4.6 Graphic presentation of interference suppression (30 dBi)

If possible, in order to minimize the implementation costs (of antennas with higher gains) in using this method, the added gain on one antenna may be deducted from the antenna at the other side of the hop, unless its field-of-sight becomes too broad to include a potential

interferer. However, in order to keep the cost minimal, the antenna with the gain along the multi-hop wireless link should be chosen the same for all of the stations.

4.9 The link establishment

In 802.11a, the transmission of handshakes or broadcast signals, such as the beacon, and RTS/CTS are all done in the lowest data rate mode.

The link establishment has a few steps as briefly described in chapter 2, before the link is ready for the actual data transmission. However, after this phase, the link stabilizes and transmission becomes smooth [See Appendix A for some field test results]. In other words, availability of SNR (at least at low levels) may not guarantee a fast link establishment according to our practical field tests. However, if the SNR assumes more significant values, then the establishment phase becomes almost transparent. One way to avoid this situation is to allow more SNR at the receiver, during the link budget analysis, by allowing higher antenna gains, more accurate alignments and (if there is a choice) addition of transmitted power.

4.10 A procedure for implementation of the wireless link

In order to implement the wireless link the following approach may be taken.

4.10.1 Frequency selection

The following guidelines are necessary for selection of the frequency of operation of the hops:

- Channels with lower frequencies are preferred since they exhibit less path-loss.
- Previous information/field measurement data to identify potential sources of interference in the planned frequency band.

- If there is an interferer that cannot be avoided, then a trade-off must be made as described in beginning of the section 4.10, to safely maintain the data transfer rate.

4.10.2 Procedure for link implementation

In order to implement the wireless link, the following procedure is proposed:

- 1) Providing an on-site measurement/survey to identify the interferers in the service area.
- 2) Start with the lowest available channel.
- 3) Measure the interference level (if any).
- 4) Estimate the total loss between the transmitter and the receiver antennas using the equation (4.1).
- 5) For a fixed transmitted power, find the lowest antenna gain (choice of parameter N using the antenna model provided in chapter 3), that provides the minimum received power for the maximum data rate. Verify if this antenna is suitable for all of the hops. If so, this will be the antenna selection for every hop of the link. Otherwise, select the next higher-gain antenna and verify again.
- 6) Calculate the HPBW of the selected antenna (using equation 4.3-4.5 for the antennas).
- 7) For the selected antenna, find the front-to-back ratio.
- 8) To avoid adjacent channel interference, skip one channel and pick a higher channel for the connecting hops. Measure the interference of each hop of interest with the neighboring hop. If the operation of a hop hampers the connected hop, a) the channels should be chosen with higher frequency separation or, b) an antenna with a higher front-to-back ratio must be chosen.

The procedure repeats from the step 2 on for the rest of the hops.

This procedure has been summarized in Figure 4.5.

Figure 4.7 Link implementation procedure

4.11 Conclusion

In this chapter, based on the detail analyses of the physical environment provided in chapter 3, a model for the wireless has been provided. The model includes the equations for the pathloss, diffraction loss, description of possible interference and the procedure to avoid interference, a procedure for implementation of the wireless link. The next chapter provides the simulation results of the model and discussed issues.

Chapter 5

Simulation results and discussions

5.1 Introduction

In this chapter, using the models in chapter 4, simulation results are presented to verify the purpose of these models. Also, to demonstrate the feasibility of the proposed wireless link, part of our field test results are presented in the form of average throughputs. Details of the test results are found in appendix A. To make a quick reference for the reader, the equations used in the simulations are quoted below:

For pathloss we use the complete version of the two-ray model:

$$\frac{P_r}{P_0} = K_f^2 |1 + \alpha \cos^N(\theta_1) \cos^N(\theta_2) (\cos \Delta\varphi + j \sin \Delta\varphi)|^2. \quad (4.1)$$

in which $\Delta\varphi = \frac{2\pi h_1 h_2}{\lambda d}$. In this equation, the index of reflection uses one of equations

depending on the type of polarization used:

$$\alpha_h = \frac{-\varepsilon_r \sin \theta_i + \sqrt{\varepsilon_r - \cos^2 \theta_i}}{\varepsilon_r \sin \theta_i + \sqrt{\varepsilon_r - \cos^2 \theta_i}} \text{ and } \alpha_v = \frac{\sin \theta_i - \sqrt{\varepsilon_r - \cos^2 \theta_i}}{\sin \theta_i + \sqrt{\varepsilon_r - \cos^2 \theta_i}}. \quad (4.2)$$

Parameter ϵ_r is the relative electric permittivity and θ_i is the incident angle.

In the equation (4.1) the model used for the antenna is described by the pattern:

$$G(\theta) = G_{\max} \cos^N(\theta). \quad (4.3)$$

Given a specific maximum gain, the HPBW of the antenna is calculated as:

$$\theta_{HPBW} = \sqrt{\frac{4\pi}{G_{\max}}}, \quad (4.4)$$

the antenna parameter N is calculated as:

$$N = \left\lceil \frac{\ln\left(\frac{1}{2}\right)}{2 \ln(\cos(\theta_{HPBW}))} \right\rceil, \quad (4.5)$$

which defines the antenna model.

5.2 Received power versus height simulations

In a setup using a transmitter/receiver pair and parabolic reflector antennas, the following simulation illustrates the received power as a function of height. This simulation has been repeated for different antenna gains for comparison. Also, to clarify the results, in each set of test, distance has been used as parameter. In all of the simulation, we have assumed a 13dBm transmitted which is a typical value and it is the same for our WLAN card [22]. Table 5.1 shows the list of default parameters for all of the simulation. Wherever necessary, default values have been replaced by appropriate values and quoted where the results have been presented. These values have been chosen as close as possible to our field test setup [shown in Appendix A]; for our simulations we have used MATLAB R14.

Table 5.1 Simulation default parameters table

Frequency (GHz)	5.25	Standard deviation of the terrain roughness (m)	0.1
TXr Antenna Height (m)	5	Electric relative permittivity (ϵ_r)	4
RX Antenna Height(m)	5	Max. Antenna gain	30 dBi
Transmitted power (dBm)	13	Default polarization	Vertical
Conductivity of ground σ	0.005	Min Rx power for 54 Mbps Data rate	-73 dBm

In Figure 5.1, using a 30 dBi antenna, the curves of the received power vs. height for different distances are illustrated. In order to discuss the significance of height, Figure 5.1 shows the change of received power versus height with distance used as a parameter, at 5.25 GHz, and 13dBm transmitted power. Equation (4.1) and its dependencies are used as model.

Figure 5.1 Received power versus height at 5.25 GHz

It is observed that at any specific distance, beyond a certain height, the received power will remain constant. To clarify how this effect changes with the antenna gain, we repeat the above test with antenna gains of 20 dBi and 40 dBi (which correspond to practical extreme values). Figure 5.2 shows the same effect for a 20 dBi antenna.

Figure 5.2 Received power versus height at 5.25 GHz (20 dBi antenna gain)

Figure 5.3 also shows the same effect for a 40 dBi antenna.

Figure 5.3 Received power versus height at 5.25 GHz (40 dBi antenna gain)

As it is seen in Figures 5.1-5.3, as the antenna gain is increased the antenna heights beyond which, for a fixed distance, the received power remains constant is reduced. The inverse of this conclusion is also correct when the antenna gain is decreased. This effect is caused by the fact that, for an antenna of a higher gain, at shorter heights, the second reflected ray in the two-ray model becomes negligible. Therefore, the conclusion from the simplified two-ray model that increasing the received power with antenna height is limited up to a certain height. In fact this effect shows that beyond a certain height for a specific antenna gain and distance, the model is reduced to the free-space model. This conclusion has been shown in Figures 5.4 and 5.5.

Figure 5.4 Received power predicted by Free Space model for different antenna gains

Figure 5.5 The received power prediction by the detailed two-ray model beyond certain heights are the same as Free Space model

As is seen from the above figures, the values predicted by the Free-space model and those of high-gain parabolic antennas are exactly the same beyond a certain height.

If the height values beyond which the received power remains constant are connected by line segments as in Figure 5.6, a decaying behavior is displayed.

Figure 5.6 Behavior of constant-power heights with distance

5.3 Effect of frequency on the received power

Since the 802.11a PHY operates over a wide range of frequencies (5.15-5.8 GHz), a similar test is repeated to observe the effect of the carrier frequency on the received power. This is shown in Figures 5.7 and 5.8. It is observed that over the 802.11a frequency band, effect of frequency is negligible.

Figure 5.7 Effect of channel frequency on received power (40 dBi antennas)

Figure 5.8 Effect of channel frequency on received power (20 dBi antennas)

5.4 Small scale effects

If the same scenario as in sections 5.2 and 5.3 is repeated over shorter range of heights to focus on the effects on a smaller scale, it may be observed that [Figure 5.9], due to severe fluctuation of the received power, due care must be taken especially when lower gain antennas are utilized gain. The horizontal dashed lines are received power thresholds for different PHY data rates.

Figure 5.9 Small-scale effects on received power

It is observed that at low heights, a variation (e.g. by 2 meters when $h=15\text{m}$ and distance is 10km) in heights may result in loss of communication, as the received power curves may cross threshold lines.

5.5 Diffraction v -parameter and Diffraction Loss

In this section we examine the diffraction model introduced in chapter 4. The model uses the equations quoted below [all referred to Figure 3.35]. The diffraction loss integral:

$$L_{Diff} = \frac{E}{E_0} = \frac{(1+j)}{2} \int_{v_0}^{\infty} e^{-j(\pi/2)v^2} dv, \quad (4.6)$$

in which: $v = h \sqrt{\frac{2c}{\lambda(c^2 - x_0^2)}}$. (4.7)

Parameter v , referring to Figure 3.35 is calculated using the following parameters:

$$A = \frac{\Delta h}{2c}, B = -\frac{h_t + h_r}{2}, c = \frac{d}{2}, \Delta h = h_t - h_r, x_0 = -\frac{ABa^2}{b^2 + A^2a^2}, h = y_0 = \frac{Bb^2}{b^2 + A^2a^2}.$$

Also:
$$a^2 = \frac{B^2 + c^2}{A^2 + 1}, \text{ and } b^2 = \frac{B^2 - c^2 A^2}{A^2 + 1}. \quad (4.8)$$

Therefore the overall loss is estimated as follows:

$$L_{total} = P.L.(dB) + L_{Diff}(dB). \quad (4.9)$$

Figure 5.10 shows change of v-parameter with height and its effect on the pathloss.

Figure 5.10 V-parameter versus height

Figure 5.11 Diffraction loss versus height

The above attenuation has been integrated in the following simulations in the following sections.

5.6 Available PHY Layer data rate over a distance

As described in the model, adding several attenuation effects [pathloss, reflection index and diffraction loss] together, the available data rates at the PHY are depicted in Figures 5.11-5.14. In this set of simulations, the available PHY data rates are illustrated using a typical minimum received power thresholds. In these scenarios, the antenna gains of 10 dBi and 20dBi at heights of 10m and 20m have been assumed, respectively [4 cases].

Figure 5.12 Received power and available PHY data transfer rate (the upper curve)

Figure 5.13 Received power and available PHY data transfer rate

Figure 5.14 Received power and available PHY data transfer rate

Figure 5.15 Received power and available PHY data transfer rate

5.7 Practical test results from field tests

Figure 5.15 shows the averages of different test results from various field test using antennas of 1.8m height [the heights included in the figure includes the site elevation] and 28dBi gain and D-Link 802.11a WLAN cards. Samples of the test results have been included in Appendix A. The points inside the ellipse refer to the test in which two wireless links were combined using a two input splitter/combiners. As it is seen, the average of the measured throughput values, has reduced significantly in parallel operation. This problem is due to the poor isolation (20dB) of the splitter/combiner and leakage of the links over each other when one of the links was in transmission mode and the other one in the reception. The upper values (connected by dotted line) are the average of the results obtained by single links.

Figure 5.16 Averages of Field Test results [The heights shown include the site elevations.]

The values shown are averages of the throughput test results. The above test results show the feasibility of connecting two nodes in a point-to-point connection despite the inconveniences associated with the existing technology for the proposed point-to-point connection.

5.8 Effects of wind vibration

Assuming a random vibration at the horizontal plane with a zero-mean normal distribution of with standard deviation of 3 degree[half of a 30 dBi HPBW as a significant value], at heights of 2 m, the receive power is shown in Figure 5.16.

Figure 5.17 Received power vs. distance in a windy environment

It is observed that when antenna gain increases, it becomes more sensitive to the vibrations. This effect can become very significant on a long haul link. However, in our proposed link model, the length of the link (or hops) may be a few hundred meters. Therefore this effect is not significant in our application in which the hop lengths are expected to be in the range of

1km or less. As it is seen, in case of the 40 dBi antenna, at distances above 10km, the fluctuation may go over 30 dB.

5.9 Interference suppression

As it was discussed in previous chapters, the effects of interferer may be minimized using antennas of higher gain. As an illustration, since the pattern of radiation of the interferer may not be known, it is safer to assume an interferer of 1dBm EIRP [a random voltage received with zero mean Gaussian distribution with 1 dBm power] at 45 degree angle to the LOS to show its effect of the SNIR pattern. If the receiver antenna is rotated from -90 degrees to 90 degree with respect to the LOS, the SNIR at the receiver is acquired according to Figure 5.17. The transmitted power is 13 dBm at antenna heights of 2m.

Figure 5.18 Variation of SNIR by rotating the receiver antenna

As it is readily expected, the SNIR is maximum, when the two antennas are aligned along the LOS. However, if an interferer is at a very close angle to the LOS, in case of 10 dBi and 20dBi antennas, the SNIR may drop significantly and the link may fail. However, the gain of the receiver antenna may be chosen to a higher value to suppress the interferer. This is illustrated in Figure 5.18. It should be noted that, this method should be used only if interferer may not be avoided e.g. by selecting another carrier frequency.

Figure 5.19 Suppression of interferer of 1 dBm EIRP at 5 degrees to LOS

As it is seen in Figure 5.18, the interference at a 5° angle is totally suppressed by using the 30 dBi antenna which demonstrates the effectiveness of the method. However, high cost of higher-gain antenna must be considered in this approach.

5.10 Possible implementations for wireless connections of sparsely-populated areas

As it was discussed in the previous chapters, different solutions are possible for providing high-speed wireless access to the areas which are sparsely scattered around the cities.

- 1) One solution is using cellular structure and connecting the cells either by wire-line or wireless to the central station in the city. In this scheme each cell will cover certain areas.

The problem with this method is that only a small percent of the radio power is received by the user communities, unless expensive adaptive antenna systems are used. In case of using omni-directional antennas in most of the areas the power is wasted and the range is reduced. Also substantial amount of investment is required, if the cells are connected by wireline.

- 2) The second solution is using a mesh network by connecting the nodes via wireless links over relatively long distances.

In the second method, the nodes of the mesh may be connected by high towers to cover the loss due to long distance pathloss. In this case, use of high towers requires expensive equipment, tower and appropriate real-estate. However, as mentioned earlier, it is possible to use multi-hop links to connect any two nodes with low-antenna multi-hop links which may be installed on roof-tops or naturally available high points. In this scenario, the links are implemented using the 802.11a WLAN technology (which operates at 5 GHz) and directional antennas. Afterwards, for the distribution system use of 802.11g access points may be made (which operates at 2.4GHz with the same PHY data rate as 802.11a and it is commercially widely available-Figure 5.19 on the next page). Since the hops are connected

to the points (such as houses) at which a population may exist, as a by-product, it is possible to provide the service to this type of users and those in their vicinity; this is not possible in the single-hop link.

Figure 5.20 A Multi-hop 802.11a link with 802.11g distribution (dotted circles)

In the following section, a comparison between the high-tower solution and the low-antenna multi-hop solutions is made and advantages and drawbacks are pointed out. Since the distribution system is not the subject of this work, we only assume its existence and analyze only the wireless hops. In all the following estimation, the electric relative permittivity has been assume to a typical value of 4.

In the following sections, the throughput calculation, these equations is used [quoted from chapter 2]:

$$Throughput_{\max} = \frac{8L_{DATA}}{\frac{T_{slot}CW_{\min}}{2} + T_{DIFS} + T_{DATA} + 2\tau + T_{SIFS} + T_{ACK}}, \quad (2.1)$$

where the L_{DATA} is the length of the actual data portion in the frame in Bytes, T_{DIFS} is the $DIFS$ interval time, T_{SIFS} is the $SIFS$ interval time, T_{DATA} is the actual time of the payload,

T_{ACK} the acknowledge time and τ is the overall propagation delay. T_{slot} and CW_{min} are values of the time-slots and length of the minimum contention window, defined in the standard.

5.11 Comparison between the high-tower solution and low-antenna multi-hop link

For the comparison of the proposed multi-hop solution, the following scenario is considered.

It is desired to connect the point X to the point Y.

Figure 5.21 A single wireless link connecting X and Y.

5.11.1 Single-hop implementation of the link

Table 5.2 shows the specifications of the wireless link to be established between points X and Y.

Table 5.2 Link parameters for the single-hop link

Distance (m)	Transmitted Power (dBm)	Height (X) (m)	Height (Y) (m)	Ground Roughness (m)	Frequency (GHz) /Channel Number	Total Received power (dBm)	Antenna gain (dBi)	Maximum sensitivity to wind vibration
8500	13	30	30	0.1	5.25	-31.2	20	RE. Figures 5.23-5.24

Figure 5.22 Received power estimate for the single hop link XY

Table 5.3 Total average delay and throughput for the single-hop link XY

Single-hop link XY	Total Delay	28.33 μ Sec
	Average Throughput	23.2 Mbps*

* From Figure 5.25.

It has to be noted that a change of distance by 500m further or roughly 700m closer to the transmitter, will attenuate the power drastically, unless a higher antenna gain is used.

5.11.2 Multi-hop implementation of the link

The same link is implemented to connect two points (here called A and D to avoid confusion) using three links AB, BC and CD in Figure 5.22, and the following table of parameters.

Figure 5.23 A multi-hop wireless link connecting A and D.

The received power estimate together with the link parameters have been shown in Table 5.4. Some of the parameters have been extracted from Figures 5.23 and 5.26. As it is observed, the with a 3-section multi-hop link, a throughput performance close to the single-ho link may be acquired. However, due to the nodes delays, the throughput may suffer to some extent which must be minimized by using faster drop-insert schemes/hardware. In our field tests due to the restriction of the resources, our use of software has been restricted to application layer which is very inefficient; the drop-insert operation must be done as close as possible to the PHY. This delay has been considered in Table 5.5. In this table, τ is the delay due to use of repeaters. [In the simulation results (Figure 5.26) effect of this delay has been shown].

Figure 5.23 shows the received power using the parameters detailed in Table 5.4; using the same curve the received power for different link segments has been shown.

Figure 5.24 Received power estimate for the multi-hop link AD

Table 5.4 Link parameters for the multi-hop link AD

The hop section	Distance (m)	TX Power (dBm)	TX Height (m)	RX Height (m)	Ground Roughness (m)	Frequency (GHz) /Channel Number	Total Received Power (dBm)	Antenna gain (dBi)	Maximum sensitivity to wind vibration
AB	2500	13	2	2	0.1	5.125	-16	20	RE. Figures 5.23-5.24
BC	3000	13	2	2	0.1	5.165	-17.8	20	
CD	4000	13	2	2	0.1	5.205	-20.2	20	

Table 5.5 Total average delay and throughput for the multi-hop link AD

Multi-hop link AD	Total Delay (μ Sec)	8.33 μ Sec + 10 μ Sec + 13.33 μ Sec + τ
	Average Throughput (Mbps)	RE. Figure 5.25

5.11.3 Sensitivity of the link to the wind vibration

In this section, using our proposed link model pathloss [equation(4.1)], a zero-mean Gaussian random angle vibration with 3° standard deviation[half of our practical antennas HPBW] is added to the transmitter antenna only (for clarity the receiver is considered fixed). Using the a 20 dBi antenna, the received power is shown in Figure 5.24.

Figure 5.25 Received power for a wind vibration SD of 3 degree

As an extreme case, the same test is repeated with a 20 dBi antenna and 10° (Figure 5.25-next page).

Figure 5.26 Received power for a wind vibration SD of 10 degree

It is observed from Figures 5.24 and 5.25 that the single-hop implementation of the link is more sensitive to the wind vibration; although for a 3 degree standard deviation, in both cases, the received power fluctuation is negligible, for higher deviation such as 10 degree, it can be come quite significant, especially if the antenna gain is higher.

5.11.4 Throughput comparison of the multi-hop and single-hop

In Figure 5.26, using the model introduced in Chapter 2, throughput is shown versus the processing time of each node.

Figure 5.27 Throughput comparison

It can be observed that for the multi-hop implementation of the wireless link, in order that the average throughput of the link does not reduce below 90% of the single-hop link, it is required that the processing delay of each node, must not exceed 12.7 μ Sec. Since the Single-hop link has no repeater, in the view of the MAC layer, the processing time has no effect on the average throughput.

5.12 Conclusion

From the above results and especially the above example, it is evident that implementing the wireless link with low-height antennas provides similar performance to a single-hop link with high antennas, provided that repeater nodes utilize faster computers. The allowed delay of the nodes must be calculated based on the specific topology and the number of available repeater nodes.

In this chapter, validity of the provided model in chapter 4 was examined and different attributes of the multi-hop wireless link were tested and analyzed.

Chapter 6 provides the general conclusion and suggestions for future work.

Chapter 6

Conclusion and Proposals for the Future Work

6.1 Conclusion

In this research, a model for a point-to-point wireless link using 802.11a WLAN standard was proposed. In this work, a brief description of the 802.11 network has been provided; it included a description of different types of the networks, the MAC layer, procedures and issue of media access, packet transmission, fragmentation, and link adaptation together with a description of the PHY was provided. In order to model the physical environment, different environmental factors which affect the radio-wave propagation were studied and modeled using the two-ray model ray-tracing technique in detail which included different propagation effects such as ground reflection and polarization, roughness of the ground and radio-wave diffraction. The provided model is able to provide an estimate for the pathloss which in turn, affects the received power and the therefore the data transfer rate. Importantly, this model includes an accurate model for the parabolic reflector antennas used in the point to point link. Using the provided model, a model for the point-to-point wireless link was provided which combines the previously described models together with the techniques to enhance the data

transfer rate by minimizing the interference. Also a procedure for the implementing the link point-to-point link was provided.

Finally, the provided models were examined to verify the expectations and their applicability in a practical environment. Together with the simulation results, the field test results were also provided to confirm the feasibility of the proposed link.

6.2 Proposals for future work

During the present research the following subjects have been found interesting either as a continuation of this work or a research topic related to the present work:

- It is proposed that a research should be done using the provided model for a mesh network which uses the repeaters in “Smart-mode”. Such a research will entail detailed analysis of the frequency planning, interference suppression and optimizing the channel capacity.
- It is proposed that the proposed work may be continued also using the techniques such as power control which currently does not exist in the 802.11a standard.
- Another potential topic for continuation of this work is modeling the parallel operation of the channels using the provided interference models. Such a work will entail substantial modification to the MAC layer to ensure the operability of the parallel link. Our test result from the field test may prove useful as a start point to this topic.

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Appendix A

Antenna Characteristics

The antennas provided from the Pacific Wave Inc has the following characteristics

(Source: www.pacwireless.com)

Parameter		Min	Typ	Max	Units
Frequency Range	DA58 DA53	5725 5150		5850 5350	MHz
Gain	DA58 DA53		29 28		dBi
Input Return Loss (S_{11})			-14		dB
VSWR			1.5:1		
3db Beam Angle			6		Deg
Cross Pole			-32		dB
Front to Back		25			dB
Side Lobe			-28		dB
Impedance			50		OHM
Input Power				100	W
Pole Diameter (OD)		1.5 (38)		3.0 (76)	Inch (mm)
Operating Temperature		-40		+70	Deg C
Rated Wind Velocity				125	mph
Focal Length			10 (254)		Inch (mm)
Weight			11 (5)		Lbs (Kg)
Dimensions (Dia)			25.5 (648)		Inch (mm)

Pattern of the above antenna (same source):

Test setup pictures

One of the test setups used in our field measurements is shown below

Sample Test Results

Table I (Environment b: $h_1 \sim 1.5\text{m}$, $h_2 \sim 35\text{m}$)

Test number	Net Id.	Power Indicator	File Size (B)	Distance (m)	Duration (Sec)	Avg. D/R (kbps)
1	Net1	.8-FP f	~194 MB	~110	138.699	11229.807
1	Net2	.8-FP f	Same	~110	126.782	12285.374
2	Net1	.8-FP f	Same	~110	169.273	9201.493
2	Net2	.8-FP f	Same	~110	159.509	9764.743
3	Net1	.8-FP f	Same	~110	138.429	11251.742
3	Net2	.8-FP f	Same <--	~110	124.819	12478.565
4	Net1	.8-FP f	Same -->	~110	333.480	5813.832 *
4	Net2	.8-FP f	Same -->	~110	307.482	5065.556
5 S	Net1	.8-FP f	Same -->	~110	38.525	40429.642
6 S	Net2	.8-FP f	Same -->	~110	38.266	40075.453
7 S	Net1	.8-FP f	Same <--	~110	42.832	36364.936
8 S	Net2	.8-FP f	Same <--	~110	41.890	37182.126

S: Single line transmissions. --> Host to Client <-- Client to Host.

Table III Turbo Mode Results (194 MB file)

Test number	Net Id.	Power Indicator	File Size (B)	Distance (m)	Duration (Sec)	Avg. D/R (kbps)
1	Net1	same	194MB →	same	57.513	27082.158
2	Net1	same	Same ←	same	73.976	21054.939
3	Net1	same	Same →	same	59.045	26379.381
4	Net1	same	Same ←	same	71.413	21810.803
5	Net1	same	Same →	same	59.986	25965.414
6	Net1	same	Same ←	same	69.680	22353.095

Test number	Net Id.	Power Indicator	File Size (B)	Distance (m)	Duration (Sec)	Avg. D/R (kbps)
1	Net1	FP/FP	Same	Same	62.710	24837.565
1	Net2	FP/FP	Same	Same	74.808	20820.995
2	Net1	FP/FP	Same	Same	64.473	24158.564
2	Net2	FP/FP	Same	Same	74.177	20998.087
3 s	Net1	FP/FP	Same	Same	38.405	40556.149
3 s	Net2	FP/FP	Same	Same	38.865	40085.752
3 s	Net2	FP/FP	Same	Same	38.676	40272.613
4 <i>Rev</i>	Net1	FP/FP	Same	Same	71.553	21768.066
4 <i>Rev</i>	Net2	FP/FP	Same	Same	86.534	17999.402
5 <i>Rev</i>	Net1	FP/FP	Same	Same	72.534	21473.538
5 <i>Rev</i>	Net2	FP/FP	Same	Same	84.572	18417.150

S: Single line transmissions.

Modulation Techniques

Data Rate (Mbps)	Modulation	Coding Rate	Coded bits per subcarrier	Coded bits per OFDM symbol	Data bits per OFDM symbol
6	BPSK	1/2	1	48	24
9	BPSK	3/4	1	48	36
12	QPSK	1/2	2	6	48
18	QPSK	3/4	2	96	72
24	16-QAM	1/2	4	192	96
36	16-QAM	1/2	4	192	144
48	16-QAM	3/4	4	288	192
54	64-QAM	2/3	6	288	216

Frame Format

PLL preamble (12 symbols)	Rate (4 bits)	Reserved (1 bit)	Length (12 bits)	Parity 1 bit	Tail 6 bits	Service (16 bits)	PSBU (payload)	Tail (6 bits)	Pad
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OFDM Operating Bands and Channels

Band	Channel numbers	Frequency (MHz)	Maximum output power
U-NII lower band 95.15 to 5.25 MHz	36	5180	40mW (2.5mW/MHz)
	40	5200	
	44	5220	
	48	5240	
U-NII lower band 95.15 to 5.25 MHz	52	5260	200mW (12.5mW/MHz)
	56	5280	
	60	5300	
	64	5320	
U-NII lower band 95.15 to 5.25 MHz	149	5745	800mW (50mW/MHz)
	153	5765	
	157	5785	
	161	5805	

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