Wireless Communications Andrea Goldsmith

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Chapters 1-16, Appendices A-C

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Chapter 1

Overview of Wireless Communications

Wireless communication is one of the most impactful technologies in history, drastically affecting the way we live, work, play, and interact with people and the world. There are billions of cellphone subscribers worldwide, and a wide range of devices in addition to phones use cellular technology for their connectivity. Wireless network technology using the Wi-Fi standard has been incorporated into billions of devices as well, including smartphones, computers, cars, drones, kitchen appliances, watches, and tennis shoes. Satellite communication systems support video, voice, and data applications for receivers on earth, in the air, and in space. Revenue across all areas of wireless technology and services is trillions of dollars annually. The insatiable demand for wireless data along with new and compelling wireless applications indicate a bright future for wireless systems. However, many technical challenges remain in designing wireless networks and devices that deliver the performance necessary to support existing and emerging applications. In this introductory chapter we will briefly review the history of wireless communications, from the smoke signals of antiquity to the rise of radio communication that underlies the Wi-Fi, cellular, and satellite networks of today. We then discuss the most prevalent wireless communication systems in operation today. The impact of spectrum properties and regulation as well as standards on the design and success of wireless systems is also illuminated. We close the chapter by presenting a vision for the wireless communication systems of the future, including the technical challenges that must be overcome to make this vision a reality. Techniques to address many of these challenges are covered in subsequent chapters. The huge gap between the capabilities of current systems and the vision for future systems indicates that much research and development in wireless communications remains to be done.

1.1 History of Wireless Communications

1.1.1 Origins of Radio Technology

The first wireless networks were developed in antiquity. These systems transmitted information visually over line-of-sight distances (later extended by telescopes) using smoke signals, torch signaling, flashing mirrors, signal flares, or semaphore flags. An elaborate set of signal combinations was developed to convey complex messages with these rudimentary signals. Observation stations were built on hilltops and along roads to relay these messages over large distances. These early communication networks were replaced first by the telegraph network (invented by Samuel Morse in 1838) and later by the telephone.

The origins of radio communications began around 1820 with experiments by Oersted demonstrating that an electric field could move a compass needle, thereby establishing a connection between electricity and magnetism. Work by Amp´ere, Gauss, Henry, Faraday, and others further advanced knowledge about electromagnetic waves, culminating in Maxwell’s theory of electromagnetism published in 1865. The first transmission of electromagnetic

1

waves was performed by Hertz in the late 1880s, after which he famously declared to his students that these waves would be “of no use whatsoever.” He was proved wrong in 1895 when Marconi demonstrated the first radio trans mission across his father’s estate in Bologna. That transmission is considered the birth of radio communications, a term coined in the early 1900s. Marconi moved to England to continue his experiments over increasingly large transmission ranges, culminating in the first trans-Atlantic radio transmission in 1901. In 1900 Fessenden became the first person to send a speech signal over radio waves, and six years later he made the first public radio broad cast. From these early beginnings, described in more detail in [1], radio technology advanced rapidly to enable transmissions over larger distances with better quality, less power, and smaller, cheaper devices, thereby enabling public and private radio communications, television, and wireless networking.

1.1.2 From Analog to Digital

Radio systems designed prior to the invention of the transistor, including AM/FM radio, analog television, and amateur radio, transmitted analog signals. Most modern radio systems transmit digital signals generated by digital modulation of a bit stream. The bit stream may represent binary data (e.g., a computer file, digital photo, or digital video stream) or it may be obtained by digitizing an analog signal (e.g., by sampling the analog signal and then quantizing each sample). A digital radio can transmit a continuous bit stream or it can group the bits into packets. The latter type of radio is called a packet radio and is characterized by bursty transmissions: the radio is idle except when it transmits a packet. When packet radios transmit continuous data such as voice and video, the delay between received packets must not exceed the delay constraint of the data. The first wireless network based on packet radio, ALOHAnet, was developed at the University of Hawaii and began operation in 1971. This network enabled computer sites at seven campuses spread out over four islands to communicate with a central computer on Oahu via radio transmission. The network architecture used a star topology with the central computer at its hub. Any two computers could establish a bi-directional communications link between them by going through the central hub. ALOHAnet incorporated the first set of protocols for channel access and routing in packet radio systems, and many of the underlying principles in these protocols are still in use today.

The U.S. military saw great potential for communication systems exploiting the combination of packet data and broadcast radio inherent to ALOHAnet. Throughout the 70’s and early 80’s the Defense Advanced Research Projects Agency (DARPA) invested significant resources to develop networks using packet radios for communi cations in the battlefield. The nodes in these packet radio networks had the ability to configure (or reconfigure) into a network without the aid of any established infrastructure. Self-configuring wireless networks without any infrastructure were later coined ad hoc wireless networks.

DARPA’s investment in packet radio networks peaked in the mid 1980’s, but these networks fell far short of expectations in terms of speed and performance. This was due in part to the limited capabilities of the radios and in part to the lack of robust and efficient access and routing protocols. Packet radio networks also found commercial application in supporting wide-area wireless data services. These services, first introduced in the early 1990’s, enabled wireless data access (including email, file transfer, and web browsing) at fairly low speeds, on the order of 20 Kbps. The market for these wide-area wireless data services did not take off due mainly to their low data rates, high cost, and lack of “killer applications”. All of these services eventually folded, spurred in part by the introduction of wireless data in 2G cellular services [2], which marked the dawn of the wireless data revolution.

1.1.3 Evolution of Wireless Systems and Standards

The ubiquity of wireless communications has been enabled by the growth and success of Wireless Local Area Net works (WLANs), standardized through the family of IEEE 802.11 (Wi-Fi) protocols, as well as cellular networks. Satellite systems also play an important role in the wireless ecosystem. The evolution of these systems and their corresponding standards is traced out in this subsection.

2

Wi-Fi Systems

The success story of Wi-Fi systems, as illuminated in [3], began as an evolution of the Ethernet (802.3) standard for wired local area networks (LANs). Ethernet technology, developed at Xerox Parc in the 1970s and standardized in 1983, was widely adopted throughout the 1980s to connect computers, servers, and printers within office buildings. WLANs were envisioned as Ethernet LANs with cables replaced by radio links. In 1985 the Federal Communications Commission (FCC) enabled the commercial development of WLANs by authorizing for unlicensed use three of the Industrial, Scientific, and Medical (ISM) frequency bands: the 900 MHz band spanning 902-928 MHz, the 2.4 GHz band spanning 2.4-2.4835 GHz, and the 5.8 GHz band spanning 5.725-5.875 GHz. Up until then, these frequency bands had been reserved internationally for radio equipment associated with industrial, scientific and medical purposes other than telecommunications. Similar rulings followed shortly thereafter from the spectrum regulatory bodies in other countries. The new rulings allowed unlicensed use of these ISM bands by any radio following a certain set of restrictions to avoid compromising the performance of the primary band users. Such radios were also subject to interference from these primary users. The opening of these ISM bands to “free” use by unlicensed wireless devices unleashed a flurry of research and commercial wireless system development, particularly for WLANs.

The first WLAN product for the ISM band, called WaveLAN, was launched in 1988 by the NCR corporation and cost several thousand dollars. These WLANs had data rates up to 2 Mbps and operated in the 900 MHz and 2.4 GHz ISM bands. Dozens of WLAN companies and products appeared over the ensuing few years, mostly operating in the 900 MHz ISM band using direct-sequence spread spectrum, with data rates on the order of 1- 2 Mbps. The lack of standardization for these products led to high development costs, poor reliability, and lack of interoperability between systems. Moreover, Ethernet’s 10 Mbps data rate and high reliability far exceeded the capabilities of these early WLAN products. Since companies were willing to run cables within and between their facilities to get this better performance, the WLAN market remained small for its first decade of existence. The initial WLAN products were phased out as the 802.11 standards-based WLAN products hit the market in the late 1990s. In Europe a WLAN standard called HIPERLAN was finalized in 1996. An improved version was launched in 2000, but HIPERLAN systems never gained much traction.

The first 802.11 standard, finalized in 1997, was born from a working group within the IEEE 802 LAN/MAN Standards Committee that was formed in 1990. This standard, called 802.11-1997, operated in the 2.4 GHz ISM band with data rates up to 2 Mbps. The standard used channel sensing with collision avoidance for medium access as well as frequency hopping or direct-sequence spread spectrum to mitigate the main sources of interference in that band which, at the time, consisted primarily of microwave ovens, cordless phones, and baby monitors. Also in 1997, the FCC authorized 200 MHz of additional unlicensed spectrum in the 5 GHz band, with certain constraints to avoid interfering with the primary users, mostly radar systems, operating in this band. Two years later, in 1999, the standard that would ignite the Wi-Fi revolution, 802.11b, was finalized. This standard increased data rates to 11 Mbps and eliminated the frequency-hop option of its predecessor. A plethora of 802.11b products soon appeared. Their interoperability, coupled with dramatically lower costs relative to earlier WLAN products, led to widespread use of this new technology. Millions of 802.11b products were shipped in 2000, just one year after the standard was finalized, and these shipments grew tenfold by 2003. The year 1999 marked two other important milestones for WLANs. That year the 802.11a standard for the 5GHz ISM frequency band was finalized to capitalize on the new unlicensed spectrum in that band. This standard enabled 54 Mbps data rates in 20 MHz channels and introduced orthogonal-frequency-division-multiplexing (OFDM) coupled with adaptive modulation as a new physical layer design. Also in 1999 the Wireless Ethernet Compatibility Alliance was formed to facilitate interoperability and certification of WLAN products. The name of this group was later abbreviated to the Wi-Fi Alliance, thereby coining the widely used moniker for WLAN technology and standards today.

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The Wi-Fi standard has evolved rapidly since its early days, with new versions developed about every five years, as described in more detail in Appendix D.1. The 802.11g standard, introduced in 2003, is essentially the same in its physical layer and multiple access design as 802.11a, but operates in the 2.4 GHz ISM frequency band. Later standards provided improvements over these first OFDM Wi-Fi systems, including wider channels, multiple transmit and receive antennas to enable multiple spatial streams and improved robustness through beam forming, larger signal constellations, improved error-correction codes, and coordinated multiple access. These improvements have led to highly-reliable Wi-Fi products that are capable of 10 Gbps data rates within a building or outdoor area. Wi-Fi has also moved into the unregulated 60 GHz frequency band through the 802.11ad standard. Even its name has evolved; In 2018 the 802.11 standards body abandoned using letter suffixes for new generations of the standard. Instead it coined the sixth generation of the WiFi standard, originally named 802.11ax, as Wi-Fi 6. Today Wi-Fi technology is pervasive indoors and out even in remote corners of the world. In addition to its pervasiveness, Wi-Fi has experienced an explosion of applications beyond its original use of connecting computers to each other and their peripheral devices. In addition to computers, smartphones, and tablets, many electronic devices today, from medical devices to refrigerators to cars, are equipped with Wi-Fi, allowing them to download new software, exchange data with other devices, and take advantage of cloud-based storage and computation.

Cellular Systems

Cellular systems are another exceedingly successful wireless technology. The convergence of radio and tele phony began in 1915, when wireless voice transmission between New York and San Francisco was first established. The first analog mobile telephone system was deployed in St. Louis Missouri in 1946, launching AT&T’s Mobile Telephone Service (MTS). Within two years AT&T had deployed MTS over approximately 100 cities and highway corridors. Only six channels were allocated by the FCC for the service and, due to their close spacing in frequency, only three were usable at any given time. Hence only three people within a city could make a call simultaneously. The monthly service and per-call cost was very high, and the equipment bulky and heavy. Evolution of the system was slow; while the equipment improved and spectrum to support up to 12 channels was added, the system capacity remained extremely limited.

Ironically, about the same time MTS was first being deployed, a solution to this capacity problem had already emerged from researchers at AT&T Bell Laboratories: the notion of cellular systems. The cellular system concept, articulated in a 1947 Bell Laboratories Technical Memo by D. H. Ring [4], exploited the fact that the power of a transmitted signal falls off with distance. Thus, channels using the same frequency can be allocated to users at spatially-separate locations with minimal interference between the users. To exploit this principle of frequency reuse, a cellular system partitions a geographical area into non-overlapping cells, as shown in Fig. 1.3 below. Sets of channels are assigned to each cell, and cells that are assigned the same channel set are spaced far enough apart so that interference between the users in these cells is small. In early cellular systems the distance between cells using the same channel set was relatively large, but today sophisticated interference mitigation techniques allow channels to be reused in every cell. As a user moves between adjacent cells, its call is handed off to a channel associated with the new cell. Frequency reuse enables much more efficient use of spectrum as the number of simultaneous users is no longer limited to the number of available channels. Indeed, while many aspects of cellular system technology have changed over time, frequency reuse remains at the heart of cellular system design.

Although the cellular concept was introduced in the late 1940s, it was not implemented for several decades, as technology was not yet ripe to realize the system in practice. In the mid-1960s engineers at Bell Laboratories began work on a design and feasibility study for a metropolitan analog cellular system. The details of the design and analysis, along with successful system tests in Newark and Philadelphia, formed the basis of an AT&T FCC proposal in 1971 to approve cellular service and allocate spectrum for it [5]. The FCC approved experimental cellular licenses to telephone companies in 1974, which led to the construction of several cellular systems. Fol

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lowing a long and convoluted process, in 1981 the FCC finalized its ruling for the issuance of commercial cellular licenses, instituting a duopoly in every metropolitan area whereby one license would be granted to a traditional phone company, and the other to a non-traditional operator. The first generation cellular system for U.S. deploy ment, called the advanced mobile phone system (AMPS) and described in [6], was launched in Chicago in 1983. A similar service had been launched in Tokyo in 1979 and in Scandinavia in 1981, although those systems were incompatible with AMPS. Cellular service in many other countries launched in the ensuing years, mostly with incompatible standards, which precluded cellular roaming between countries. These first generation systems are referred to as 1G systems.

Like Wi-Fi, the exponential growth of the cellular industry exceeded all expectations, increasing by at least an order of magnitude every decade, from a modest 100,000 subscribers in 1984 to tens of millions in 1994, hundreds of millions in 2004, and billions in 2014. To meet the growing demand for wireless data along with a diverse requirements for different types of wireless devices, a new generation of cellular systems and standards has emerged approximately every decade. In fact, that timeline was compressed for 1G systems, since the analog cellular system deployed in Chicago in 1983 was already saturated by 1984. At that point the FCC increased the cellular spectral allocation from 40 MHz to 50 MHz. As cellular systems throughout more and more cities became saturated with demand, the development of digital cellular technology for increased capacity and better perfor mance became essential. These enhanced system requirements launched the process to create a second generation (2G) cellular standard in the late 1980s with deployments in the early 1990s. In addition to voice communication, the move to digital technology paved the way for these systems to support low-rate data as well, in particular short texts, voice mail, and paging services. Unfortunately, the great market potential for cellular phones led to a proliferation of 2G cellular standards, with three different standards in the U.S. alone. One of these matched the European standard; Japan adopted a separate standard. Hence global roaming required a multi-mode phone. This deficiency was corrected for the third generation of cellular standards (3G), for which seven telecommunications standards bodies across Asia, Europe and the United States formed the third-generation partnership project (3GPP) to develop a single worldwide standard. The 3G cellular systems based on the 3GPP standard, whose deployments began in the early 2000s, provided an order of magnitude higher peak data rates than 2G systems. Indeed, it was the capabilities of 3G systems that transitioned the “killer application” of cell phones from voice to wireless data. The proliferation of smart phone technology in the mid-2000s, which were designed to consume vast amounts of wireless data, greatly stressed the capacity of 3G networks. Moreover, the 3G networks had far lower data rates than Wi-Fi, which became the access mode of choice for high-speed data. These developments paved the way for the next-generation 4G “long term evolution (LTE)” cellular standard. These systems, deployed in the early 2010s, supported an order-of-magnitude peak data rate increase over 3G systems. The 5G cellular standard supports higher data rates than 4G systems, as well as lower latency and better energy efficiency. System deployments for 5G began in 2019, continuing the trend of new cellular systems and standards every decade. Starting with the 3G systems, each generation of cellular standard has been developed to meet a set of International Mobile Telecom munications (IMT) requirements outlined by the International Telecommunications Union (ITU). The ITU certifies which cellular standards meet its IMT requirements; the 3GPP standard is designed to meet these requirements, as are alternate standards developed by specific countries or other standards organizations. More details on current cellular technology will be given in Chapter 15, with the evolution of cellular standards described in Appendix D.2.

Satellite Systems

Satellite communication systems are another major component of today’s wireless communications infras tructure. Commercial satellite systems can provide broadcast services over very wide areas, fill the coverage gap in locations without cellular service, and provide connectivity for aerial systems such as airplane Wi-Fi. Satellite systems are typically categorized by the height of the satellites’ orbit: low-earth orbit (LEO) satellites operate at

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roughly 2000 km altitude, medium-earth orbit (MEO) satellites at roughly 9000 km altitude, and geosynchronous orbit (GEO) satellites at roughly 40,000 km altitude. GEO satellites with geostationary orbits are seen as station ary from the earth, whereas GEO satellites with other orbits (such as elliptical) have their coverage area change over time. The disadvantage of high altitude orbits is that it takes a great deal of power to reach the satellite, and the propagation delay is typically too large for two-way delay-constrained applications like voice, video, and gaming. However, satellites at these orbits tend to have larger coverage areas, so fewer satellites are necessary to provide wide-area or global coverage. In addition to communication services, satellites are used for many other applications including weather monitoring, earth observation, surveillance, imaging, navigation, and localization.

The concept of using GEO satellites for communications was first suggested by Herman Potocnik in 1928 and later popularized through a 1945 article in Wireless World written by the science fiction writer Arthur C. Clarke. However, the first deployed satellites, the Soviet Union’s Sputnik in 1957 and the Nasa/Bell Laboratories’ Echo-1 in 1960, were not geosynchronous due to the difficulty of lifting a satellite into such a high orbit. Following these launches, in 1962 the Communications Satellite Corporation (Comsat) was formed in the United States to develop commercial communication satellite systems. Two years later, the International Telecommunications Satellite Consortium (IntelSat) emerged as a public-private consortium of 18 countries with the goal of enabling global telecommunications connectivity. SYNCOM 2, the first communication satellite to successfully reach geosyn chronous orbit, was launched in 1963. The following year SYNCOM 3 was launched into geostationary orbit, providing a two-way 10 MHz communication channel at a carrier frequency of 1.8 GHz for the satellite-to-earth link (downlink) and at 7.3 GHz for the reverse link (uplink). Shortly after its launch, SYNCOM 3 was used to provide live television coverage to US viewers of the 1964 Summer Olympics in Tokyo. The IntelSat consortium launched a number of satellites during the late 1960s in order to reach near-global coverage, culminating in the broadcasting of the 1969 landing of the first human on the moon to 600 million viewers. In addition to television broadcasting, these early systems also supported voice, teletype, and paging.

GEO satellite technology has evolved continuously since its early days. The C band of the radio spectrum (4-8 GHz) was used in early GEO systems, but subsequent systems have taken advantage of the larger bandwidths available at the higher frequency X (8-12.5 GHz), Ku (12.5-18 GHz), and Ka (26.5-40 GHz) bands in order to provide higher data rates. Antenna system technology for GEOs has also evolved to focus the transmission energy more precisely, leading to higher signal quality and hence higher data rates. There are hundreds of GEO satellites deployed today, with some capable of a total data rate in excess of one hundred Gbps. For individual users, typical data rates are on the order of ten Mbps in the downlink and several Mbps in the uplink [7]. Due to their large coverage regions, geosynchronous satellites are the primary communications mechanism for vessels at sea and in the air. They are also well-suited for broadcast entertainment where round-trip delay is not a consideration; they support hundreds of channels of digital television and radio at very high quality.

As cellular systems were rolled out in the late 1980s, the focus for satellite technology turned to building LEO systems that might compete with them. This led to the launch of several LEO communication systems in the late 1990s, including Globalstar and Iridium [8]. These LEOs provided global coverage but the link rates remained low due to power and bandwidth constraints. The handsets for these systems were also much larger and heavier than their cellular counterparts, primarily due to their large batteries. The LEO satellite systems deployed in the 1990s did not experience significant commercial success as, for most users, they provided worse performance and coverage than the competing 2G cellular phone technology at a higher cost. While these satellite systems provided coverage to remote areas without access to cellular service, the cost was prohibitive for the large majority of people in these areas. As a result, pretty much all the initial LEO companies, including Globalstar and Iridium, filed for bankruptcy protection within a few years of launching their service. Teledesic, which was planning a very ambitious LEO system with more than 800 satellites, declared bankrupcy before a single launch. After several iterations of restructuring, Iridium and Globalstar emerged as profitable services and today provide global Internet connectivity, albeit for a relatively small number of subscribers compared to that of cellular service. Interest in

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LEO satellites resurfaced as 4G cellular systems rolled out due to increased demand for connectivity in remote locations not served by cellular, better technology for both satellites and ground transceivers, and reduced launch costs. Significant commercial development of LEO satellites emerged concurrent with 5G cellular rollouts as a means to provide broadband Internet service to the billions of people worldwide that do not have access to high speed connectivity. It is anticipated that such systems will each deploy hundreds to thousands of satellites.

Today we take for granted the magic of wireless communications, which allows us and our devices to be connected anywhere in the world. This magic has been enabled by the pioneers that contributed to advances in wireless technology underlying the powerful systems ubiquitous today.

1.2 Current Systems

This section provides a design overview of the most prevalent wireless systems in operation today. The design details of these systems are constantly evolving, incorporating new technologies and innovations. This section will focus mainly on the high-level design aspects of these systems. More details on these systems and their underlying technologies will be provided in specific sections of the book. A summary of Wi-Fi, cellular, and short-range networking standards can be found in Appendix D with more complete treatments of recent standards in [9, 10, 11, 12, 13].

1.2.1 Wireless Local Area Networks

WLANs support data transmissions for multiple users within a “local” coverage area whose size depends on the radio design, operating frequency, propagation characteristics, and antenna capabilities. A WLAN consists of one or more access points (APs) connected to the Internet that serve one or more WLAN devices, also called clients. The basic WLAN architecture is shown in Figure 1.1. In this architecture clients within the coverage area of an AP connect to it via single-hop radio transmission. For better performance and coverage, some WLANs use relay nodes to enable multi-hop transmissions between clients and APs, as shown in Figure 1.2. WLANs today follow the 802.11 Wi-Fi family of protocols. Handoff of a moving client between APs is not supported by the 802.11 protocol, however some WLAN networks, particularly corporate and campus systems, incorporate this feature into their overall design.

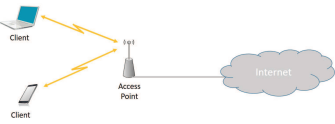
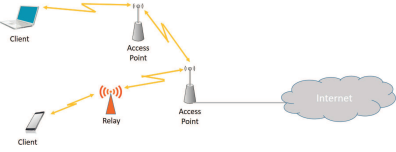


Figure 1.1: Basic wireless LAN architecture

While early WLANs were exclusively indoor systems, current WLANs operate both indoors and outdoors. Indoor WLANs are prevalent in most homes, offices, and other indoor locations where people congregate. Outdoor systems are generally deployed in areas with a high density of users such as corporate and academic campuses, sports stadiums, and downtown areas. The range of an indoor WLAN is typically less than 50 meters and can

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Figure 1.2: Wireless LAN architecture with multihop transmissions

be confined to a single room for those operating at 60 GHz. Outdoor systems have a bigger range than indoor systems due to their higher power, better antennas, and lower density of obstructions between the transmitter and receiver. WLANs generally operate in unlicensed frequency bands, hence share these bands with other unlicensed devices, such as cordless phones, baby monitors, security systems, and Bluetooth radios. Interference between WLAN devices is controlled by the WLAN access protocol, whereas interference between WLAN devices and other unlicensed devices is mitigated by a limit on the power per unit bandwidth for such devices.

WLANs use packet data transmission for better sharing of the network resources. Hence, data files are seg mented into packets. Sharing of the available bandwidth between different APs and, for most systems, between clients accessing the same AP is typically controlled in a distributed manner using the carrier sense multiple access with collision avoidance (CSMA/CA) protocol. In CSMA/CA, energy from other transmissions on a given channel is sensed prior to a transmission and, if the energy is sensed above a given threshold, the transmitter waits a random backoff time before again attempting a transmission. A transmitted packet is acknowledged by the receiver once received. If a transmitted packet is not acknowledged, it is assumed lost and hence retransmitted. As discussed in more detail in Chapter 14, the CSMA/CA access protocol is very inefficient, which leads to poor performance of WLANs under moderate to heavy traffic loads. As a result, some WLANs use more sophisticated access protocols, similar to those in cellular systems, that provide centralized scheduling and resource allocation either to all clients served by a given AP or, more generally, to all APs and clients within a given system.

1.2.2 Cellular Systems

Cellular systems provide connectivity to a wide range of devices, both indoors and out, with a plethora of features and applications including text messages, voice calls, data and video transfer, Internet access, and mobile “apps” (application software tailored to run on mobile devices). Cellular systems typically operate in licensed frequency bands, whereby the cellular operator must purchase or lease the spectrum in which their system operates. These licenses typically grant the operator exclusive use of the licensed spectrum. Starting in 2017 regulators in several countries began allowing cellular systems to operate in the 5 GHz unlicensed band in addition to within their licensed spectrum. Given the large amount of spectrum in these unlicensed bands compared to what is available in the licensed bands, this development has the potential to drastically increase cellular system capacity. However, it will also increase interference in the unlicensed bands, particularly for Wi-Fi systems.

The basic premise behind cellular system design is frequency reuse, which exploits the fact that signal power falls off with distance to reuse the same frequency spectrum at spatially separated locations. Specifically, a cellular

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system consists of multiple cells, where each cell is assigned one or more channels to serve its users. Each channel may be reused in another cell some distance away, and the interference between cells operating on the same channel is called intercell interference. The spatial separation of cells that reuse the same channel set, the reuse distance, should be as small as possible so that frequencies are reused as often as possible, thereby maximizing spectral efficiency. However, as the reuse distance decreases, the intercell interference increases owing to the smaller propagation distance between interfering cells. Early cellular systems had reuse distances greater than one, but current systems typically reuse each channel in every cell while managing the resulting interference through sophisticated mitigation techniques.

The most basic cellular system architecture consists of tessellating cells, as shown in Figure 1.3, where Ci denotes the channel set assigned to cell i. The two-dimensional cell shapes that tessellate (cover a region without gaps or overlap) are hexagons, squares, and triangles. Of these, the hexagon best approximates an idealized omni directional transmitter’s circular coverage area. Early cellular systems followed this basic architecture using a relatively small number of cells to cover an entire city or region. The cell base stations in these systems were placed on tall buildings or mountains and transmitted at very high power with cell coverage areas of several square miles. These large cells are called macrocells. The first few generations of macrocell base stations used single antenna omni-directional transmitters, so a mobile moving in a circle around these base station had approximately constant average received power unless the signal was blocked by an attenuating object. In deployed systems cell coverage areas overlap or have gaps as signal propagation from a set of base stations never creates tessellating shapes in practice, even if they can be optimally placed. Optimal base station placement is also impractical, as zoning restrictions, rooftop availability, site cost, backhaul and power availability, as well as other considerations influence this placement.

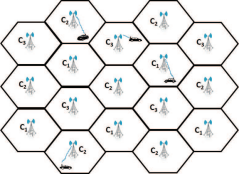


Figure 1.3: Cellular network architecture (homogeneous cell size).

Macrocells have the benefit of wide area coverage, but they can often become overloaded when they contain more users than channels. This phenomenon has led to hierarchical cellular system architectures, with macrocells that provide wide area coverage and small cells embedded within these larger cells to provide high capacity, as shown in Figure 1.4. Not only do the small cells provide increased capacity over a macrocell-only network, but they also reduce the transmit power required at both the base station and mobile terminal, since the maximum transmission distance within the small cell is much less than in the macrocell. Cellular systems with heterogeneous cells sizes are referred to as Hetnets. In current cellular systems, channels are typically assigned dynamically based on interference conditions. Another feature of many current cellular system designs is for base stations in

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adjacent macrocells to operate on the same frequency, utilizing power control, adaptive modulation and coding, as well as interference mitigation techniques to ensure the interference between users does not preclude acceptable performance.

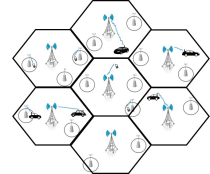


Figure 1.4: Hierarchical cellular network architecture (a Hetnet with macrocells and small cells).

All base stations in a given geographical area, both macrocells and small cells, are connected via a high-speed communications link to a mobile telephone switching office (MTSO), as shown in Figure 1.5. The MTSO acts as a central controller for the cellular system, allocating channels within each cell, coordinating handoffs between cells when a mobile traverses a cell boundary, and routing calls to and from mobile users. The MTSO can route calls to mobile users in other geographic regions via the local MTSO in that region or to landline users through the public switched telephone network (PSTN). In addition, the MTSO can provide connectivity to the Internet.

A new user located in a given cell requests a channel by sending a call request to the cell’s base station over a separate control channel. The request is relayed to the MTSO, which accepts the call request if a channel is available in that cell. If no channels are available then the call request is rejected. A call handoff is initiated when the base station or the mobile in a given cell detects that the received signal power for that call is approaching a given minimum threshold. In this case the base station informs the MTSO that the mobile requires a handoff, and the MTSO then queries surrounding base stations to determine if one of these stations can detect that mobile’s signal. If so then the MTSO coordinates a handoff between the original base station and the new base station. If no channels are available in the cell with the new base station then the handoff fails and the call is dropped. A call will also be dropped if the signal strength between a mobile and its base station falls below the minimum threshold needed for communication due to signal propagation effects such as path loss, blockage, or multipath fading. These propagation characteristics are described in Chapters 2-3.

Spectral sharing in cellular systems, also called multiple access, is done by dividing the signaling dimensions along the time, frequency, code, and/or the spatial dimensions. Current cellular standards use a combination of time and frequency division for spectral sharing within a cell. However, there are still 2G and 3G systems in operation that use code-division multiple access based on direct sequence spread spectrum. Channels assigned to users within a cell may be orthogonal or non-orthogonal. In the latter case the channels of two users within a cell will overlap in time, frequency, code or spatial dimensions, which is referred to as intracell interference. More details on multiple access techniques and their performance analysis will be given in Chapters 13 and 14.

Efficient cellular system designs are interference limited – that is, interference from within and outside a cell dominates the noise floor, since otherwise more users could be added to the system. As a result, any tech

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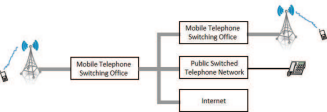


Figure 1.5: Cellular network architecture evolution.

nique to reduce interference in cellular systems leads directly to an increase in system capacity and performance. Some methods for interference reduction in cellular systems include cell sectorization, power control, directional antennas and antenna-array signal processing, multiuser detection and interference cancellation, base station co operation, and user scheduling. Details of these techniques will be given in Chapters 14 and 15.

1.2.3 Satellite Systems

Satellite systems are another major component of the wireless communications infrastructure [14, 15]. There is a plethora of satellite communications services available today, including mobile service to airplanes, ships, vehicles, and hand-held terminals, fixed satellite service to earth stations, as well as broadcast radio and television service. Fig. 1.6 illustrates the satellite architecture supporting all such services. One of the biggest advantages of satellites over terrestrial wireless systems is their ability to provide coverage in remote areas.

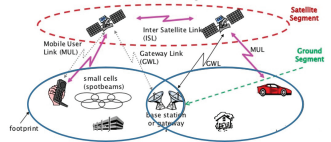


Figure 1.6: Satellite System Architecture

Satellite communication systems consist of one or more satellites communicating with stations on earth or in the air. Communication satellites typically serve as relays between stations on the ground and in some cases provide direct links to other satellites as well. Ground stations may be at fixed locations, in which case their antennas may be large to maximize received signal strength. Mobile stations and indoor fixed stations have smaller antennas, typically on the order of 1-3 meters. The coverage area or footprint of a satellite depends on its orbit. A GEO satellite can cover one or more continents, hence only a handful are needed to provide coverage for the entire globe. However, the transmission latency between the earth and a GEO satellite is high, on the order of 300 ms.

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MEO satellites have a coverage area of around ten thousand square kilometers and a latency of around 80 ms. A LEO coverage area is about a thousand square kilometers with a latency of around 10 ms. A GEO satellite provides continuous coverage to any station within its footprint, whereas a satellite with a different orbit will have a fixed station within its footprint for only part of a 24 hour period. GEO satellites are heavy, sophisticated, and costly to build and launch. Portable stations or handsets for GEOs tend to be large and bulky due to the power required to reach the satellite. In addition, the large round-trip propagation delay to a GEO satellite is quite noticeable in two-way voice communication. The most common service offered by GEO satellites is radio and TV broadcasting since delay is not a constraint and the satellite construction, launch, and operational costs are amortized over the many users within the GEO footprint. GEO satellites are also commonly used as a backbone link for terrestrial networks, for airplane and maritime communications, and for connectivity in remote locations.

LEO satellites are much lighter and lower in cost than GEOs to manufacture, launch, and operate. In addition, the stations and handsets in a LEO system have smaller size, transmit power, and latency than GEO systems due to the closer proximity of LEO satellites to the earth. Hence, most mobile services use LEO satellites, typically in a constellation of dozens to hundreds of satellites whose total footprint covers the locations around the globe supported by the system. Continuous coverage of a given location in a LEO system requires handoff between satellites as a fixed location on earth is within the footprint of a given LEO satellite for only 10-40 minutes. Hence, when the footprint of a LEO satellite moves away from a given station or handset, its connection is handed off to another LEO satellite so as to maintain continuous connectivity. Sophisticated antennas on LEO satellites can create very small spotbeams for mobile terminals that focus that transmission energy within a small footprint. These spotbeams allow for frequency reuse similar to that of cellular systems.

For the reasons outlined in the previous paragraph, LEOs tend to support the highest data rates, best per formance, and lowest cost among satellite services. As expected, MEO satellite systems provide a compromise between the system requirements, performance, and costs compared with those of GEO and LEO systems. MEO satellite services mainly compete with GEO systems by offering better performance at a lower cost for users in relatively remote locations that are not well served by LEO or terrestrial systems.

1.2.4 Fixed Wireless Access

Fixed wireless access (FWA) systems support wireless communications between a fixed access point and multiple terminals. The FWA system architecture is shown in Fig 1.7. The access point transmits to the receivers of multiple terminals in the downlink direction, and receives signals from the transmitters of multiple terminals in the uplink direction. Different multiple access techniques can be used to share the system spectrum between the multiple terminals.

FWA systems provide an alternative to the wired broadband options of DSL, cable and fiber, whose availability may be limited in rural areas. In the United States, two frequency bands were set aside for these systems in the late 1990s: part of the 28 GHz spectrum for local distribution systems (local multipoint distribution service, LMDS) and a band in the 2 GHz spectrum for metropolitan distribution service (multichannel multipoint distribution services, MMDS). MMDS systems were never widely deployed, and are used today in sparsely populated rural areas, where laying cables is not economically viable. The initial LMDS systems were expensive and performed poorly due to the challenges of providing service at such high frequencies. As a result these systems did not succeed commercially and were discontinued. The LMDS spectrum is just below the millimeter wave band of 30-300 GHz. Recent interest in utilizing the large amount of unregulated millimeter wave spectrum has led to advances in system, circuit, and antenna design at these high frequencies, which in turn has renewed interest in FWA systems using the LMDS and millimeter wave bands.

One of the main growth drivers of FWA was the IEEE 802.16 (WiMAX) standard, finalized in 2001. The first WiMAX standard defined operation between 2 GHz and 11 GHz for non-line-of-sight links and between

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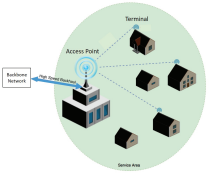


Figure 1.7: Fixed Wireless Access System

10 GHz and 66 GHz for line-of-sight links, however WiMAX systems generally operate below 6 GHz. The first commercial WiMAX systems followed the 802.16d standard with data rates of around 40 Mbps. A later version, 802.16e, was developed in the mid-2000s to support mobile users with a system design similar to that of cellular. The 802.16e systems offered 15 Mbps data rates, much higher than that of 3G cellular systems. The significantly higher data rates of 802.16e over 3G led to speculation that it would be adopted for 4G cellular. However, in the end 4G cellular adopted the LTE standard, which ultimately pushed WiMAX systems out of most mobile service markets. WiMAX is still used to support mobile services in areas where LTE has not been deployed, as well as in industrial markets such as aviation, utilities, and transportation. FWA systems based on the LTE and 5G cellular standards have been deployed to provide high-speed connectivity for homes, apartments, and office buildings.

1.2.5 Short Range Radios with Multihop Routing

As radios decrease their cost and power consumption, it becomes feasible to embed them into more types of electronic devices, which enables applications such as smart homes, sensor networks, vehicular networks, and other interconnected systems. The most common radio standards that have emerged to support this trend are Bluetooth, ZigBee, and Z-Wave. All of these radios support a multihop routing protocol, whereby a given radio can communicate with any other radio in its transmission range. If the destination radio is not within this range, intermediate radios relay the message to this destination. Radios with multihop routing form an ad hoc wireless network since they can reconfigure and have no established infrastructure. Such ad hoc wireless networks are commonly used by the military as well as for emergency response. In principle a multihop routing protocol can support hundreds or even thousands of nodes, but network performance generally degrades as the number of nodes increases. The short range radios described in this section have not yet demonstrated that their multihop routing protocols are feasible in practice for large numbers of nodes. In addition to Bluetooth, Zigbie, and Z Wave, proprietary radios have been used in a number of devices, products, and military systems to support their communication requirements.

The Bluetooth standard is based on a small radio transceiver microchip built into digital devices.1 The standard and device certification is managed by the Bluetooth Special Interest Group. In most applications the Bluetooth



1The Bluetooth standard is named after Harald I Bluetooth, the king of Denmark between 940 and 985 A.D. who united Denmark and Norway. Bluetooth proposes to unite devices via radio connections, hence the inspiration for its name.

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radio takes the place of a connecting cable for electronic devices such as cell phones, tablets, headsets, audio equipment, cameras, watches, and smart meters. Bluetooth has also been integrated into larger devices such as cars and medical equipment. Bluetooth is mainly for short-range communications – for example, from a laptop to a nearby printer or from a cell phone to a wireless headset. Its normal range of operation is 10 m (at 2.5-mW transmit power), and this range can be increased to 100 m by increasing the transmit power to 100 mW. The system operates in the unlicensed 2.4-GHz frequency band, so it can be used worldwide without any licensing issues. The Bluetooth standard provides one data channel at 723.2 kbps. In this mode, there is a reverse channel with a data rate of 57.6 kbps. The specification also allows up to three additional channels each at a rate of 64 kbps, which are primarily used for voice connections with headsets. These different modes result in an aggregate bit rate of approximately 1 Mbps. An enhanced data rate mode provides up to 3 Mbps, while a low-energy mode significantly reduces power consumption at the expense of range or data rate.

Bluetooth uses frequency hopping for multiple access with a carrier spacing of 1 MHz. Typically, up to eighty different frequencies are used for a total bandwidth of 80 MHz. At any given time, the bandwidth available is 1 MHz. Bluetooth radios form small ad hoc networks of up to eight devices sharing the same logical channel (same hop sequence), which is called a piconet. Different channels (different hopping sequences) can simultaneously share the same 80 MHz bandwidth. Collisions will occur when devices in different piconets that are on different logical channels happen to use the same hop frequency at the same time. As the number of piconets in an area increases, the number of collisions increases and performance degrades. The original Bluetooth standard was developed jointly by 3 Com, Ericsson, Intel, IBM, Lucent, Microsoft, Motorola, Nokia, and Toshiba. Many additional manufactures have contributed to each new generation of the standard. Bluetooth is integrated into a wide range of electronic devices with several billion Bluetooth-enabled devices shipped annually.

The ZigBee2 radio specification is designed for lower cost and power consumption than Bluetooth. It follows the IEEE 802.15.4 standard with device certification managed by the Zigbie Alliance. The radio operates in the same 2.4 GHz ISM band as Bluetooth. Zigbee radios support data rates of up to 250 kbps at a range of up to 30 m. These data rates are slower than Bluetooth, but in exchange the radio consumes significantly less power with a larger transmission range. Zigbee also operates in a “green” mode whereby the radio is powered through energy harvesting of its environment, reducing or in some cases completely elimiating the need for battery power. The goal of ZigBee is to provide radio operation for months or years without recharging, thereby targeting devices such as smart tags, meters, lights, and thermostats, as well as those used for sensing and automation.

Z-Wave radios are designed primarily for smart home applications with operation in the 900 MHz ISM band. Since power falls off more slowly in this band than at the higher 2.4 GHz spectrum, Z-Wave radios have a higher range than standard Zigbee or Bluetooth radios, on the order of 100 m. In addition, there is no interference between Z-Wave radios and those operating in the crowded 2.4 GHz band, including Wi-Fi as well as Bluetooth and Zigbee. On the downside, Z-Wave has significantly lower data rates than either Bluetooth or Zigbee, ranging from 10 to 100 Kbps. The Z-Wave standard and device certification is managed by the Z-Wave Alliance.

1.3 Wireless Spectrum

1.3.1 Regulation

Most countries have government agencies responsible for allocating and controlling use of the radio spectrum. In the United States, spectrum is allocated by the Federal Communications Commission (FCC) for commercial use and by the Office of Spectral Management (OSM) under the auspices of the National Telecommunications and Information Administration (NTIA) for government use. Countries throughout the world have similar regula



2ZigBee takes its name from the dance that honey bees use to communicate information about newly found food sources to other members of the colony.

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tory agencies to regulate spectrum within their borders. Certain regions of the world have a common regulatory agency for spectrum allocation, e.g. commercial spectral allocation across Europe is governed by the European Telecommunications Standards Institute (ETSI). Satellite systems cover large areas spanning many countries and sometimes the globe. Globally, spectrum is allocated by the International Telecommunications Union Radio Com munications group (ITU-R) through its World Radiocommunication Conferences (WRC). The standards arm of this body, ITU-T, adopts telecommunication standards for global systems that must interoperate across national boundaries. Regulatory agencies typically have many competing considerations in deciding how to allocate any given block of spectrum, including whether to allocate it for commercial, military, or shared use. These deci sions are generally driven by a broad mandate to regulate spectrum “in the public interest,” as stated in the U.S. Communications Act of 1934 establishing the FCC.

Historically government agencies allocated spectral blocks for specific uses and assigned licenses to use these blocks to specific groups or companies. For example, in the 1980s the FCC allocated spectrum around 850 MHz for analog cellular phone service, in particular 824-849 MHz for the downlink (base station to mobile) and 869-894 MHz in the uplink (mobile to base station). Spectral licenses were provided to two operators in each geographical area based on a number of criteria. While this method of licensed spectral allocation is still used in some circum stances, a fundamental shift occurred worldwide in the early 1990s for licensed spectrum to be auctioned to the highest bidder, with some restrictions in place to ensure fairness and competitive use of the spectrum. The basis for the shift was the market-based reasoning that auctions provide the fairest and most efficient way for governments to allocate the limited spectral resource and, moreover, this method provides significant revenue to the government. However, auctions are not universally supported for spectral allocation based on the contention that they can stifle innovation, limit competition, and hurt technology adoption. Specifically, the high cost of spectrum dictates that only large companies or conglomerates can purchase it. Moreover, the large investment required to obtain spectrum can delay, sometimes indefinitely, the ability to invest in infrastructure for system rollout. Finally, high spectral cost is usually passed on to the end user. The early 3G spectral auctions, with suspected collusion between bidders, low bids, and several auction winners that ultimately defaulted, provided ammunition to the opponents of spectral auctions. Lessons learned from these early auctions were adopted in the design of subsequent auctions, which generally went smoothly and raised up to tens of billions of dollars. In addition, reverse or incentive auctions were initiated starting in 2016, whereby license holders could sell back their spectrum to regulatory bodies for future auctioning. A comprehensive treatment of spectrum regulation and its allocation through auctions can be found in [16].

In addition to spectral auctions, spectrum can be set aside in specific frequency bands, called unlicensed bands, that are free to use without a license according to a specific set of rules. The rules may correspond to a spe cific access protocol to ensure fairness, restrictions on power levels, and so forth. The purpose of these unlicensed bands is to encourage innovation and low-cost implementation. Wi-Fi is often associated with the unlicensed fre quency bands, however it is just one of the hundreds of successful unlicensed systems, which include standardized short-range radio systems as well as proprietary radios in cordless phones, wireless home security systems, baby monitors, medical equipment, inventory systems, smart meters, and keyless automobile entry systems. Indeed, it is estimated that unlicensed wireless devices contribute tens of billions of dollars annually to the US economy alone. A major difficulty of unlicensed bands is that they can be killed by their own success. If many unlicensed devices on the same channel are used in close proximity then they interfere with each other, which can make the band unusable. Cellular systems can also operate in the unlicensed bands by following the unlicensed spectrum rules. However, since these systems can use their licensed bands for control, they have an advantage over systems such as Wi-Fi that use the unlicensed bands for both control and data transmission and can crowd out Wi-Fi users as a result [17]. Much of the spectrum above 30 GHz is unregulated or lightly regulated, and hence can be used by unlicensed users with little to no restriction.

Underlay systems are another alternative for allocating spectrum. An underlay system operates as a secondary 15

user in a licensed frequency band simultaneous with the licensed users in a manner such that the licensed users experience minimal interference from them. This is usually accomplished by spreading the signal over a very wide bandwidth, typically more than 500 MHz, and restricting its power per Hertz. The first underlay standard approved for operation was ultrawideband (UWB) communications [18]. Specifically, in 2002 the FCC approved 7500 MHz of spectrum for the use of UWB devices, with the very stringent power restriction of no more than 75 nW/MHz. This sparked regulatory activities in countries throughout Europe and Asia to also enable UWB, albeit with different restrictions than those in the US. Indeed, regulatory approval of UWB proved to be quite contro versial given the complexity of characterizing how interference affects the primary band users, and the fact the UWB transmissions span many licensed users across both commercial and government domains. The regulatory challenges coupled with the severe power constraints on UWB systems ultimately proved insurmountable for most commercial systems, hence the technology failed to achieve much success [19]. The interference constraint for underlay users may alternatively be met without restricting power per Hertz by using multiple-antenna techniques to guide the underlay signals away from the spatial dimensions occupied by licensed users [20].

Following the introduction of underlay systems, regulatory bodies began exploring other innovative technolo gies that could make spectrum utilization in the licensed bands more flexible and efficient. This push for innovation was long overdue; other than spectral auctions and underlay systems, the basic mechanisms for licensed spectral allocation had not changed much since the inception of regulatory bodies in the early to mid-1900s. Many of the compelling ideas for exploiting technology to better utilize licensed spectrum fall under the notion of a cognitive radio. A cognitive radio utilizes advanced radio and signal processing technology along with novel spectrum al location policies to support unlicensed users operating in the existing licensed spectrum, without degrading the performance of the licensed users. In particular, a cognitive radio “learns” about coexisting licensed users within its spectrum and then uses this information to utilize the spectrum without degrading the transmissions of these users [21]. Based on the nature of the coexisting user information the cognitive radio can collect, as well as a priori rules about spectrum usage, a cognitive radio seeks to overlay or interweave its signal with the transmissions of licensed nodes. Spatial underlay systems that dynamically avoid the spatial dimensions of licensed users also fall within the paradigm of cognitive radios. Cognitive radio techniques can also be applied in the unlicensed bands to reduce interference between users and thus improve spectral efficiency in these bands as well.

Interweave cognitive radios utilize unused parts of the licensed spectrum. The idea of interweave radios came about after studies conducted by the FCC and industry showed the existence of space-time-frequency voids, referred to as spectrum holes, in both the licensed and unlicensed bands that are not in constant use. These spectrum holes, which can change with time and geographic location, can be exploited by interweave radios to support their communication. In some cases the spectrum holes are permanent, either because a certain block of licensed spectrum is unused, or because “guard channels” between occupied channels are needed in the overall design to reduce interference between the licensed users. Such guard channels, or “white spaces,” were specified in the analog television broadcast standard. As analog television bands transitioned to digital, arguments were made that such guard bands were no longer needed, and hence these white spaces could be freed up for other uses. Despite resistance from the television broadcasters, unlicensed use of television white spaces was approved by regulatory bodies in multiple countries starting in 2010. White space devices must generally consult a database of available spectrum holes in their geographical area before using them, and such databases are relatively static. A more sophisticated interweave radio looks for dynamic spectrum holes by periodically monitoring the radio spectrum, detecting spectrum holes as they occur in time, space, and/or frequency, and then using such holes opportunistically to communicate over them. Such opportunistic use of white spaces, which was the original motivation behind the notion of cognitive radio [22], has yet to be approved for spectrum allocation.

In overlay systems the cognitive transmitter has some knowledge about the transmissions of noncognitive users in the band. This knowledge can be exploited in two ways: to mitigate the interference caused to licensed users and to improve the performance of licensed users by amplifying (relaying) their signals. In particular, an

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overlay user with knowledge of the licensed user’s transmitted signal can use a special type of signal encoding to completely cancel the interference caused by this licensed user at its own receiver. While this coded signal will cause interference to the licensed user, by also using part of its power to amplify the licensed user’s signal, the impact of this interference can be completely mitigated and, in fact, the licensed user might even experience better performance than without the existence of the overlay user. The overlay cognitive radio paradigm was originally proposed in [23] and capacity limits as well as practical implementations of overlay systems have been extensively studied since then. However, regulatory agencies have not yet considered allowing overlay systems to operate in the licensed spectrum.

Overlay, underlay, and interweave radio innovations could make spectrum utilization far more efficient than it is today, in both the licensed and unlicensed bands, thereby enabling new wireless systems, products, and applica tions. More details on the technologies behind these cognitive radio paradigms will be discussed in Chapter 16.7. However, even once these and other cognitive radio technologies are fully developed, changing regulatory policies to include these novel mechanisms will likely be fraught with conflicting opinions from government, industry, and the end users about what best serves the public interest in the allocation of spectrum.

1.3.2 Properties and Existing Allocations

Most wireless communication systems operate in the radio spectrum between 30 MHz and 30 GHz, with some in the millimeter wave frequency band (30 GHz-300 GHz) as well. For communication systems at frequencies below 800 MHz, antenna sizes are too large for small devices, and at frequencies above 3 GHz, signal attenuation with distance precludes long-range transmission. Given these tradeoffs, the primary frequency bands that fueled the extraordinary grown of cellular, Wi-Fi, and short-range radios like Bluetooth were in the 0.8-3 GHz range. As those frequency bands became congested, these systems moved to adjacent bands (0.6-0.8 for cellular and 3-5 GHz for cellular and Wi-Fi). Cellular and Wi-Fi systems started to expand into the millimeter wave bands as well due to the plentiful spectrum there. A similar evolution into higher frequency bands occured in satellite systems. In particular, early satellite systems used the 4-8 GHz C band, but over time moved into the X (8-12.5 GHz), Ku (12-18 GHz), K (18-26 GHz) and Ka (26-40 GHz) bands. The K bands were assigned to terrestrial fixed wireless services as well, including LMDS. Note that the required antenna size for good reception is inversely proportional to the signal frequency, so moving systems to a higher frequency allows for more compact antennas. However, received signal power with nondirectional antennas is proportional to the inverse of frequency squared, so it is harder to cover large distances with high-frequency signals.

As discussed previously, spectrum is allocated either in licensed bands (which regulatory bodies assign to specific operators) or in unlicensed bands (which can be used by any system subject to certain operational re quirements). Often different countries try to match their frequency bands for licensed and unlicensed use so that technology developed for that spectrum is compatible worldwide, however that isn’t possible if a country has al located one of these frequency bands to another use. Figure 1.8 shows the unlicensed spectrum allocations in the United States. In most cases there are similar frequency allocations in Europe and Asia. ISM Band I at 900 MHz has limited spectrum availability as well as licensed users transmitting at high power who interfere with the unlicensed users. Since performance in this band is somewhat poor, it is not heavily utilized by unlicensed devices. The U-NII bands have a total of 555 MHz of spectrum in four separate bands, with different power restrictions. In addition, some countries have imposed restrictions on parts of the U-NII bands so that unlicensed systems must avoid interference with radar systems licensed to operate in those bands.

Figure 1.9 shows the frequency bands allocated to current and legacy cellular systems in the United States along with their band name abbreviations3. Note from this table that 4G (LTE) systems operate in the 3.5 GHz



3Specialized Mobile Radio (SMR); Personal Communication System (PCS); Advanced Wireless Services (AWS); Digital Dividend (DD); Wireless Communications Services (WCS); Broadband Radio Service (BRS); Citizen’s Broadband Radio Service (CBRS).

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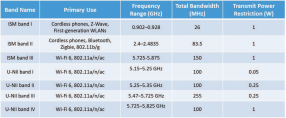


Figure 1.8: Unlicensed Frequency Bands

CBRS band and the 5.2 GHz U-NII band, which are unlicensed bands, and hence must be shared with other unlicensed users. Prior to 4G, cellular systems operated only in dedicated licensed spectrum and hence interference was centrally managed. Under this new paradigm of spectrum sharing, the licensed and unlicensed users using the same band must manage shared access and interference between the different systems.

Figure 1.9: Cellular Frequency Band Allocations in the US and their Band Abbreviations: 2G-5G Systems.

The licensed and unlicensed frequencies allocated to different wireless systems are constantly evolving to meet their capacity demands. For example, once the unlicensed spectrum in the 900 MHz and 2.5 GHz spectrum opened up, many new products and services were launched to exploit this spectrum. This led to the opening of the 3 U-NII bands for unlicensed use. TV broadcasters that were allocated spectrum in the 700 MHz band back in the 20th century were gradually moved out of these bands to open up this spectrum for cellular systems. Cellular systems are now operating in several unlicensed bands and may move into more of them. Millimeter wave spectrum above the 39 GHz Ka band is also being considered for 5G cellular systems.

1.4 Communication Standards

Communication systems that interact with each other require standardization. Standards are typically decided on by national or international committees; in the United States this role is played by the Telecommunications Indus try Association (TIA) while ETSI plays this role in Europe. The IEEE is the major player for WLAN standards development through its 802.11 working group. Other IEEE 802 working groups develop standards for wire less networks, such as short-range or fixed-wireless access networks, to complement WLANs. Cellular system standards are primarily developed by 3GPP. Standards groups typically follow a lengthy process for standards development that entails input from companies and other interested parties as well as a long and detailed review process. The standards process is a large time investment, but companies participate because incorporating their ideas into the standard gives them an advantage in developing the resulting system. In general, standards do not include all the details of the system design, rather only those needed for interoperability. This allows companies to innovate and differentiate their products from other standardized systems. The main goal of standardization is

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enabling systems to interoperate.

In addition to ensuring interoperability, standards also allow economies of scale and pressure prices lower. For example, WLANs typically operate in the unlicensed spectral bands, so they are not required to follow a specific standard. The first generation of WLANs were not standardized and so specialized components were needed for many systems, leading to excessively high cost that, when coupled with poor performance, led to limited adoption. This experience resulted in a strong push to standardize the next WLAN generation, which yielded the highly successful IEEE 802.11 family of standards. Adherence of products to the 802.11 standard is certified by the Wi-Fi Alliance.

There are, of course, disadvantages to standardization. The standards process is not perfect, as company participants often have their own agendas, which do not always coincide with the best technology or the best interests of consumers. In addition, the standards process must be completed at some point, after which it becomes more difficult to add new innovations and improvements to an existing standard. Finally, the standards process can become quite politicized. This happened in Europe with 1st generation cellular systems, where each country had its own standard, and with the second generation of cellular phones in the United States which ultimately adopted three different standards. The formation of the 3GPP standards body to create a single unified cellular standard throughout much of the world was a response to these pitfalls in earlier cellular standards, and is largely responsible for the massive growth in and success of cellular technology. Hence, despite its flaws, standardization is often an essential component of wireless system design and operation in order to ensure its success.

1.5 Wireless Vision

“It is always wise to look ahead, but difficult to look further than you can see.” - Winston Churchill

Wireless communication is ubiquitious in the world we live in, enabling vast connectivity among people and devices as well as rapid access to information. Wireless technology impacts every aspect of modern life: culture, business, politics, economics, health, entertainment, and education. So what might the future bring?

Demand for higher data rates seems unending, hence future wireless systems could support peak speeds of hundreds or perhaps thousands of Gigabits per second. There are billions of people and locations throughout the world today without wireless (or wired) connectivity, so perhaps future systems will fill in these coverage holes so that no person or place on the planet lacks wireless connectivity. Wireless devices might shrink to such small sizes that they can be deployed within any object or living being. Some wireless devices may be powered from a very small battery or even self-power through energy harvesting or wireless charging, eliminating the need to ever plug in. Wireless technology might evolve to support the “five-nines” standard in reliability for tradition telephone service, meaning that the service is reliable 99.999% of the time in any location, indoors and out. Finally, wireless systems must be extremely secure against natural impairments as well as eavesdroppers, attackers and spoofers.

If this vision of wireless technology comes to pass, what will it enable? In addition to providing people with voice, high-speed data, and broadcast entertainment, future wireless networks will also support machine-to machine communications for tens of billions of devices. In the home these networks will enable a new class of intelligent electronic devices that can interact with each other and with the Internet. Such “smart” homes will drastically improve energy efficiency, security, emergency response, as well as help the elderly and disabled with assisted living. Other applications of these networks include sensing and data collection in the power grid to improve robustness and efficiency, “smart cities” that provide services such as trash collection and road mainte nance when the need is detected, and in-body communications for medical devices, biosensors, and targeted drug delivery. Wireless video and virtual reality will permeate the home and any place that people congregate with entertainment, and also enable remote classrooms, remote training facilities, and remote hospitals anywhere in the world. Wireless sensor networks will improve monitoring of and response to fire hazards, toxic waste sites,

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stress and strain in buildings and bridges, carbon dioxide movement, and the spread of chemicals and gases at disaster sites. Finally, wireless networks with very low latency will enable distributed control systems with remote devices, sensors, and actuators linked together via wireless communication channels. Such systems will in turn en able intelligent transportation systems including self-driving vehicles, mobile robots and drones, as well as easily reconfigurable industrial automation.

The exponential growth of smartphone use and wireless Internet access has led to great optimism about wireless technology in general. Obviously not all wireless applications will flourish. While many wireless systems and companies have enjoyed spectacular success, there have also been many failures along the way, including the first generation of wireless LANs and LEO satellite systems, as well as wide area data services, and fixed wireless access to the home. Indeed, it is impossible to predict what wireless failures and triumphs lie on the horizon. Moreover, there must be sufficient flexibility and creativity among both engineers and regulators to allow for accidental successes. It is clear, however, that the current and emerging wireless systems of today – coupled with the vision of applications that wireless can enable – ensure a bright future for wireless technology.

1.6 Technical Challenges

This section provides an overview of the many technical challenges that must be addressed to make the wireless vision a reality. These challenges extend across all aspects of the system, including hardware design, channel characterization, physical layer and multiple access techniques as well as networking protocols and architectures. Techniques to address many of these challenges are described in subsequent chapters of the book.

The design of wireless systems begins with a model for the underlying channel through which the signals will propagate. In all such channels, signal power decreases with distance due to the physics of propagation as well as attenuation from blocking objects. These signal propagation models are developed in Chapter 2. If the transmitter, receiver, or surrounding objects are moving, the channel changes randomly with time due to changing reflections and attenuation. These random channel variations, whose statistical models are developed in Chapter 3, make it difficult to design reliable systems with guaranteed performance. Channel characteristics, including signal attenu ation, also depend on the frequency of operation. In particular, received power generally decreases with the carrier frequency and, in the case of free space propagation with omnidirectional antennas, it is inversely proportional to the square of this frequency. Thus, most wireless systems today operate at carrier frequencies below 5 GHz to ensure good coverage, leading to a spectrum shortage in this range of frequencies. Moving to higher carrier frequencies, such as millimeter wave (30-300 GHz) or terahertz (.3-3 THz), provides much more spectrum than what is available in the lower frequency bands. However, these higher frequencies of operation reduce range unless energy is directionally focused using multiple or directional antenna techniques. Signal propagation characteristics at these frequencies create challenges in designing reliable communication links, as described in Chapter 2.9.5. In addition, hardware components are expensive and power hungry.

The maximum data rate that can be reliably sent over a wireless (or wireline) channel is its Shannon capacity, which is derived in Chapter 4. This capacity is directly proportional to the channel bandwidth, i.e. the amount of spectrum allocated to the channel. This rate also depends on the number of antennas at the transmitter and receiver, as multiple-input multiple-output (MIMO) techniques allow for independent data streams to be transmitted along the independent spatial dimensions these multiple antennas create, as described in Chapter 10. Achievable data rates for a given wireless system also depend on signal propagation and interference characteristics. Due to these challenges, data rates for both cellular and Wi-Fi systems are several orders of magnitude lower than for a fiber optic cable, but that could change as millimeter wave systems are deployed.

In terms of hardware challenges, as the size of wireless devices shrink, breakthroughs are needed to make both the analog and digital circuitry significantly smaller, cheaper, and more energy efficient. Wireless systems operating at millimeter wave and higher frequencies require cheap and reliable RF components, which remains a

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significant challenge. The size, power consumption, and precision of analog-to-digital converters is also becoming a bottleneck as systems move to larger bandwidths. Finally, large antenna arrays that improve signal propagation and mitigate interference require hardware innovations, including hybrid analog and digital processing, to reduce their footprint, power consumption, and cost.

The physical layer design of wireless systems today is quite sophisticated, with dense-constellation modula tion (Chapters 5-6), diversity and adaptive modulation techniques to combat random signal variations (Chapters 7 and 9), powerful error-correction coding (Chapter 8), as well as techniques to combat intersymbol interference caused by delayed signal components that arise from channel reflections (Chapters 11-13). Systems operating in rapidly changing environments require new modulation, coding, detection, and multi-antenna techniques that are robust to such changes when adaptation to them is infeasible. Machine learning may play a role in improving physical layer techniques for channel estimation, signal detection, decoding, and equalizaiton, particularly when channels are hard to model, hard to estimate, or rapidly varying [24, 25]. In addition, algorithms that best exploit the many spatial degrees of freedom offered by large antenna arrays are needed. For energy-constrained systems, particularly those running off non-rechargeable batteries, communication schemes must be developed that can meet performance requirements while minimizing total system power consumption (for signal transmission, reception, and processing).

Multiple access techniques, developed in Chapter 14, allow users to share the same system bandwidth. This bandwidth sharing is done either through coordinated access, as in cellular systems and current Wi-Fi standards, or through distributed techniques, as used in early generations of Wi-Fi systems. Most wireless systems use access schemes that assign orthogonal time and frequency slices of the total system bandwidth to different users. For MIMO systems, independent spatial dimensions can be used as separate channels as well. Non-orthogonal access designs improve spectral efficiency by overlapping channels in the time, frequency, code, or spatial dimensions while managing the interference that results. Channel assignment is done by the base stations in cellular systems and by access points in Wi-Fi systems. Wireless access techniques are ripe for innovation, in the centralized and distributed mechanisms used to assign channels as well as in slicing up channels in a non-orthogonal manner to increases spectral efficiency with minimal interference. Machine learning is also being applied to multiple access, resource allocation, and scheduling [24].

The network architectures and protocols for future wireless systems must support a much wider range of devices and applications than exist today. For infrastructure-based systems like cellular networks (Chapter 15), this will require a more flexible architecture, with a range of cell sizes, greater densification of cells, as well as significant data processing, dynamic optimization, and resource allocation that may be centralized, decentralized, or centralized within a subsets of nodes (coined neighborhood or “fog-based” optimization). Rethinking cellular system architectures to take into account novel forms of cooperation across base stations and users to exploit rather than mitigate interference may lead to significantly higher capacity and robustness. Reduction of latency and over head for applications with short delay-constrained messages is also needed, as are innovations in energy-efficient architectures, e.g. for systems whose backbone is powered by batteries, solar, other forms of energy harvesting, or wireless power transfer. In ad hoc wireless networks (Chapter 16), advances in cooperative techniques for trans mission, reception, and relaying have the potential to significantly improve coverage, reliability, and data rates.

Seamless handoff between different wireless networks, such as Wi-Fi and cellular, is precluded by existing protocols and addressing mechanisms, which must evolve to allow a device to continuously connect on any and all available networks. How best to utilize a multiplicity of available networks for a given device or application is an open challenge. Many wireless systems consist of predominantly wired links, with wireless as the last hop only. Hence, challenges in wired networking, such as latency and bandwidth shortages, impact wireless networks as well. Deployment of caching and computation capabilities in base stations and Wi-Fi access points give rise to new design challenges; should applications utilize these edge capabilities or rely on more powerful capabilities situated farther from the network edge, which entail more latency to access. Security is also more difficult to implement for

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signals traveling over wireless channels, since the airwaves are susceptible to snooping and jamming by anyone with an RF antenna.

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Chapter 1 Problems

1. As storage capability increases, we can store larger and larger amounts of data on smaller and smaller storage devices. Indeed, we can envision microscopic computer chips storing terraflops of data. Suppose this data is to be transfered over some distance. Discuss the pros and cons of putting a large number of these storage devices in a truck or drone and driving or flying them to their destination rather than sending the data electronically.

2. Describe two technical advantages and disadvantages of wireless systems that use bursty data transmission rather than continuous data transmission.

3. Fiber optic cable typically exhibits a probability of bit error of Pb = 10−12. A form of wireless modulation, DPSK, has Pb = 1/2γ in some wireless channels, where γ is the average SNR. Find the average SNR required to achieve the same Pb in the wireless channel as in the fiber optic cable. Because of this extremely high required SNR, wireless channels typically have Pb much larger than 10−12.

4. Find the round-trip delay of data sent between a satellite and the earth for LEO, MEO, and GEO satellites assuming the speed of light is 3 · 108 m/s. If the maximum acceptable delay for a voice system is 30 ms, which of these satellite systems would be acceptable for two-way voice communication?

5. What applications might significantly increase the demand for wireless data?

6. This problem illustrates some of the economic issues facing service providers for mixed-media systems. Suppose you are a service provider with 120 kHz of bandwidth that you must allocate between voice and data users. The voice users require 20 kHz of bandwidth and the data users require 60 kHz of bandwidth. So, for example, you could allocate all of your bandwidth to voice users, resulting in six voice channels, or you could divide the bandwidth into one data channel and three voice channels, etc. Suppose further that this is a time-division system with timeslots of duration T. All voice and data call requests come in at the beginning of a timeslot, and both types of calls last T seconds. There are six independent voice users in the system: each of these users requests a voice channel with probability .2 and pays $.20 if his call is processed. There are two independent data users in the system: each of these users requests a data channel with probability .5 and pays $.50 if his call is processed. How should you allocate your bandwidth to maximize your expected revenue?

7. Describe three disadvantages of using a fixed wireless access systems instead of DSL or cable. Describe three scenarios where the disadvantages override the advantages.

8. Cellular systems have migrated to Hetnets consisting of a mix of macrocells and small cells in order to increase system capacity and energy efficiency. Name at least three design issues that are complicated by this trend.

9. Why does minimizing the reuse distance maximize the spectral efficiency of a cellular system?

10. This problem demonstrates the capacity increase associated with a decrease in cell size. Consider a square city of 100 square kilometers. Suppose you design a cellular system for this city with square cells, where every cell (regardless of cell size) has 100 channels and so can support 100 active users. (In practice, the number of users that can be supported per cell is mostly independent of cell size as long as the propagation model and power scale appropriately.)

(a) What is the total number of active users that your system can support for a cell size of 1 km2? 23

(b) What cell size would you use if your system had to support 250,000 active users?

Now we consider some financial implications based on the fact that users do not talk continuously. Assume that Friday from 5–6 P.M. is the busiest hour for cell-phone users. During this time, the average user places a single call, and this call lasts two minutes. Your system should be designed so that subscribers need tolerate no greater than a 2% blocking probability during this peak hour. (Blocking probability is computed using the Erlang B model: Pb = (AC/C!)/!"ck=0 Ak/k!#, where C is the number of channels and A = UµH for U the number of users, µ the average number of call requests per unit time per user, and H the average duration of a call [5, Chap. 3.6].

(c) How many total subscribers can be supported in the macrocell system (1-km2 cells) and in the microcell system (with cell size from part (b))?

(d) If a base station costs $500,000, what are the base station costs for each system?

(e) If the monthly user fee in each system is $50, what will be the monthly revenue in each case? How long will it take to recoup the infrastructure (base station) cost for each system?

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Chapter 2

Path Loss, Shadowing, and Multipath

The wireless radio channel poses a severe challenge as a medium for reliable high-speed communication. Not only is it susceptible to noise, interference, and other channel impediments, but these impediments change over time in unpredictable ways as a result of user movement and environment dynamics. In this chapter we characterize the primary phenomena that affect signal propagation: path loss and shadowing, as well as signal reflection, diffraction, and scattering. Path loss characterizes how a signal’s received power decreases with transmit-receive distance. It is caused by dissipation of the power radiated by the transmitter as well as by effects of the propagation channel. Path-loss models assume that path loss is the same at a given transmit–receive distance (assuming that the path loss model does not include shadowing effects). Shadowing is the attenuation caused by obstacles between the transmitter and receiver that absorb the transmitted signal. When the attenuation is strong, the signal is blocked. The number and type of objects that cause shadowing at any given receiver location is typically unknown. Hence attenuation due to shadowing is modeled as a random parameter. Unlike path loss, shadowing does not depend on the transmit-receive distance itself but rather on the objects between the transmitter and receiver. Reflection, diffraction, and scattering are caused by a transmitted signal interacting with objects in the environment around the transmitter or receiver. The signal components that arise due to these objects are called multipath components. Different multipath components arrive at the receiver with different time delays and phase shifts. When the phase shifts are aligned, the multipath components add constructively; when they are not aligned, they add destructively. This constructive and destructive addition of multipath components leads to significant variations in the received signal power.

Received power variation due to path loss occurs over long distances (100-1000 m), whereas variation due to shadowing occurs over distances that are proportional to the length of the obstructing object (10-100 m in outdoor environments and less in indoor environments). Since variations in received power due to path loss and shadowing occur over relatively large distances, these variations are sometimes referred to as large-scale propagation effects. The received power variations due to constructive and destructive addition of multipath components occur over very short distances, on the order of the signal wavelength, since each component’s phase rotates 360 degrees over that distance. Hence, power variations due to multipath are sometimes referred to as small-scale propagation effects. Figure 2.1 shows an example of the received-to-transmit power ratio in decibels1 (dB) versus log distance for the combined effects of path loss, shadowing, and multipath. As indicated in the figure, received power due to path loss alone is generally modeled as decreasing linearly with respect to the log of the transmit-receive distance, with additional slow variations due to shadowing and fast variations due to multipath.

After a brief introduction to propagation and a description of our signal model, we present the simplest model for signal propagation: free-space path loss. A signal propagating between two points with no attenuation or reflection follows the free-space propagation law. We then describe the two-ray multipath model, which augments



1The decibel value of x is 10 log10 x.

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Figure 2.1: Effects of path loss, shadowing, and multipath on received power as a function of distance.

the free-space model with a single reflected ray. The two-ray model introduces the notion of a variable path-loss exponent, which gives rise to more general path loss exponent models. After introducing these models, we discuss the commonly-used log-normal model for shadowing. General ray tracing is then introduced to model the multipath components that arise due to signal reflections, diffraction, and scattering. These models approximate signal propagation according to Maxwell’s equations and depend heavily on the geometry and dielectric properties of the region through which the signal propagates. If the number of multipath components is large or if the geometry and dielectric properties of the propagation environment are unknown, then statistical multipath models must be used instead of ray tracing. These statistical multipath models will be described in Chapter 3. We close the chapter by describing empirical channel models with parameters for path loss and shadowing based on measurements for both indoor and outdoor channels.

Although this chapter gives a brief overview of channel models for path loss, shadowing, and multipath, comprehensive coverage of channel and propagation models at different frequencies of interest merits a book in its own right, and in fact there are many excellent references on this topic including [1, 2, 3, 4, 10, 6]. Models specialized to multiple antenna, ultrawideband, and millimeter wave channels can be found in [7], [8], and [9], respectively.

2.1 Radio Wave Propagation

The initial understanding of radio wave propagation goes back to the pioneering work of James Clerk Maxwell, who in 1864 formulated a theory of electromagnetic propagation that predicted the existence of radio waves. In 1887, the physical existence of these waves was demonstrated by Heinrich Hertz. However, Hertz saw no practical use for radio waves, reasoning that since audio frequencies were low, where propagation was poor, radio waves could never carry voice. In 1894 Oliver Lodge used these principles to build the first wireless communication system, though its transmission distance was limited to 150 meters. By 1897 the entrepreneur Guglielmo Marconi had managed to send a radio signal from the Isle of Wight to a tugboat eighteen miles away, and in 1901 Marconi’s wireless system could traverse the Atlantic ocean. These early systems used telegraph signals for communicating information. The first transmission of voice and music was made by Reginald Fessenden in 1906 using a form of amplitude modulation, which circumvented the propagation limitations at low frequencies observed by Hertz by translating signals to a higher frequency, as is done in all wireless systems today.

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Electromagnetic waves propagate through environments where they are reflected, scattered, and diffracted by walls, terrain, buildings, and other objects. The ultimate details of this propagation can be obtained by solving Maxwell’s equations with boundary conditions that express the physical characteristics of these obstructing objects. This often requires the calculation of the radar cross-section (RCS) of large and complex structures. Since these calculations are difficult and since the necessary parameters are often not available, approximations have been developed to characterize signal propagation without resorting to Maxwell’s equations.

The most common signal propagation approximations use ray-tracing techniques based on ray-optic theory [10]. Ray-tracing approximates the propagation of electromagnetic waves by representing the wavefronts as dis crete narrow beams or rays. This approximation determines the reflection and refraction effects on the wavefront but ignores the more complex scattering phenomenon predicted by Maxwell’s coupled differential equations. The ray-tracing model most accurately approximates Maxwell’s equations when the wavelength of the signal is much less than the size of the objects off of which it is reflected, refracted, or scattered. The simplest ray-tracing model is the two-ray model, which accurately describes signal propagation when there is one direct path between the trans mitter and receiver and one reflected path. The reflected path typically bounces off the ground, and the two-ray model is a good approximation for propagation along highways or rural roads and over water. We will analyze the two-ray model in detail, as well as more complex models with additional reflected, scattered, or diffracted compo nents. Many propagation environments are not accurately characterized by ray-tracing models. In these cases it is common to develop analytical models based on empirical measurements, and we will discuss several of the most common of these empirical models.

Often the complexity and variability of the radio channel make it difficult to obtain an accurate deterministic channel model. For these cases, statistical models are often used. The attenuation caused by signal path obstruc tions such as buildings or other objects is typically characterized statistically, as described in Section 2.7. Statistical models are also used to characterize the constructive and destructive interference for a large number of multipath components, as described in Chapter 3. Indoor environments tend to be less regular than outdoor environments, since the geometric and dielectric characteristics change dramatically depending on whether the indoor environ ment is an open factory, cubicled office, or metal machine shop. For these environments computer-aided modeling tools are available to predict signal propagation characteristics [11].

2.2 Transmit and Receive Signal Models

The transmitted and received signals in any wireless system are real-valued. The channel introduces an amplitude and phase change at each frequency of the transmitted signal so that the received signal is also real-valued. Real modulated and demodulated signals are often represented as the real part of a complex signal in order to facilitate analysis. This model gives rise to the equivalent lowpass representation of bandpass signals, which we use for our transmitted and received signals. More details on the equivalent lowpass representation of bandpass signals and systems can be found in Appendix A.

We model the transmitted signal at carrier frequency fc as

s(t) = Re{u(t)ej2πfct}

= Re{u(t)} cos(2πfct) − Im{u(t)} sin(2πfct)

= sI (t) cos(2πfct) − sQ(t) sin(2πfct), (2.1)

where u(t) = sI (t)+jsQ(t) is a complex baseband signal with in-phase component sI (t) = Re{u(t)}, quadrature component sQ(t) = Im{u(t)}, bandwidth Bu, and power Pu. The signal u(t) is called the complex envelope or equivalent lowpass signal of s(t). We call u(t) the complex envelope of s(t) because the magnitude of u(t) is the magnitude of s(t). The phase of u(t) includes any carrier phase offset. The equivalent lowpass representation of

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band-pass signals with bandwidth B ≪ fc allows signal manipulation via u(t) irrespective of the carrier frequency. The power in the transmitted signal s(t) is Pt = Pu/2.

For time-invariant channels, the received signal is the convolution of s(t) with the channel impulse response h(t) plus an additional noise component n(t) introduced by the channel: r(t) = s(t) ∗ h(t) + n(t). It can be written in a similar form as the transmitted signal as:

r(t) = Re{v(t)ej2πfct} + n(t), (2.2)

where v(t) is the equivalent lowpass signal for the received signal without the noise. This equivalent signal depends on h(t). In particular, as discussed in Appendix A, v(t) = u(t) ∗ c(t), where c(t) is the equivalent lowpass channel impulse response for h(t). Time-varying channels will be treated in Chapter 3.

The received signal in (2.2) consists of two terms, the first term corresponding to the transmitted signal after propagation through the channel, and the second term corresponding to the noise added by the channel. The signal to-noise power ratio (SNR) of the received signal is defined as the power of the first term divided by the power of the second term. In this chapter (and in Chapter 3) we will neglect the random noise component n(t) in our analysis, since these chapters focus on signal propagation, which is not affected by noise. However, noise will play a prominent role in the capacity and performance of wireless systems studied in later chapters.

When the transmitter or receiver is moving, the received signal will have a Doppler shift of fD = vλ cos θ associated with it, where θ is the arrival angle of the received signal relative to the direction of motion, v is the receiver velocity toward the transmitter in the direction of motion, and λ = c/fc is the signal wavelength (c = 3 · 108 m/s is the speed of light). The geometry associated with the Doppler shift is shown in Figure 2.2. The Doppler shift results from the fact that transmitter or receiver movement over a short time interval ∆t causes a slight change in distance ∆d = v∆t cos θ that the transmitted signal needs to travel to the receiver. The phase change due to this path-length difference is ∆φ = 2πv

relationship between signal frequency and phase:

λ ∆t cos θ. The Doppler frequency is then obtained from the

fD = 12π∆φ 

∆t = vλ cos θ. (2.3)

If the receiver is moving toward the transmitter (i.e., if −π/2 ≤ θ ≤ π/2) , then the Doppler frequency is positive; otherwise, it is negative. We will ignore the Doppler term in the path loss models of this chapter, since for typical vehicle speeds (75 km/hr) and frequencies (about 10 GHz) it is small, on the order of 1 KHz [1]. However, we will include Doppler effects in Chapter 3 on statistical fading models where it is used to characterize a random channel’s rate of change.

Suppose a signal s(t) of power Pt is transmitted through a given channel with corresponding received signal r(t) of power Pr, where Pr is averaged over any random variations due to shadowing. We define the linear path loss of the channel as the ratio of transmit power to receive power:

PL = Pt 

Pr. (2.4)

We define the path loss of the channel as the value of the linear path loss in decibels or, equivalently, the difference in dB between the transmitted and received signal power:

PL dB = 10 log10

$Pt Pr

%

dB. (2.5)

The dB path loss is a nonnegative number since the channel does not contain active elements, and thus it can only attenuate the signal. The dB path gain is defined as the negative of the dB path loss: PG = −PL = 10 log10(Pr/Pt) dB. Due to the laws of physics underlying signal propagation, the dB path loss generally increases

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Figure 2.2: Geometry associated with Doppler shift.

with distance, as illustrated in Figure 2.1 (note that this figure shows a decreasing path gain or, equivalently, an increasing path loss). The path loss generally increases with distance, which means the received power in dB decreases with distance, as illustrated in Figure 2.1. With shadowing, also illusrated in Figure 2.1, the received power is random owing to random blockage from objects, as we discuss in Section 2.6.

2.3 Free-Space Path Loss

Consider a signal transmitted through free space to a receiver located at distance d from the transmitter. Assume there are no obstructions between the transmitter and receiver and that the signal propagates along a straight line between the two. The channel model associated with this transmission is called a line-of-sight (LOS) channel, and the corresponding received signal is called the LOS signal or ray. Under free-space path loss the received signal is given by [2]:

&'

r(t) = Re

λ√GtGru(t − τl)e−j2πd/λ 4πd

(

ej2πfct

)

, (2.6)

where Gt and Gr are, respectively, the transmit and receive antenna power gains in the LOS direction relative to a unity gain isotropic antenna, τl = d/c is the signal propagation delay of the LOS signal traveling the distance d, and the phase shift e−j2πd/λ is due to the distance d that the wave travels2. Transmit and receive directional antennas have gains Gt and Gr greater than unity in one or more angular directions relative to the idealized isotropic antenna gains Gt = Gr = 1. Directional antenna gains can range from 2.15 dB for a half-wavelength dipole to tens of dB in horn or dish antennas. More details on directional antenna designs and their gains can be found in [12].



2When the transmit and receive antennas are at the same height, the distance d equals the horizontal separation distance between the transmitter and receiver. When the transmitter and receiver are at different heights, the distance the wave travels exceeds this separation distance. These different distances will be characterized in the two-ray model (Section 2.4).

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The power in the transmitted signal s(t) is Pt, so from (2.6), the linear path loss for free-space propagation, a formula first introduced by Friis [13], is

Pr

Pt= GtGr

\* λ

4πd

+2

. (2.7)

The Friis formula (2.7) dictates that the receive power Pr equals the transmit power per unit area under free space propagation that is incident on the receive antenna, given by PtGt/(4πd2), multiplied by that antenna’s effective area Ar, a quantity that determines how effectively the receive antenna captures this incident power. From (2.7) this effective area equals

Ar = λ2Gr 

4π . (2.8)

The dependence of Ar on λ2 is due to the resonance of a signal with a given wavelength on a linear conductor half that size. Since λ = c/fc, the Friis formulae indicates that, as the carrier frequency increases, the received power decreases as the square of the increase. Antenna arrays, discussed in Chapter 10, create a different effective area than single-element antennas, which can mitigate or remove the dependence of received power on λ [14]. The Friis formula also indicates that the received signal power falls off in inverse proportion to the square of the distance d between the transmit and receive antennas. We will see in the next section that, for other signal propagation environments, the received signal power falls off more quickly relative to this distance.

The received power can be expressed in dBm as3

Pr (dBm) = Pt (dBm) + 10 log10(GtGr) + 20 log10(λ) − 20 log10(4π) − 20 log10(d). (2.9)

Equation (2.9) is an example of a link budget equation, which expresses the received power of a signal transmitted through a given channel (or link) as a function of the transmit power and all the losses the signal experiences on that link. Free-space path loss is defined as the path loss of the free-space model:

PL (dB) = 10 log10

The free-space path gain is thus

$Pt Pr

%

= −10 log10

,

GtGr

\* λ

4πd

+2-

. (2.10)

PG = −PL = 10 log10

,

GtGr

\* λ

4πd

+2-

. (2.11)

Example 2.1: Consider an outdoor small cell with fc = 2.5 GHz, cells of radius 10 m, and isotropic antennas. Under the free-space path loss model, what transmit power is required at the base station in order for all terminals within the cell to receive a minimum power of 0.1 µW? How does this change if the system frequency is 5 GHz?

Solution: We must find a transmit power such that the terminals at the cell boundary receive the minimum required power. We obtain a formula for the required transmit power by inverting (2.7) to obtain:

Pt = Pr



\* 4πd √GtGrλ

+2

.

3The dBm value of a power x is its dB value relative to a milliwatt: 10 log10(x/.001). The dBW value of a power x is its dB value relative to a watt: 10 log10 x, so 1W corresponds to 30 dBm.

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Substituting in Gt = Gr = 1 (isotropic antennas), λ = c/fc = 0.12 m, d = 10 m, and Pr = 0.1 µW yields Pt = .1097 W. At 5 GHz only λ = .06 changes, so Pt = .4388 W. We see that doubling the carrier frequency leads to a requirement for four times more transmit power, illustratingthe power consumption challenges in moving wireless systems to high-frequency spectral bands.



2.4 Two-Ray Multipath Model

The two-ray model is used when a single ground (or other) reflection dominates the multipath effect, as illustrated in Figure 2.3. The received signal consists of two components: the LOS component or ray, which corresponds to the transmitted signal propagating through free space, and a reflected component or ray, which is the transmitted signal reflected off the ground. The received LOS ray is given by the free-space propagation loss formula (2.6) with the distance the signal travels set to d0. The reflected ray shown in Figure 2.3 travels distance d1 = d11 +d12. Since the two signal components in this model travel different distances, we use d in this model and all subsequent propagation models in this chapter to denote the horizontal distance between the transmitter and receiver. The distances d0 and d1 then depend on d as well as the transmitter and receiver heights ht and hr. We see from Figure 2.3 that when ht = hr, d = d0, and when d is very large relative to ht and hr, d ≈ d0 ≈ d1.

If we ignore the effect of surface wave attenuation4 then, by superposition, the received signal for the two-ray model is

r2-ray(t) = Re

&

λ

4π

'√G0u(t − τ0)e−j2πd0/λ 

d0+R√G1u(t − τ1)e−j2πd1/λ d1

(

ej2πfct

)

, (2.12)

where τ0 = d0/c is the time delay of the LOS ray, τ1 = d1/c is the time delay of the ground reflection ray, √G0 = .Gt0Gr0 is the product of the transmit and receive antenna field radiation patterns in the LOS direction,

R is the ground reflection coefficient, and √G1 = .Gt1Gr1 is the product of the transmit and receive antenna field radiation patterns corresponding to the ground reflection. The delay spread of the two-ray model equals the difference between the delay of the LOS ray and that of the reflected ray: (d1 − d0)/c.

If the transmitted signal is narrowband relative to the delay spread such that (τ1−τ0 ≪ B−1 u ) then u(t−τ1) ≈ u(t − τ0). With this approximation, the received power of the two-ray model for narrowband transmission is



Figure 2.3: Two-ray model.



4This is a valid approximation for antennas located more than a few wavelengths from the ground.

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Pr = Pt

\* λ 4π

+2 ////√G0 

d0+R√G1e−j∆φ d1

////2, (2.13)

where ∆φ = 2π(d1 − d0)/λ is the phase difference between the two received signal components. Equation (2.13) has been shown [15] to agree closely with empirical da

.ta. From the geometry of the two-ray model, d1 = (ht + hr)2 + d2 and d0 = .(ht − hr)2 + d2. Hence 

d1 − d0 = .(ht + hr)2 + d2 − .(ht − hr)2 + d2. (2.14)

When d is very large compared to ht + hr , we can use a Taylor series approximation in (2.14) to get ∆φ = 2πλ0.(ht + hr)2 + d2 − .(ht − hr )2 + d21≈ 4πhthr 

λd . (2.15)

The ground reflection coefficient is given by [1, 16]

R = sin θ − Z 

sin θ + Z , (2.16)

where

&√εr − cos2 θ/εr for vertical polarization,

Z =

√εr − cos2 θ for horizontal polarization, (2.17)

and εr is the dielectric constant of the ground. For earth or road surfaces this dielectric constant is approximately that of a pure dielectric (for which εr is real with a value of about 15).

Since for asymptotically large d, d1 ≈ d0 ≈ d, G1 ≈ G0 where we denote this approximate antenna gain as G, and θ ≈ 0 which by (2.16) then implies that R ≈ −1. Substituting these approximations into (2.13) yields

'

Pr ≈ Pt

λ√G 4πd

(2

|1 − e−j∆φ|2. (2.18)

Using the approximation for ∆φ in (2.15) at large d we get the final approximation for received power at large distance d as

(2

, (2.19)

or, in dB,

'

Pr ≈ Pt

λ√G 4πd

(2\*4πhthr λd

+2

= Pt

'√Ghthr d2

Pr dBm = Pt dBm + 10 log10(G) + 20 log10(hthr) − 40 log10(d). (2.20)

Thus, in the limit of asymptotically large d, the received power falls off inversely with the fourth power of d and is independent of the wavelength λ. This is in contrast to free-space propagation where power is inversely proportion to d2 and proportional to λ2. The received signal power becomes independent of λ because directional antenna arrays have a received power that does not necessarily decrease with λ, and combining the direct path and reflected signal at the receiver effectively forms an antenna array.

A plot of (2.13) as a function of the transmit-receive separation distance d is shown in Figure 2.4 for fc = 900 MHz, R = −1, ht = 50 m, hr = 2 m, G0 = G1 = 1, and received power normalized to 0 dB at a reference distance of 1 m. This plot reflects the case where the transmit antenna height ht is much larger than the receiver

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Figure 2.4: Received power versus distance for two-ray model. The received power is normalized to 0 dB at a reference distance dr = 1 m (so 10 log(d/dr)=0 dB for d = 1 m).

antenna height hr, as is typical in an urban cellular system where macrocell base stations are located on top of tall buildings with mobile devices near street level. The plot can be separated into three segments. For small distances (d<ht) the two rays add constructively as d increases from zero. The local maximum over this segment occurs for d such that ∆φ = π in (2.15). At that distance the phases of the two multipath components are perfectly aligned (Re−j∆φ = 1). With this phase alignment the path loss over this segment is roughly proportional to 1/(d2 + h2t)

since both multipath components travel roughly the same distance .d2 + h2t . That is because, for hr << ht, d1 = .d2 + (ht + hr )2 ≈ d0 = .d2 + (ht − hr)2 ≈2d2 + h2t . (2.21)

After this first local maximum and up to a certain critical distance dc, the wave experiences constructive and destructive interference of the two rays as ∆φ changes with d, resulting in a wave pattern with a sequence of maxima (when |∆φ| is an odd integer multiple of π) and minima (when |∆φ| is an even integer multiple of π). These maxima and minima are also referred to as multipath fading, discussed in more detail in the next chapter. At the critical distance dc, the final maximum is reached, after which the signal power falls off proportionally with d−4. This rapid falloff with distance is due to the fact that, for d>dc, the signal components only combine destructively since ∆φ → 0 as d → ∞. Since the final maximum is achieved for d such that ∆φ = π, an approximation for dc can be obtained by setting ∆φ = π in (2.15), obtaining

dc = 4hthr/λ. (2.22)

For the parameters used to generate Figure 2.4, dc = 1.2 Km (log10 dc = 3.08), which is also shown in the figure. The power falloff with distance in the two-ray model can be approximated by averaging out its local maxima and minima. This results in a piecewise linear model with three segments, which is also shown in Figure 2.4 slightly offset from the actual power falloff curve for illustration purposes. In the first segment, in order to get a

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constant power falloff we assume d0 ≈ d1 ≈ ht, leading to a power falloff proportional to 1/h2t that is independent of d; for distances between ht and dc, power falls off at −20 dB/decade; and at distances greater than dc, power falls off at –40 dB/decade.

The critical distance dc is proportional to the transmit antenna height ht and to the carrier frequency fc = c/λ. Thus, for small cells and indoor systems where ht is ten meters or less, for fc on the order of 1 GHz, dc is hundreds of meters, as illustrated in Example 2.2. On the other hand, for millimeter wave systems where fc is tens of GHz, dc typically well exceeds a kilometer. The value of dc can be used for system design. For example, if propagation in a cellular system obeys the two-ray model then the critical distance would be a natural size for the cell radius, since the path loss associated with interference outside the cell would be much larger than path loss for desired signals inside the cell. However, setting the cell radius to dc could result in very large cells, as illustrated in Figure 2.4 and in the next example. Since smaller cells are more desirable – both to increase capacity and reduce transmit power – cell radii are typically much smaller than dc. Thus, with a two-ray propagation model, power falloff within these relatively small cells goes as distance squared. Moreover, propagation in cellular systems rarely follows a two-ray model, since cancellation by reflected rays rarely occurs in all directions.

Example 2.2: Determine the critical distance for the two-ray model in an outdoor cell with ht = 10 m, hr =3m and in an indoor cell with ht =3m, hr =2m for fc = 2 GHz.

Solution: dc = 4hthr/λ = 800 m for the outdoor cell and 160 m for the indoor cell. A cell radius of 800 m in an outdoor cell is typical for a macrocell but a bit large for a small cell which today are on the order of 10-100 m to maintain large capacity. However, for a macrocell with an 800 m radius under these system parameters, signal power would fall off as d2 inside the cell, while interference from neighboring cells would fall off as d4 and thus would be greatly reduced. Similarly, 160 m is quite large for the cell radius of an indoor system, as there would typically be many walls the signal would have to penetrate for an indoor cell radius of that size. Hence an indoor system would typically have a smaller cell radius: on the order of 10–20 m, the size of one room or a few adjacent rooms in a typical building.



2.5 Path Loss Exponent Models

2.5.1 Single-Slope

The complexity of signal propagation makes it difficult to obtain a single model that characterizes path loss accu rately across a range of different environments and frequencies. Accurate path loss models can be obtained from complex analytical models or empirical measurements when tight system specifications must be met or the best locations for base stations or access-point layouts must be determined. However, for general trade-off analysis of various system designs it is sometimes best to use a simple model that captures the essence of signal propagation without resorting to complicated path loss models, which are only approximations to the real channel anyway. The following single-slope model for path loss as a function of distance is a simple model that captures several important propagation characteristics.

Pr = PtK 36

\*dr d

+γ

, (2.23)

Table 2.1: Typical path loss exponents

Environment γ range 

Urban macrocells 3.7–6.5 

Urban microcells 2.7–3.5 

Office building (same floor) 1.6–3.5 

Office building (multiple floors) 2–6 

Store 1.8–2.2 

Factory 1.6–3.3 

Home 3 



where dr is a reference distance for the antenna far field, γ is the path-loss exponent, and K is a unitless constant equal to the path gain Pr/Pt at distance d = dr. The dB attenuation is thus

Pr (dBm) = Pt (dBm) + K dB − 10γ log10 (d/dr), (2.24)

and the path loss consists of two terms, where all environment and antenna parameters are captured by the first term, and the path loss due only to distance is captured by the second term:

PL (dB) = 10 log10 (Pt/Pr) = −10 log10 K + 10γ log10(d/dr). (2.25)

The values for K, dr, and γ can be obtained to approximate either an analytical or empirical model. In particular, the free-space path-loss model and some of the empirical models described in Section 2.7 are all of the same form as (2.24). Because of antenna near field effects at transmit distances on the order of a signal wavelength, the model (2.24) is generally valid only at transmission distances d>dr, where dr is typically assumed to be 1 m for systems transmitting over distances greater than this nominal value.

When the single-slope model is used to approximate empirical measurements, the value of K < 1 is some times set to the free-space path gain at distance dr assuming isotropic antennas:

K dB = 20 log10λ 

4πdr, (2.26)

and this assumption is supported by empirical data for free-space path loss at a transmission distance of 100 m [17]. Alternatively, K can be determined by measurement at dr or optimized (alone or together with γ) to minimize the mean-square error (MSE) between the model and the empirical measurements [17]. The value of γ depends on the propagation environment: for propagation that approximately follows a free-space or two-ray model, γ is set to 2 or 4 respectively. The value of γ for more complex environments can be obtained via a minimum mean-square error (MMSE) fit to empirical measurements, as illustrated in Example 2.3. Alternatively, γ can be obtained from an empirically based model that takes into account frequency and antenna height [17]. Table 2.6 summarizes values of the path loss exponent γ for different environments based on data from [11, 17, 18, 19, 20, 21, 52]. Empirical measurements indicate that path loss exponents at higher frequencies tend to be higher [19, 23, 21, 22, 57, 59] whereas path loss exponents at higher antenna heights tend to be lower [17, 57, 59]. The more complex empirical models described below in Section 2.7, particularly the widely used 3GPP and WINNER II channel models [57, 59], have additional terms that explicitly capture the dependence of path loss on frequency and antenna height. Note that the wide range of empirical path-loss exponents for indoor propagation may be due to attenuation caused by floors, objects, and partitions (see Section 2.5.5).

Example 2.3: Consider the set of empirical measurements of Pr/Pt given in Table 2.2 for an indoor system at 900

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MHz. Find the path loss exponent γ that minimizes the MSE between the single-slope model (2.24) and the empir ical dB power measurements, assuming that dr =1m and K is determined from the free-space path-gain formula at this dr. Find the received power at 150 m for the single-slope path loss model with this path loss exponent and a transmit power of 1 mW (0 dBm). Note that, since the path loss model in dB is linear, minimizing the MSE of the empirical data in dB (versus in linear units) relative to this model is a simple linear regression.

Solution: We first set up the MMSE error equation for the dB power measurements as

F(γ) = 35 i=1

[Mmeasured(di) − Mmodel(di)]2,

where Mmeasured(di) is the path loss measurement in Table 2.2 at distance di and where Mmodel(di) = K − 10γ log10(di) is the path loss at di based on (2.24) since dr = 1. Now using the free-space path gain formula yields K = 20 log10(.3333/(4π)) = −31.53 dB. Thus

Table 2.2: path loss measurements

Distance from transmitter M = Pr/Pt 

10 m –70 dB 

20 m –75 dB 

50 m –90 dB 

100 m –110 dB 

300 m –125 dB 



F(γ)=(−70 + 31.53 + 10γ)2 + (−75 + 31.53 + 13.01γ)2

+ (−90 + 31.53 + 16.99γ)2 + (−110 + 31.53 + 20γ)2

+ (−125 + 31.53 + 24.77γ)2

= 21682.50 − 11656.60γ + 1571.47γ2. (2.27)

Differentiating F(γ) relative to γ and setting it to zero yields

∂F(γ)

∂γ = −11656.60 + 3142.94γ = 0 ⇒ γ = 3.71.

For the received power at 150 m under the single-slope path loss model with K = −31.53, γ = 3.71, and Pt = 0 dBm, we have Pr = Pt + K − 10γ log10(d/dr)=0 − 31.53 − 10 · 3.71 log10(150) = −112.26 dBm. Clearly the measurements deviate from the single-slope path loss model; this variation can be attributed to shadow fading, described in Section 2.7.



2.5.2 Multi-Slope

The multi-slope path loss model, which is piecewise linear in dB, generalizes the single-slope model by allowing for different path loss exponents at different distances. Given this greater flexibility, the multi-slope model is commonly used to analytically model empirical measurements. The multi-slope path loss model and a set of

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measurement data on which it would be based is illustrated in Figure 2.5 for dB attenuation versus log distance, where the dots represent hypothetical measurements and the multi-slope model is the analytical approximation to these measurements. A multi-slope model with N segments must specify N − 1 breakpoints d1,...,dN−1 as well as the slopes corresponding to each segment s1,...,sN . The slope of the ith segment is si = −10γi for γi the path loss exponent on that segment. Different methods can be used to determine the number and location of breakpoints to be used in modeling the empirical data. Once these are fixed, the slopes corresponding to each segment can be obtained by linear regression on the data. The transition to a new slope might be instantaneous at a given breakpoint distance or entail some smoothing. Multi-slope models capture the analytical two-ray propagation model of Section 2.4. They also fit well to empirical measurements in propagation environments where the path loss exponent increases with distance [52], where the LOS path loss differs from that of non-LOS path loss [98], and in heterogeneous cellular networks where users may connect to a relatively distant macrocell base station while experiencing interference from a nearby small cell [99].

A special case of the multi-slope model is the dual-slope model. The dual-slope model is characterized by a constant path loss factor K and a path loss exponent γ1 above some reference distance dr and up to some breakpoint distance dBP, after which power falls off with path loss exponent γ2. The dB path loss is thus given by:

&

Pr(d) (dB) =

Pt + K − 10γ1 log10(d/dr), dr ≤ d ≤ dBP, Pt + K − 10γ1 log10(dBP/dr) − 10γ2 log10(d/dBP), d>dBP. (2.28)

The path loss exponents, K, and dBP are typically obtained via a regression fit to empirical data [17, 92]. The two ray model described in Section 2.4 for d>ht can be approximated by the dual-slope model, with the breakpoint distance dBP = dc given by (2.22) and path loss exponents γ1 = 2 and γ2 = 4. Many of the measurement-based models described in Section 2.9 use the dual-slope model for path loss.

The transition between the multiple equations in the dual-slope model (2.28) can be smoothed by the following model [34, 93]:

Pr = PtK 

L(d), (2.29)

where

L(d) ∆=

$ d dr

%γ1,

1 +

$ d dBP

%(γ1−γ2)q-1/q

. (2.30)

In this expression, q is a parameter that determines the smoothness of the path loss at the transition region close to the breakpoint distance dBP. This smoothing in this model can be extended to more than two regions [91].

2.6 Shadowing

A signal transmitted through a wireless channel will typically experience random variation due to blockage from objects in the signal path, giving rise to random variations of the received power at a given distance. Such variations are also caused by changes in reflecting surfaces and scattering objects. Thus, a model for the random attenuation due to these effects is also needed. The location, size, and dielectric properties of the blocking objects – as well as the changes in reflecting surfaces and scattering objects that cause the random attenuation – are generally unknown, so statistical models must be used to characterize this attenuation.

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Figure 2.5: Multi-slope model for path loss.

The most common model for this additional attenuation is log-normal shadowing; under this model the dB value of the random attenuation is Gauss-distributed. The log-normal shadowing model has been empirically con firmed to model accurately the variation in received power in both outdoor and indoor radio propagation environ ments (see, e.g., [17, 102]). In the log-normal shadowing model, the ratio of transmit-to-receive power ψ = Pt/Pr is assumed to be random with a log-normal distribution given by

'

p(ψ) = ξ 

√2πσψdB ψ exp

−(10 log10 ψ − µψdB )2 2σ2ψdB 

(

, ψ > 0, (2.31)

where ξ = 10/ ln 10, µψdB is the mean of ψdB = 10 log10 ψ in decibels, and σψdB is the standard deviation of ψdB (also in dB). The mean can be based on an analytical model or empirical measurements. For empirical mea surements µψdB equals the empirical path loss, since average attenuation from shadowing is already incorporated into the measurements. For analytical models, µψdB must incorporate both the path loss (e.g., from a free-space or ray-tracing model) as well as average attenuation from blockage. Alternatively, path loss can be treated separately from shadowing, as described in the next section.

A random variable with a log-normal distribution is called a log-normal random variable. Note that if ψ is log-normal then the received power and received SNR will also be log-normal, since these are just constant multiples of ψ. For received SNR the mean and standard deviation of this log-normal random variable are also in decibels. For log-normal received power the random variable has units of power, so its mean and standard deviation will be in dBm or dBW instead of dB. The mean of ψ (the linear average path gain) can be obtained from (2.31) as

'

µψ = E[ψ] = exp

ξ + σ2ψdB 

µψdB

2ξ2

(

. (2.32)

The conversion from the linear mean (in dB) to the log mean (in dB) is derived from (2.32) as 10 log10 µψ = µψdB + σ2ψdB 

2ξ . (2.33)

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Performance in log-normal shadowing is typically parameterized by the log mean µψdB , which is referred to as the average dB path loss and is given in units of dB. We can compute the distribution of the dB value of ψ, ψdB = 10 log10 ψ from the distribution of ψ given in (2.31). This computation reveals that ψdB is Gauss-distributed with mean µψdB and standard deviation σψdB:

'

p(ψdB) = 1 

√2πσψdBexp

−(ψdB − µψdB )2 2σ2ψdB 

(

. (2.34)

The log-normal distribution of ψ or, equivalently, the Gaussian distribution of ψdB is defined by two parameters: µψdB and σψdB . Given that ψdB is based on the ratio of transmit and receive powers, it can be expressed as a difference of these powers in either dBW or dBm, as follows:

ψdB = 10 log10(Pt/Pr) = Pt dBW − Pr dBW = Pt dBm − Pr dBm. (2.35)

Since ψdB is Gauss-distributed, by (2.35) and properties of Gaussian random variables we have that Pr dBW is also Gauss-distributed with mean Pt dBW − µψdB and standard deviation σψdB, i.e. the same standard deviation as ψdB. Similarly, Pr dBm is Gauss-distributed with mean Pt dBm − µψdB and standard deviation σψdB.

The log-normal distributionis an imperfect approximation to the effect of shadowing because it has a non-zero probability that the received power exceeds the transmit power, which violates the laws of physics. In particular, since the transmit power should always exceed the received power, ψ = Pt/Pr should always be greater than unity. Thus shadowing models set µψdB to be greater than zero. Note, however, that the log-normal distribution (2.31) takes values for 0 ≤ ψ ≤ ∞. Hence, for ψ < 1 we have Pr > Pt, which is physically impossible. However, this probability will be very small when µψdB is large and positive. Thus, the log-normal model captures the underlying physical model for shadowing most accurately when µψdB ≫ 0.

If the mean and standard deviation for the shadowing model are based on empirical measurements, then the question arises as to whether they should be obtained by taking averages of the linear or rather the dB values of the empirical measurements. In other words, given empirical (linear) path loss measurements {pi}Ni=1, should the mean

path loss be determined as µψ = (1/N)"Ni=1 pi or as µψdB = (1/N)"Ni=1 10 log10 pi? A similar question arises for computing the empirical variance. In practice it is more common to determine mean path loss and variance based on averaging the dB values of the empirical measurements for several reasons. First, as we shall see, the mathematical justification for the log-normal model is based on dB measurements. In addition, the literature shows that obtaining empirical averages based on dB path loss measurements leads to a better approximation for the physical channel being modeled [103]. Finally, as we saw in Section 2.5.4, power falloff with distance models are often obtained by a piecewise linear approximation to empirical measurements of dB power versus the log of distance [11].

Most empirical studies for outdoor channels support a standard deviation σψdB ranging from 4 dB to 13 dB [1, 34, 104, 105, 106]. Larger deviations are associated with environments that contain a high density of blocking objects such as buildings and foliage outdoors, or walls and furniture indoors. Moreover, since attenuation by ob jects is more severe and more variable at high frequencies, and there is more variance in the number of attenuating objects at large distances, σψdB generally increases with both frequency and distance [72].The mean power µψdB depends on the path loss and building properties in the area under consideration. The mean power µψdB varies with distance; this is due to path loss and to the fact that average attenuation from objects increases with distance owing to the potential for a larger number of attenuating objects.

The Gaussian model for the distribution of the mean received signal in dB can be justified by the following attenuation model when shadowing is dominated by the attenuation from blocking objects. The attenuation of a signal as it travels through an object of depth d is approximately equal to

s(d) = ce−αd, (2.36)

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where α is an attenuation constant that depends on the object’s interior materials and associated dielectric properties while c is a fixed attenuation constant that doesn’t depend on the object’s depth (e.g. the penetration loss the signal experiences in entering and exiting the object). If we assume that α is approximately equal for all blocking objects and that the ith blocking object has a random depth di, then the attenuation of a signal as it propagates through this region is

s(dt) = ce−αΣidi = ce−αdt , (2.37)

where dt = "i di is the sum of the random object depths through which the signal travels. If there are many objects between the transmitter and receiver, then by the central limit theorem we can approximate dt by a Gaussian random variable. Thus, ln s(dt) = −αdt will have a Gaussian distribution with mean µ and standard deviation σ. The value of σ will depend on the environment.

Example 2.4: In Example 2.3 we found that the exponent for the single-slope path loss model that best fits the measurements in Table 2.2 was γ = 3.71. Assuming the single-slope path loss model with this exponent, dr = 1m, and the same K = −31.53 dB, find σ2ψdB , the variance of log-normal shadowing about the mean path loss based on these empirical measurements.

Solution:

The sample variance relative to the simplified path loss model with γ = 3.71 is

σ2ψdB = 1535 i=1

[Mmeasured(di) − Mmodel(di)]2,

where Mmeasured(di)is the path loss measurement in Table 2.2 at distance di and Mmodel(di) = K−37.1 log10(di)since dr = 1m. This yields

σ2ψdB = 15[(−70 + 31.53 + 37.1)2 + (−75 + 31.53 + 48.27)2

+ (−90 + 31.53 + 63.03)2 + (−110 + 31.53 + 74.2)2

+ (−125 + 31.53 + 91.90)2]

= 13.28.

Thus, the standard deviation of shadow fading on this path is σψdB = 3.64 dB. Note that the bracketed term in the displayed expression equals the MMSE formula (2.27) from Example 2.3 with γ = 3.71.

Extensive measurements have been taken to characterize the empirical autocorrelation function of the shadow fading process over distance for different environments at different frequencies (see e.g. [105, 107, 108, 109, 110]). The most common analytical model for this function, first proposed by Gudmundson [105] and based on empirical measurements, assumes that the shadowing ψ(d) is a first-order autoregressive process where the covariance between shadow fading at two points separated by distance δ is characterized by

A(δ) = E[(ψdB(d) − µψdB )(ψdB(d + δ) − µψdB )] = σ2ψdB ρδ/D

D , (2.38)

where ρD is the normalized covariance between two points separated by a fixed distance D. This covariance must be obtained empirically, and it varies with the propagation environment and carrier frequency. Measurements indicate that for suburban macrocells with fc = 900 MHz, ρD = 0.82 for D = 100 m and for urban microcells

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with fc = 2 GHz, ρD = 0.3 for D = 10 m [105, 109]. This model can be simplified and its empirical dependence removed by setting ρD = 1/e for distance D = Xc, which yields

A(δ) = σ2ψdB e−δ/Xc . (2.39)

The decorrelation distance Xc in this model is the distance at which the signal autocovariance equals 1/e of its maximum value and is on the order of the size of the blocking objects or clusters of these objects. For outdoor systems, Xc typically ranges from 50 m to 100 m [108, 109]. For users moving at velocity v, the shadowing decor relation in time τ is obtained by substituting vτ = δ in (2.38) or (2.39). For a linear array of antennas, discussed in Chapter 10, shadowing can vary with the angle between the array and the signal’s LOS path. Autocorrelation relative to this angle has been investigated in [107, 109].

The first-order autoregressive correlation model (2.38) and its simplified form (2.39) are easy to analyze and to simulate. Specifically, one can simulate ψdB by first generating a white Gaussian noise process with power σ2ψdB and then passing it through a first-order filter with response ρδ/D

D for a covariance characterized by (2.38)

or response e−δ/Xc for a covariance characterized by (2.39). The filter output will produce a shadowing random process with the desired correlation properties [105, 106].

2.7 Combined Path Loss and Shadowing

Models for path loss and shadowing can be combined to capture power falloff versus distance along with the random attenuation about this path loss from shadowing. In this combined model, average dB path loss (µψdB ) is characterized by the path loss model while shadow fading, with a mean of 0 dB, creates variations about this path loss, as illustrated by the path loss and shadowing curve in Figure 2.1. Specifically, this curve plots the combination of the single-slope path loss model (2.23) and the log-normal shadowing random process defined by (2.34) and (2.39).

2.7.1 Single-Slope Path Loss with Log-Normal Shadowing

For the combined model of single-slope path loss (2.24) and log-normal shadowing (2.33), the ratio of received to transmitted power in dB is given by

Pr

PtdB = 10 log10 K − 10γ log10 (d/dr) − ψdB, (2.40)

where ψdB is a Gauss-distributed random variable with variance σ2ψdB . The mean of ψdB is assumed to be zero when the term 10 log10 K captures average shadowing. When this is not the case, for example when K is calculated from the free-space path gain formula at d = dr, then the mean of ψdB is positive and equal to the average shadowing loss over all distances. In (2.40), and as shown in Figure 2.1, the path loss increases linearly relative to log10 d/dr with a slope of −10γ dB/decade, where γ is the path loss exponent. The variations due to shadowing change more rapidly, on the order of the decorrelation distance Xc.

Examples 2.3 and 2.4 illustrated the combined model for single-slope path loss and log-normal shadow ing based on the measurements in Table 2.2, where path loss obeys the single-slope path loss model with K = −31.53 dB and path loss exponent γ = 3.71 and where shadowing obeys the log-normal model with mean given by the path loss model and standard deviation σψdB = 3.65 dB.

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2.7.2 Outage Probability

The combined effects of path loss and shadowing have important implications for wireless system design. In wireless systems there is typically a target minimum received power level Pmin or, equivalently, a minimum SNR SNRmin = Pmin/(N0B) below which performance becomes unacceptable (e.g., the caller’s voice on a cell phone is unintelligible). However, with shadowing the received power at any given distance from the transmitter is log normally distributed with some probability of falling below Pmin. We define outage probability Pout(Pmin, d) to be the probability that the received power at a given distance d, Pr(d), falls below Pmin: Pout(Pmin, d) = p(Pr(d) < Pmin). For the combined path loss and shadowing model of Section 2.8 this becomes

p(Pr(d) ≤ Pmin)=1 − Q

$Pmin − (Pt + 10 log10 K − 10γ log10(d/dr)) σψdB

%

, (2.41)

where the Q-function is defined as the probability that a Gaussian random variable X with mean 0 and variance 1 is greater than z:

Q(z) ! p(X>z) =

4 ∞ z

1

√2πe−y2/2dy. (2.42)

The conversion between the Q-function and complementary error function is

Q(z) = 12erfc $ z√2%. (2.43)

We will omit the arguments Pmin and d of Pout when the context is clear or in generic references to outage proba bility.

Example 2.5: Find the outage probability at 150 m for a channel based on the single-slope path loss and shadowing models of Examples 2.3 and 2.4, assuming a transmit power of Pt = 10 mW and minimum power requirement of Pmin = −110.5 dBm.

Solution: We have Pt = 10 mW = 10 dBm. Hence,

Pout(−110.5 dBm, 150 m)

= p(Pr(150 m) < −110.5 dBm)

= 1 − Q

$Pmin − (Pt + 10 log10 K − 10γ log10(d/dr))

%

σψdB

= 1 − Q

$−110.5 − (10 − 31.53 − 37.1 log10(150))

%

3.65

= 0.0120.

An outage probability of l% is a typical target in wireless system designs.



2.7.3 Cell Coverage Area and Percentage

The coverage area of a given cell in a cellular system is defined as the area of locations within the cell where the received power is above a given minimum. Consider a base station inside a circular cell of a given radius R. All

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Figure 2.6: Contours of constant received power.

mobiles within the cell require some minimum received SNR for acceptable performance. Assuming a given model for noise, the SNR requirement translates to a minimum received power Pmin throughout the cell. The transmit power at the base station is designed for an average received power at the cell boundary of P R, averaged over the shadowing variations. In the absence of shadowing, the coverage area of this system is πR2 since all locations have received power above the required minimum. However, shadowing will cause some locations within the cell to have received power below P R, and others will have received power exceeding P R. This is illustrated in Figure 2.6, where we show contours of constant received power based on a fixed transmit power at the base station for path loss and average shadowing and for path loss and random shadowing. For path loss and average shadowing, constant power contours form a circle around the base station because combined path loss and average shadowing is the same at a uniform distance from the base station. For path loss and random shadowing, the contours form an amoeba-like shape due to the random shadowing variations about the average.

The constant power contours for combined path loss and random shadowing indicate the challenge that shad owing poses in cellular system design. Specifically, it is not possible for all users at the cell boundary to receive the same power level. Thus, either the base station must transmit extra power to ensure users affected by shadowing receive their minimum required power Pmin, which causes excessive interference to neighboring cells, or some users within the cell will find their minimum received power requirement unmet. In fact, since the Gaussian distri bution has infinite tails, under this model any mobile in the cell has a nonzero probability of experiencing received power below its required minimum, even if the mobile is close to the base station. The model matches propagation scenarios in practice since a mobile may be in a tunnel or blocked by a large building, regardless of its proximity to the base station.

The cell coverage percentage is defined as the expected percentage of locations within a cell where received power exceeds Pmin. The cell coverage percentage under path loss and shadowing, also referred to as the fraction of useful service area, was derived by Reudink in [1, Chapter 2.5.3] as follows. The percentage of area within a cell where the received power exceeds the minimum required power Pmin is obtained by taking an incremental area dA at radius r from the base station in the cell, as shown in Figure 2.6. Let Pr(r) be the received power in dA from combined path loss and shadowing. Then the total area within the cell where the minimum power requirement is exceeded is obtained by integrating over all incremental areas where this minimum is exceeded:

45

C = E

\* 1

πR2 4

4

cell area

+

1[Pr(r) > Pmin in dA] dA

= 1 πR2 

E[1[Pr(r) > Pmin in dA]] dA, (2.44) cell area

where 1[·] denotes the indicator function. Define PA(r) = p(Pr(r) > Pmin) in dA. Then PA(r) = E[1[Pr(r) > Pmin in dA]]. Making this substitution in (2.44) and using polar coordinates for the integration yields

C = 1 πR2 

4

cell area

PA(r) dA = 1 πR2 

4 2π 0

4 R 0

PA(r)r dr dθ. (2.45)

The outage probability within the cell is defined as the percentage of area within the cell that does not meet its minimum power requirement Pmin; that is, Pcell

we have

out = 1 − C. Given the log-normal distribution for the shadowing,

PA = p(Pr(r) ≥ Pmin) = Q

$Pmin − (Pt + 10 log10 K − 10γ log10(r/d0))

%

σψdB

= 1 − Pout(Pmin, r), (2.46)

where Pout is the outage probability defined in (2.41) with d = r. Locations within the cell with received power below Pmin are said to be outage locations.

Combining (2.45) and (2.46) yields5

a + b ln rR1dr, (2.47)

where

C = 2R24 R 0

0

rQ

a = Pmin − Pr(R) 

σψdB, b = 10γ log10(e)

σψdB, (2.48)

and Pr(R) = Pt + 10 log10 K − 10γ log10(R/d0) is the received power at the cell boundary (distance R from the base station) due to path loss alone. Applying integration by parts to solve this integral yields a closed-form solution for C in terms of a and b:

C = Q(a) + exp

\*2 − 2ab b2

+

Q

$2 − ab b

%

. (2.49)

If the target minimum received power equals the average power at the cell boundary, Pmin = Pr(R), then a = 0 and the cell coverage percentage simplifies to

C = 12 + exp \* 2b2+Q$2b%. (2.50)



5Recall that (2.46) is generally valid only for r ≥ d0, yet to simplify the analysis we have applied the model for all r. This approximation will have little impact on cell coverage percentage, since d0 is typically very small compared to R and the outage probability for r<d0 is negligible.

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Note that with this simplification C depends only on the ratio γ/σψdB. Moreover, owing to the symmetry of the Gaussian distribution, under this assumption the outage probability at the cell boundary Pout(Pr(R), R)=0.5.

Example 2.6: Find the cell coverage percentage for a cell with the combined path loss and shadowing models of Examples 2.3 and 2.4, a cell radius of 600 m, a base station transmit power of Pt = 100 mW = 20 dBm, and a minimum received power requirement of Pmin = −110 dBm and also one of Pmin = −120 dBm.

Solution: We first consider Pmin = −110 and check if a = 0 to see whether we should use the full formula (2.49) or the simplified formula (2.50). We have Pr(R) = Pt + K − 10γ log10(600) = 20 − 31.53 − 37.1 log10(600) = −114.6 dBm ̸= −110 dBm, so we use (2.49). Evaluating a and b from (2.48) yields a = (−110 + 114.6)/3.65 = 1.26 and b = (37.1 · 0.434)/3.65 = 4.41. Substituting these into (2.49) yields

C = Q(1.26) + exp

\*2 − 2(1.26 · 4.41) 4.412

+

Q

$2 − (1.26)(4.41) 4.41

%

= .59,

which would be a very low cell coverage percentage value for an operational cellular system (lots of unhappy customers). Now considering the less stringent received power requirement Pmin = −120 dBm yields a = (−120 + 114.6)/3.65 = −1.479 and the same b = 4.41. Substituting these values into (2.49) yields C = .988, a much more acceptable value for cell coverage percentage.



Example 2.7: Consider a cellular system designed so that Pmin = Pr(R). That is, the received power due to path loss and average shadowing at the cell boundary equals the minimum received power required for acceptable per formance. Find the cell coverage percentage for path loss values γ = 2, 4, 6 and σψdB = 4, 8, 12, and explain how cell coverage percentage changes as γ and σψdB increase.

Solution: For Pmin = Pr(R) we have a = 0, so cell coverage percentage is given by the formula (2.50). The cell coverage percentage thus depends only on the value for b = 10γ log10(e)/σψdB, which in turm depends only on the ratio γ/σψdB. Table 2.3 contains cell coverage percentage evaluated from (2.50) for the different γ and σψdB values.

Table 2.3: cell coverage percentage for different γ and σψdB 

γ σψdB =4 dB σψdB =8 dB σψdB =12 dB 

2 .77 .67 .63

4 .85 .77 .71 

6 .90 .83 .77 



Not surprisingly, for fixed γ the cell coverage percentage increases as σψdB decreases; this is because a smaller σψdB means less variation about the mean path loss. Without shadowing we have 100% cell coverage percentage (since Pmin = Pr(R)) and so we expect that, as σψdB decreases to zero, cell coverage percentage increases to 100%. It is a bit more puzzling that for a fixed σψdB the cell coverage percentage increases as γ increases, since a larger γ implies that received signal power falls off more quickly. But recall that we have set Pmin = Pr(R), so the faster power falloff is already taken into account (i.e., we need to transmit at much higher power with γ = 6 than with γ = 2 for this equality to hold). The reason cell coverage percentage increases with path loss exponent under this assumption is that, as γ increases, the transmit power must increase to satisfy Pmin = Pr(R). This results in higher average power throughout the cell, yielding a higher percentage of locations in the cell that have the desired minimum power.

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2.8 General Ray Tracing

In a typical urban or indoor environment, a radio signal transmitted from a fixed source will encounter multiple objects in the environment, as shown in Figure 2.7. These objects produce reflected, diffracted, or scattered copies of the transmitted signal, which are not captured by the free-space and two-ray models discussed earlier. These additional copies of the transmitted signal, known as multipath components, can be attenuated in power, delayed in time, and shifted in phase and/or frequency with respect to the LOS signal path at the receiver. The transmitted signal and its multipath components are summed together at the receiver, which can produce rapid fluctuations in received signal power due to constructive and destructive combining of the components. This multipath fading was also exhibited in the two-ray model. If the arriving signal components have delay differences that exceed a symbol time, this leads to distortion in the received signal relative to the transmitted signal.



Figure 2.7: Reflected, diffracted, and scattered wave components.

In ray tracing we assume a finite number of reflectors with known location and dielectric properties. The details of the multipath propagation can then be solved using Maxwell’s equations with appropriate boundary conditions. However, the computational complexity of this solution makes it impractical as a general modeling tool. Ray-tracing techniques approximate the propagation of electromagnetic waves by representing the wavefronts as simple particles. Thus, the effects of reflection, diffraction, and scattering on the wavefront are approximated using simple geometric equations instead of Maxwell’s more complex wave equations. The error of the ray tracing approximation is smallest when the receiver is many wavelengths from the nearest scatterer and when all the scatterers are large relative to a wavelength and fairly smooth. Comparison of the ray-tracing method with certain empirical data sets shows that it can accurately model received signal power in rural areas [24], along city streets when both the transmitter and receiver are close to the ground [24, 26, 27], and in indoor environments with appropriately adjusted diffraction coefficients [28].

If the transmitter, receiver, and reflectors are all immobile, then the characteristics of the multiple received signal paths are fixed. However, if the transmitter or receiver are moving then the characteristics of the multiple paths vary with time. These time variations are deterministic when the number, location, and characteristics of the reflectors are known over time. Otherwise, statistical models must be used. Similarly, if the number of reflectors is large, or if the reflector surfaces are not smooth such that each reflector generates many signal paths, then we must use statistical approximations to characterize the received signal. We will discuss statistical fading models

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for propagation effects in Chapter 3. Hybrid models, which combine ray tracing and statistical fading, can also be found in the literature [29, 30]; however, we will not describe them here.

The most general ray tracing model includes all attenuated, diffracted, and scattered multipath components. This model uses all of the geometrical and dielectric properties of the objects surrounding the transmitter and receiver [37, 38, 39, 31]. Since this information is site-specific, general ray tracing methods are not used to obtain general theories about system performance and layout but rather for modeling propagation for a given transmit-receive configuration in a given environment. General ray tracing uses geometrical optics to trace the propagation of the LOS and reflected signal components as well as signal components from object diffraction and diffuse scattering. There is no limit to the number of multipath components at a given receiver location: the strength of each component is derived explicitly based on the locations and dielectric properties of the objects that generate them. In general, the LOS and reflected paths provide the dominant components of the received signal, since diffraction and scattering losses are high. However, in regions close to scattering or diffracting surfaces these other multipath components may dominate, especially if the LOS and reflected rays are blocked. Open source and commercial software for ray tracing has been developed for propagation modeling in both indoor and outdoor environments. In some of these programs, computer graphics are combined with aerial photographs (outdoor channels) or architectural drawings (indoor channels) to obtain a three-dimensional geometric picture of the environment.

The following section describes a ray tracing model capturing reflected rays only, including the commonly used 6-ray and 10-ray models for a signal propagating along a straight street or hallway. We then present ray tracing models for signals that are reflected as well as diffracted and scattered. We also define the important parameter of local mean received power associated with these ray tracing models.

2.8.1 Multi-Ray Reflections

Most wireless channels have more than just the single reflected ray captured in the two-ray model described in Section 2.4. The multi-ray reflection model, developed in [25, 26], is a generalization of the two-ray model based on a propagation environment called a dielectric “canyon.” This model captures all rays that experience one or more reflections on the path between the transmitter and receiver up to a certain maximum number of reflections.

In an outdoor setting, a dielectric canyon approximates propagation in a city with rectilinear streets6 with buildings along both sides of the street as well as transmitter and receiver antenna heights that are close to street level. Theoretically, an infinite number of rays can be reflected off the building fronts to arrive at the receiver; in addition, rays may also be back-reflected from buildings behind the transmitter or receiver. However, since some of the signal energy is dissipated with each reflection, signal paths corresponding to more than two or three reflections can generally be ignored. When the street layout is relatively straight, back reflections are usually negligible also. Experimental data show that the dielectric canyon model of ten rays closely approximates signal propagation through cities with a rectilinear street layout [26]. The ten-ray model incorporates the LoS path as well as all nine paths with one, two, or three reflections: specifically, there is the ground-reflected, single-wall reflected, double-wall reflected, triple-wall reflected, wall-ground reflected, and ground-wall reflected paths. There are two of each type of wall-reflected path, one for each side of the street. An overhead view of the ten-ray model is shown in Figure 2.8. A simpler model, the six-ray model, includes a subset of these ten rays: the LOS, ground reflection, two single-wall reflections, and two double-wall reflections.

For the ten-ray model, the received signal is given by

r10-ray(t) = Re

&

λ

4π

'39 i=0

Ri√Giu(t − τi)e−j2πdi/λ di

(

ej2πfct

)

, (2.51)



6A rectilinear city is flat and has linear streets that intersect at 90◦ angles, as in midtown Manhattan. 49



Figure 2.8: Overhead view of the ten-ray model. Includes the line-of-sight (LOS), ground-reflected (GR), two single-wall(SW) reflected, two double-wall(DW) reflected, two triple-wall(TW) reflected, wall-ground (W G) reflected, and ground-wall (GW) reflected rays

where di denotes the total distance traveled by the ith ray, τi = di/c, and Gi is the product of the transmit and receive antenna power gains corresponding to the ith ray. The first two terms in the summation of (2.51) correspond to the LoS and ground reflected path in the two-ray model (2.12), hence R0 = 1 since there is no reflection of the LoS path. For the ith reflection path, i > 0, the coefficient Ri is either a single reflection coefficient given by (2.16) or, if the path corresponds to multiple reflections, the product of the reflection coefficients corresponding to each reflection. The dielectric constants used in (2.51) are approximately the same as the ground dielectric, so εr = 15 is used for all the calculations of Ri. The delay spread of the 10-ray and 6-ray models equals the difference between the delay of the LOS ray and that of the reflected ray with the largest delay: maxi τi − τ0.

If we again assume a narrowband model such that u(t − τ0) ≈ u(t − τi) for all i, then the received power

corresponding to (2.51) is

Pr = Pt

where ∆φi = 2π(di − d0)/λ.

\* λ 4π

+2/////39 i=0

Ri√Gie−j∆φi di

/////2, (2.52)

Power falloff with distance in both the ten-ray model (2.52) and urban empirical data [15, 32, 33] for transmit antennas both above and below the building skyline is typically proportional to d−2, even at relatively large dis tances. Moreover, the falloff exponent is relatively insensitive to the transmitter height. This falloff with distance squared is due to the dominance of the multipath rays, which decay as d−2, over the combination of the LOS and ground-reflected rays (two-ray model), which decays as d−4. Other empirical studies [34, 35, 36] have obtained power falloff with distance proportional to d−γ , where γ lies anywhere between 2 and 6.

2.8.2 Diffraction

The propagation model for the LOS and reflected paths was outlined in the previous section. Diffraction allows a radio signal to “bend around” an object in its path to the receiver, as shown in Figure 2.9 for a wedge-shaped object. This bending phenomenon can be explained using Huygen’s principle, which states that all points on the signal wavefront can be considered as point sources for a secondary wavefront called a wavelet. When a signal is diffracted, these wavelets combine together to produce a wavefront in the new direction of propagation caused by the bending [29, Chapter 4.3]. Diffraction results from many phenomena, including hilly or irregular terrain, building and rooftop edges, or obstructions blocking the LOS path between the transmitter and receiver [2, 11, 16]. Diffraction can be accurately characterized using the geometrical theory of diffraction (GTD) [40], but the complexity of this approach has precluded its use in wireless channel modeling. Wedge diffraction simplifies the GTD by assuming the diffracting object is a wedge rather than a more general shape. This model has been

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used to characterize the mechanism by which signals are diffracted around street corners, which can result in signal attenuation exceeding 100 dB for some incident angles on the wedge [28, 39, 41, 42]. Although wedge diffraction simplifies the GTD, it still requires a numerical solution for the resulting attenuation [40, 43] and thus is not commonly used. Diffraction is most commonly modeled by the Fresnel knife-edge diffraction model because of its simplicity. The geometry of this model is shown in Figure 2.9, where the diffracting object is assumed to be asymptotically thin, which is not generally the case for hills, rough terrain, or wedge diffractors. In particular, this model does not consider diffractor parameters such as polarization, conductivity, and surface roughness, which can lead to inaccuracies [41].



Figure 2.9: Knife-edge diffraction.

The diffracted signal of Figure 2.9 travels a distance d∗ +d⋆, resulting in a phase shift of φ = 2π(d∗ +d⋆)/λ. For a LOS path of distance d, the geometry of Figure 2.9 indicates that, for h small relative to d∗ and d⋆, the signal must travel an additional distance relative to the LOS path of approximately

∆d = d∗ + d⋆ − d ≈ h2(d∗ + d⋆) 

2d∗d⋆ ;

the corresponding phase shift relative to the LOS path is approximately

∆φ = 2π∆d 

λ ≈ π2v2, (2.53)

where

v = h

52(d∗ + d⋆)

λd∗d⋆ (2.54)

is called the Fresnel–Kirchhoff diffraction parameter. The path loss of a signal at distance d∗ + d⋆ experiencing knife-edge diffraction at distance d∗ is generally a function of v. However, computing this path loss is fairly complex, requiring the use of Huygens’s principle, Fresnel zones, and the complex Fresnel integral [2]. Moreover, the resulting diffraction loss cannot generally be found in closed form. Approximations for the additional path loss due to knife-edge diffraction relative to free-space path loss (in dB) as a function of v in (2.54) are given in [16, 44]

as

⎧⎪⎪⎪⎪⎪⎪⎨

L(v)dB =

⎪⎪⎪⎪⎪⎪⎩

0, v< −1, 20 log10[.5 − .62v], −0.8 ≤ v < 0, 20 log10[.5e−.95v], 0 ≤ v < 1, 20 log10[.4 − ..1184 − (.38 − .1v)2], 1 ≤ v ≤ 2.4, 20 log10[.225/v], v> 2.4.

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(2.55)

The knife-edge diffraction received signal model is obtained by adding this loss to the free-space path loss of (2.11), yielding the following formula for the received diffracted signal:

r(t) = Re

&

λ

4π

'

L(v)√Gdu(t − τ )e−j2π(d∗+d⋆)/λ d∗ + d⋆

(

ej2πfct

)

, (2.56)

where √Gd is the antenna gain and τ = (d∗ + d⋆)/c is the delay associated with the defracted ray. In addition to diffracted rays, there may also be rays that are diffracted multiple times, or rays that are both reflected and diffracted. Models exist for including all possible permutations of reflection and diffraction [45]; however, the attenuation of the corresponding signal components is generally so large that these components are negligible relative to the noise. Diffraction models can also be specialized to a particular environment or frequency band. For example, a model for diffraction from rooftops and buildings in cellular systems was developed by Walfisch and Bertoni in [46]. This rooftop diffraction model was extended to multi-antenna base stations in [47], where it produced a “keyhole” phenomenon that reduced the effective spatial dimensions of the channel. Two dimensional diffraction models for urban small cells, where signals propagate below rooftop levels, is developed in [48]. Application of knife-edge and more general diffraction models to millimeter wave propagation is discussed in [49, 50].

2.8.3 Scattering



Figure 2.10: Scattering.

A scattered ray, shown in Figure 2.10 by the segments sα and sβ, has a path loss proportional to the product of sα and sβ. This multiplicative dependence is due to the additional spreading loss that the ray experiences after scattering. The received signal due to a scattered ray is given by the bistatic radar equation [51]:

r(t) = Re

&

λ

4π

'√Gsσu(t − τ )e−j2π(s∗+s⋆)/λ √4πs∗s⋆

(

ej2πfct

)

, (2.57)

where τ = (s∗ +s⋆)/c is the delay associated with the scattered ray; σ (in square meters) is the radar cross-section of the scattering object, which depends on the roughness, size, and shape of the scatterer; and Gs is the antenna power gain. The model assumes that the signal propagates from the transmitter to the scatterer based on free-space propagation and is then re-radiated by the scatterer with transmit power equal to σ times the received power at the scatterer. From (2.57), the path loss associated with scattering is

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Pr dBm = Pt dBm + 10 log10(Gs) + 20 log10(λ) + 10 log10(σ)

−30 log(4π) − 20 log10(s∗) − 20 log10(s⋆). (2.58)

Empirical values of 10 log10 σ were determined in [52] for different buildings in several cities. Results from this study indicate that 10 log10 σ in dBm2 ranges from −4.5 dBm2 to 55.7 dBm2, where dBm2 denotes the dB value of the σ measurement with respect to one square meter.

2.8.4 Multipath Model with Reflection, Diffraction, and Scattering

The received signal is determined from the superposition of all the components due to the multiple rays. Thus, if we have a LOS ray, Nr reflected rays, Nd diffracted rays, and Ns scattered rays, the total received signal is

: λ

\*√G0u(t)e−j2πd0/λ

Nr

Ri√Giu(t − τi)e−j2πdi/λ

rtotal(t) = Re

4π

d0+3 i=1

di

Nd

+3 j=1

Ns

Lj (v).Gdju(t − τj )e−j2π(d∗j +d⋆j )/λ d∗jd⋆j .Gsk σku(t − τk)e−j2π(s∗k+s⋆k)/λ

+

;

(2.59)

+3 k=1

√4πs∗ks⋆k

ej2πfct

,

where τj and τk are, respectively, the time delays of the given diffracted and scattered rays, (dαj , dβj ) and (sαk , sβk) are, respectively, the distance pair the signal travels before and after the object causing the diffrac tion or scattering, and the other parameters are as defined in the model (2.51) with the LoS and reflections only. The received power Pr of rtota1(t) and the corresponding path loss Pr/Pt are then obtained from (2.59). The delay spread of this model equals the difference between the delay of the LOS ray and that of the reflected, diffracted, or scattered ray that has the largest delay.

Any of these multipath components may have an additional attenuation factor if its propagation path is blocked by objects such as buildings outdoors or walls indoors. In this case, the attenuation factor of the obstructing object multiplies the component’s path loss term in (2.59). This attenuation loss will vary widely, depending on the material and depth of the object [11, 53]. Models for random loss due to attenuation are described in Section 2.6.

2.8.5 Multi-Antenna and MIMO Systems

The ray tracing techniques described in Sections 2.8.1- 2.8.4 assume single antennas at the transmitter and receiver. The same ray tracing techniques can be used for systems with multiple antennas at the transmitter and/or receiver by superposition, whereby the single-antenna analysis is applied to each transmit-receive antenna pair and the results are summed together. When both the transmitter and receiver have multiple antennas, the system is called multiple input multiple output (MIMO) system. While the application of ray tracing to MIMO systems is straightforward, the computational complexity of this approach grows linearly with the product of the number of transmit and receive antennas. As a result, ray tracing approximations to lower this complexity have been developed, including clustering of rays [54] and preprocessing environmental data [56]. A summary of ray tracing methods for MIMO systems can be found in [55].

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2.8.6 Local Mean Received Power

The path loss computed from all ray-tracing models is associated with a fixed transmitter and receiver location. In addition, ray tracing can be used to compute the local mean received power Pr in the vicinity of a given receiver location by adding the squared magnitude of all the received rays. This has the effect of averaging out local spatial variations due to phase changes around the given location. Local mean received power is a good indicator of link quality and is often used in cellular system functions like power control and handoff [83].

2.9 Measurement-Based Propagation Models

Most mobile communication systems operate in complex propagation environments that cannot be accurately modeled by free-space path loss or ray tracing. A number of analytical path loss models have been developed over the years by researchers as well as standards bodies to model path loss in typical wireless environments including urban macrocells and small cells, suburban areas, rural areas, and inside buildings with different characteristics [11, 57, 59, 74, 17]. These models are generally based on large empirical measurement campaigns that can range over a variety of distances, frequency ranges, geographical regions for outdoor models, and building types for indoor models. These analytical models have the highest accuracy when they are applied to propagation conditions similar to those under which the empirical measurements that the models are based on were made. In this section we describe the most common analytical path loss models based on empirical measurements for both indoor and outdoor systems.

Analytical models characterize Pr/Pt as a function of distance, so path loss is well-defined. In contrast, empirical measurements of Pr/Pt as a function of distance include the effects of path loss, shadowing, multipath, and other site-specific factors that affect propagation. In order to remove multipath effects, empirical measurements for path loss typically average their received power measurements and the corresponding path loss at a given distance over several wavelengths. This average path loss is called the local mean attenuation (LMA) at distance d, and it generally decreases with d owing to free-space path loss and signal obstructions. The LMA in a given environment, such as a city, depends on the specific location of the transmitter and receiver corresponding to the LMA measurement. To characterize LMA more generally, measurements are typically taken throughout the environment and possibly in multiple environments with similar characteristics. Thus, the empirical path loss PL(d) for a given environment (a city, suburban area, or office building) is defined as the average of the LMA measurements at distance d averaged over all available measurements in the given environment. For example, empirical path loss for a generic downtown area with a rectangular street grid might be obtained by averaging LMA measurements in New York City, downtown San Francisco, and downtown Chicago. The empirical path loss models given in this section are all obtained from average LMA measurements. Empirical path loss models can also be developed or refined using measurement-driven learning applied to continuous LMA data collection [75].

2.9.1 Okumura Model

One of the most well-known models for signal prediction in large urban macrocells is the Okumura model [84]. This model is applicable over distances of 1–100 km and frequency ranges of 150–1500 MHz. Okumura used extensive measurements of base station-to-mobile signal attenuation throughout Tokyo to develop a set of curves giving median attenuation relative to free space of signal propagation in irregular terrain. The base station heights for these measurements were 30–100 m, a range whose upper end is higher than typical base stations today. The empirical path loss formula of Okumura at distance d parameterized by the carrier frequency fc is given by

PL(d) dB = L(fc, d) + Aµ(fc, d) − G(ht) − G(hr) − GAREA, (2.60) 54

where L(fc, d) is free-space path loss at distance d and carrier frequency fc, Aµ(fc, d) is the median attenuation in addition to free-space path loss across all environments, G(ht) is the base station antenna height gain factor, G(hr) is the mobile antenna height gain factor, and GAREA is the gain due to the type of environment. The values of Aµ(fc, d) and GAREA are obtained from Okumura’s empirical plots [84, 11]. Okumura derived empirical formulas for G(ht) and G(hr) as follows:

G(ht) = 20 log10(ht/200), 30 m < ht < 1000 m; (2.61)

&

G(hr) =

10 log10(hr/3), hr ≤ 3 m,

20 log10(hr/3), 3 m < hr < 10 m. (2.62)

Correction factors related to terrain are also developed in [84] that improve the model’s accuracy. Okumura’s model has a 10–14 dB empirical standard deviation between the path loss predicted by the model and the path loss associated with one of the measurements used to develop the model. The expected error in using Okumura’s model for environments not based on these measurements is generally higher.

2.9.2 Hata Model

The Hata model [85] is an empirical formulation of the graphical path loss data provided by Okumura and is valid over roughly the same range of frequencies, 150–1500 MHz. This empirical model simplifies calculation of path loss because it is a closed-form formula and is not based on empirical curves for the different parameters. The standard formula for empirical path loss in urban areas under the Hata model is

PL,urban(d) dB = 69.55 + 26.16 log10(fc) − 13.82 log10(ht) − a(hr)

+ (44.9 − 6.55 log10(ht)) log10(d). (2.63)

The parameters in this model are the same as under the Okumura model, and a(hr) is a correction factor for the mobile antenna height based on the size of the coverage area [85, 11]. For small to medium-sized cities, this factor is given by

a(hr) = (1.1 log10(fc) − .7)hr − (1.56 log10(fc) − .8) dB,

and for larger cities at frequencies fc > 300 MHz by

a(hr)=3.2(log10(11.75hr))2 − 4.97 dB.

Corrections to the urban model are made for suburban and rural propagation, so that these models are (respec tively)

PL,suburban(d) dB = PL,urban(d) dB − 2[log10(fc/28)]2 − 5.4 (2.64)

and

PL,rural(d) dB = PL,urban(d) dB − 4.78[log10(fc)]2 + 18.33 log10(fc) − K, (2.65) where K ranges from 35.94 (countryside) to 40.94 (desert). Unlike the Okumura model, the Hata model does not provide for any path-specific correction factors. The Hata model well approximates the Okumura model for distances d > 1 km. Hence it is a good model for early-generation cellular systems operating in cities, but it

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does not model propagation well in cellular systems with smaller cell sizes, multiple antennas, or higher frequen cies. Indoor environments are also not captured by the Hata model. As cellular systems evolved beyond their early deployments, research into more complex propagation models was undertaken by multiple organizations, as described in the next section.

2.9.3 Cellular System Models

Following the success of early cellular systems, funding agencies and standards bodies worldwide launched projects to develop standardized cellular system channel models for the environments, carrier frequencies, cell sizes, and multiple antenna characteristics of these evolving systems. These standardized channel models can be used to evaluate different candidate cellular technologies in a uniform manner, and also provide guidelines for expected system performance in typical environments and operating conditions. Cellular system channel models span frequency ranges from .5 GHz to 100 GHz, indoor and outdoor environments with both LOS and non-LOS propagation, small and large cells sizes, and multiple antenna transmitters and receiver of different antenna heights, elevation angles, and polarizations. Models for all of these scenarios have been developed by 3GPP and the ITU [57, 58], whereas other standardized models are applicable to a subset of these settings. To illustrate some basic properties of these standardized models, in this section we provide the LOS path loss formulas for a few stan dardized models, which follow either the single-slope or dual-slope formulas ( (2.24) and (2.28), respectively) with model-specific constants. Extensions of these models to more complex environments is also discussed. More details on the unique characteristics of millimeter wave propagation are provided in Section 2.9.5.

An early project to standardize channel models for cellular systems was undertaken by the European Cooper ation in Science and Technology (COST). The COST 207 program, launched in 1984, brought together industry, government, and academia under a common umbrella to develop common propagation models for the emerging 2G cellular systems. The subsequent COST 231 project was focused on extending earlier channel models to carrier frequencies up to 2 GHz. The resulting model, referred to as the COST 231 extension to the Hata model [86], has the following path loss expression:

PL, urban(d) dB = 46.3 + 33.9 log10(fc) − 13.82 log10(ht) − a(hr)

+ (44.9 − 6.55 log10(ht)) log10(d) + CM , (2.66)

where a(hr) is the same correction factor as in the Hata model and CM = 0 dB for medium-sized cities and suburbs while CM = 3 dB for metropolitan areas. The COST 231 extension to the Hata model is restricted to the following range of parameters: 1.5 GHz < fc < 2 GHz, 30 m < ht < 200 m, 1 m < hr < 10 m, and 1 km <d< 20 km. Given the Okumura measurements on which it is based, this model is most accurate for heterogeneous macrocell-only cellular system architectures with single-antenna base stations and terminals. The COST 231 model was further extended to incorporate new features and environments of the evolving cellular systems and standards, including diffraction and non-LOS propagation (COST-231-WI)[29], multiple base station antennas (COST 259) [88], and multiple antennas at both the base stations and the mobile terminals (COST 273, COST 2100) [89, 90].

The 3GPP standards body maintains a family of channel models characterizing signal propagation for the different environments and frequencies in which their cellular systems operate [57]. These include path loss models for both LOS and non-LOS propagation in outdoor macrocells and small cells as well as for indoor of fice buildings. The 3GPP outdoor propagation models follow the dual-slope path loss model (2.28); they have a distance-dependent pathloss exponent that takes on two different values, γ1 and γ2, for distances that are, re spectively, smaller or greater than the breakpoint distance. The breakpoint distance is set to the critical distance dc = 4hthr/λ given by (2.22) in the two-ray model based on the effective transmit (ht) and receive (hr) antenna

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heights7. In particular, the 3GPP path loss for LOS propagation in outdoor macrocells and small cells is given by &

PL(d) dB =

20 log10(fc) + κ + 10γ1 log10(d), d ≤ dc, 20 log10(fc) + κ + 10γ2 log10(d) + η log10 !d2c + (ht − hr)2#, d>dc(2.67)

for d in meters and fc in GHz. The 3GPP indoor office model is governed by the first terms of (2.67) for all distances (i.e., there is no breakpoint distance in the indoor model due to the shorter distances involved). At distances below dc this model has three separate terms: the constant κ that captures environment and antenna parameters independent of distance and frequency, the term 20 log10 fc that captures frequency-dependent path loss, and the term 10γi log10(d) that captures distance-dependent path loss. The frequency-dependent path loss 20 log10 fc is the same as in free space (2.10), and this term along with κ are captured by the constant K in the single-slope model (2.24). At distances above dc the macrocell and small cell models have a fourth term that captures path loss dependence on antenna heights. The values of κ, γi, and η for the macrocell, small cell, and indoor office models are given in Table 2.4.

Table 2.4: Parameters for 3GPP LOS Path Loss Models

Parameter Macrocell Small Cell Indoor Office 

κ 28 32.4 32.4 

γ γ1 = 22, γ2 = 40 γ1 = 21, γ2 = 40 γ1 = 17.3 

η -9 -9.5 NA 



The 3GPP family of channel models also include non-LOS propagation models that take path loss to be the maximum of the LOS path loss model and a different non-LOS path loss model. The non-LOS path loss model is of the same form as (2.67) but with different parameters and a fourth term that depends only on the receiver height hr. There is also a 3GPP rural macrocell model that includes additional parameters of the environment such as building heights and street widths. The 3GPP Spatial Channel Model (SCM) incorporates MIMO parameters. Details can be found in [57]. The ITU channel models are similar to the 3GPP channel family and cover the same frequency range of .5-100 GHz [58].

The European WINNER (Wireless World Initiative New Radio) I and II projects created a similar family of channel models as 3GPP for cellular system propagation under different conditions and environments [59]. The WINNER path loss model is similar to the single-slope model (2.24) with the constant K broken into 3 separate terms. Specifically, the path loss in the WINNER propagation model 10 log10(Pt/Pr) is given by

PLdBm = 10γ log10(d) + κ + β log10(.2fc) + X (2.68)

for d in meters and fc in GHz, where γ is the path loss exponent, κ is the path loss at the antenna near-field reference distance, typically a few signal wavelengths [12], which is independent of frequency, C log10(.2fc) is the frequency-dependence term, and X is an optional environment-specific term that may, for example, capture a fixed attenuation at all distances. The Winner II model includes MIMO channel parameters.

The COST, 3GPP, ITU, and WINNER families of channel models are widely used by researchers and prac titioners for modeling path loss in cellular systems at frequencies below 6 GHz, with the 3GPP and ITU models covering frequencies up to 100 GHz. These models also include delay spread parameters ranging from tens of nanoseconds for indoor environments to hundreds of nanoseconds for outdoor environments. Alternative empiri cal channel models specifically for MIMO systems include the WIMAX 802.16 and Stanford University Interim



7In the 3GPP model the effective antennas heights are the actual heights reduced by 1m.

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(SUI) channel models [76]. As cellular system architectures have evolved to include large and small cells, device to-device communications, base stations mounted on UAVs and drones, large antenna arrays, millimeter wave frequencies, and very high speed mobile devices, there have been extensive measurement campaigns to charac terize propagation under these new configurations and conditions [77, 78, 79]. In addition, learning algorithms that continually update channel models based on collected data have also been developed [80, 81, 82]. Clearly measurement-based models for cellular systems will continue to evolve as new system architectures, frequency bands, device requirements, and channel measurement tools are introduced.

2.9.4 Wi-Fi Channel Models

Prior to 802.11n, performance evaluation of 802.11 WLAN systems was done using channel models for 2.4 GHz (802.11g) and 5 GHz (802.11a) that had been developed by individual researchers or companies. There was no general consensus on which models to use for performance evaluation of these earlier systems. To address this shortcoming, the 802.11n task group (TG) was formed to develop a standard set of models, called the TGn models, to evaluate 802.11n systems. As per the specifications of the 802.11n standard, these models include both 20 MHz and 40 MHz channels with up to 4 antennas at the transmitter and receiver. Six different TGn channel models (models A-F) comprise the TGn family to characterize propagation in different size homes and offices as well as outdoors. All six follow the dual-slope path loss model (2.28) with a path loss exponent γ1 = 2 prior to the breakpoint distance dBP and γ2 = 3.5 above this distance. The breakpoint distance dBP = 5 m in models A, B, and C, which correspond to LOS only, small home, and large home, respectively. This distance ranges from 10-30 m in models D-F (corresponding to small office, large office, and outdoors, respectively). The delay spread of the multipath, normalized with respect to the LOS path, becomes the difference between the delay of the multipath component with the largest delay and the LOS path. In these models this normalized delay spread ranges from 0 ns for the A model to 150 ns for the F model. Multipath reflections are assumed to arrive in clusters based on the indoor propagation model developed in [60]. Antenna correlation is modeled using the Kronecker model described in [61]. Shadow fading follows the log-normal model with standard deviation σψdB = 3 for distances d<dBP. For d>dBP, σψdB = 4 for Models A and B, σψdB = 5 for Models C and D, and σψdB = 6 for Models E and F. More details on the TGn models can be found in [62].

The TGac family of channels extends the TGn model to support performance evaluation of 802.11ac sys tems. This model increases the number of transmit and receive antennas using the same correlation model as TGn, increases channel bandwidth up to 160 MHz, and modifies the multipath clustering models to account for simulta neous transmission to multiple users (multiuser MIMO). The TGax family of channel models provides additional environmental scenarios for dense deployments both indoors and outdoors. Details of these extensions can be found in [63, 64].

2.9.5 Millimeter Wave Models

The millimeter (mmW) frequency band, spanning the 30-300 GHz range, was not used in the first few generations of Wi-Fi and cellular systems due to its high path loss as well as the high cost and poor performance of mmW hardware. However, as demand for increased data rates in wireless systems continued unabated with each new generation, utilizing the large amount of available spectrum in this band grew increasingly appealing. Moreover, in 2005 the FCC in the US created more flexible rules for spectrum allocation in the lower part of this band to increase its use [65]. Following this ruling, a few commercial mmW communication systems were developed for in-home use, but the technology was not yet competitive with Wi-Fi. Interest in utilizing mmW wave spectrum for cellular communications also started to grow, leading to outdoor channel measurement campaigns as well as cellular system designs to overcome the propagation challenges at these high frequencies. This section provides an overview of mmW propagation characteristics.

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Since path loss is inversely proportional to frequency squared in the free space path loss formula (2.6), this loss is much higher at mmW frequencies than at lower frequencies. However, this large free-space path loss can be compensated by high gains associated with directional or multiple antenna technologies. In addition to free-space path loss, signal propagation in the mmW band experiences significant attenuation at specific carrier frequencies due to interactions with oxygen, water, and other molecular components in the air. Attenuation due to shadowing from objects and people is more severe at these frequencies and shadowing can also cause scattering of directed beams. As a result, there is typically only a small number of multipath components in mmWave channels whose received power is above the noise floor.

The resonant frequency of oxygen (O2) molecules leads to heavy absorption, up to 15 dB/km, for signals transmitted through the air at carrier frequencies around 60 GHz. There is also significant absorption due to the resonant frequencies of O2 or water vapor (H2O) at 119, 183, and 325 GHz, with a small peak at 24 GHz. The combined effects of free-space path loss and the O2 and H2O absorption in non-rainy conditions is illustrated in Fig. 2.11 from [67]. It can be seen from this figure that atmospheric absorption is somewhat larger at sea level than at higher altitudes due to increased water vapor concentration.



Figure 2.11: Path Loss, O2, and H2O Absorption in Millimeter Wave Propagation [67].

Millimeter wave propagation is also affected by rain absorption. In particular, rain causes scattering since the wavelength of a mmW signal is roughly the same size as a raindrop. The higher the rain rate, the more scattering occurs and hence the higher the absorption. A model for attenuation at different rain rates for mmW propagation was developed by the ITU [68], as illustrated in Fig. 2.12 from [66]. A simple approximation, developed in [69], gives the rain attenuation as .95R.77 for R the rain rate in mm/hr.

While atmospheric absorption and rain attentuation entail very large path loss over large distances, most 59

mmWave systems are designed to operate over relatively short distances, less than 500 m. The losses illustrated in Figs. 2.11-2.12 for these short distances can be compensated through a combination of increased transmit power, antenna techniques, modulation choice, and channel coding.



Figure 2.12: Rain Attenuation in Millimeter Wave Propagation [66].

As interest in mmW frequency utilization for cellular and Wi-Fi grew, extensive measurement campaigns were undertaken in different indoor and outdoor environments [70], with analytical channel models developed to approximate these empirical measurements. The most widely known of these models, summarized in [71], are based on work done by four different organizations: the 3GPP standards body, the 5G Channel Model (5GCM) ad hoc group, the Mobile and wireless communications Enablers for the Twentytwenty (2020) Information Soci ety (METIS) research project, and the Millimeter-Wave Based Mobile Radio Access Network for 5G Integrated Communications (mmMAGIC) research project. The models for each group capture propagation for four different categories of environments: outdoor macrocell, outdoor small cell, home, and commercial building. In addition to models for path loss and shadowing, these mmW propagation models add a new component called LOS proba bility, which captures the probability that the LOS signal path between the transmitter and receiver is not blocked or heavily attenuated. While this probability is different for the four environment categories within each family of models, in all cases it depends on the transmit-receive distance and does not depend on the carrier frequency [72]. Millimeter wave path loss in all these models is assumed to follow the dual-slope model (2.28), where the breakpoint distance is different for each of the four environments. The shadowing standard deviation for each of the four environments is a fixed constant. As expected, shadowing in these models is larger for non-LOS com pared to LOS environments, and ranges from about 1-8 dB. A summary of the differences between mmW and microwave channel models can be found in [72]. The sensitivity of performance analysis to the particular channel model adopted is demonstrated in [79]. The terahertz (THz) band (.3-3 THz) offers even more spectrum than the mmw band, along with more challenging propagation characteristics [73].

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Table 2.5: Typical partition losses at fc ≈ 1 GHz 

Partition type Partition loss (dB)

Cloth partition 1.4 

Double plasterboard wall 3.4 

Foil insulation 3.9 

Concrete wall 13 

Aluminum siding 20.4 

All metal 26 



2.9.6 Indoor Attentuation Models

Indoor environments differ widely in the materials used for walls and floors, the layout of rooms, hallways, win dows, and open areas, the location and material in obstructing objects, the size of each room, and the number of floors. All of these factors have a significant impact on path loss in an indoor environment. While the TGn, TGac, and TGax Wi-Fi models capture propagation in several generic indoor environments, the nature of these generic models make them inaccurate in characterizing propagation for any particular indoor environment. In this section we describe indoor models that can be tailored to such environments.

Indoor path loss models must accurately capture the effects of attenuation across floors due to partitions as well as between floors. Measurements across a wide range of building characteristics and signal frequencies indicate that the attenuation per floor is greatest for the first floor that is passed through and decreases with each subsequent floor. Specifically, measurements in [18, 94, 19, 20] indicate that, at 900 MHz, the attenuation when transmitter and receiver are separated by a single floor ranges from 10–20 dB, while subsequent attenuation is 6–10 dB per floor for the next three floors and then a few decibels per floor for more than four floors. At higher frequencies the attenuation loss per floor is typically larger [94, 95]. The attenuation per floor is thought to decrease as the number of attenuating floors increases because of the scattering up the side of the building and reflections from adjacent buildings. Partition materials and dielectric properties vary widely and thus so do partition losses. Measurements for the partition loss at different frequencies for different partition types can be found in [11, 18, 96, 97, 23], and Table 2.5 indicates a few examples of partition losses measured at around 1 GHz from this data. The partition loss obtained by different researchers for the same partition type at the same frequency often varies widely, so it is difficult to make generalizations about partition loss from a specific data set.

The experimental data for floor and partition loss can be added to an analytical or empirical dB path loss model PL(d) as

Nf

Pr dBm = Pt dBm − PL(d) −3 i=1

Np

FAFi −3 i=1

PAFi, (2.69)

where FAFi represents the floor attenuation factor for the ith floor traversed by the signal and PAFi represents the partition attenuation factor associated with the ith partition traversed by the signal. The number of floors and partitions traversed by the signal are Nf and Np, respectively.

Another important factor for indoor systems whose transmitter is located outside the building is outdoor-to indoor penetration loss. Measurements indicate that this penetration loss is a function of frequency, height, and the exterior wall or window material. For carrier frequencies from .9-3 GHz penetration loss measurements range from 8 dB to 20 dB [2, 21, 100, 101]. Penetration loss for mmWave signals generally increases with frequency. Hence, the 3GPP, ITU, 5GCM, and mmMAGIC propagation models include a component for outdoor-to-indoor penetration loss as a function of both frequency and the exterior wall or window material [57, 58, 70]. The ITU penetration loss model as a function of frequency fc and for different wall and window materials is given in Table 2.6. In general penetration loss decreases by about 1.4 dB per floor at floors above the ground floor. This

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Table 2.6: ITU penetration loss model

Material Penetration loss (dB) 

Standard multi-pane glass 2+0.2fc 

Infrared reflective glass 23 + .3fc 

Concrete 5+4fc 

Wood 4.85 + 0.12fc 



decrease is due to reduced clutter at higher floors and the higher likelihood of a LOS path. 62

Chapter 2 Problems

1. Under the free-space path loss model, find the transmit power required to obtain a received power of 1 dBm for a wireless system with isotropic antennas (Gt = Gr = 1) and a carrier frequency fc = 5 GHz, assuming a distance d = 10 m. Repeat for d = 100 m.

2. For the two-ray model with transmitter–receiver separation d = 100 m, ht = 10 m, and hr = 2m, find the delay spread between the two signals.

3. For the two-ray model, show how a Taylor series approximation applied to (2.14) results in the approximation ∆φ = 2π(d1 − d0) 

λ ≈ 4πhthr

λd .

4. For the two-ray model, derive an approximate expression for the distance values below the critical distance dc at which signal nulls occur.

5. Find the critical distance dc under the two-ray model for a large macrocell in a suburban area with the base station mounted on a tower or building (ht = 20 m), the receivers at height hr = 3m, and fc = 2 GHz. Is this a good size for cell radius in a suburban macrocell? Why or why not?

6. Suppose that, instead of a ground reflection, a two-ray model consists of a LOS component and a signal reflected off a building to the left (or right) of the LOS path. Where must the building be located relative to the transmitter and receiver for this model to be the same as the two-ray model with a LOS component and ground reflection?

7. Consider a two-ray channel with impulse response h(t) = α1δ(t) + α2δ(t − .022 µs), so for input signal s(t) to the channel, the output is r(t) = h(t) ∗ s(t). Find the distance separating the transmitter and receiver, as well as α1 and α2, assuming free-space path loss on each path with a reflection coefficient of −1. Assume the transmitter and receiver are located 8 m above the ground and that the carrier frequency is 900 MHz.

8. Directional antennas are a powerful tool to reduce the effects of multipath as well as interference. In par ticular, directional antennas along the LOS path for the two-ray model can reduce the attenuation effect of ground wave cancellation, as will be illustrated in this problem. Assume the reference distance dr = 1 m. Plot the dB power (10 log10 Pr) versus log distance (log10 d) for the two-ray model with parameters fc = 900 MHz, R = −1, ht = 50 m, hr = 2 m, G0 = 1, and the following values for G1 : G1 = 1, .316, .1, and .01 (i.e., G1 = 0, −5, −10, and − 20 dB, respectively). Each of the four plots should range in distance from d = 1 m to d = 100 km. Also calculate and mark the critical distance dc = 4hthr/λ on each plot, and normalize the plots to start at approximately 0 dB. Finally, show the piecewise linear model with flat power falloff up to distance ht, falloff 10 log10(d−2) for ht <d<dc, and falloff 10 log10(d−4) for d ≥ dc. (On the power loss versus log distance plot, the piecewise linear curve becomes a set of three straight lines of slope 0, 2, and 4, respectively.) Note that at large distances it becomes increasingly difficult to have G1 ≪ G0 because this requires extremely precise angular directivity in the antennas.

9. Under what conditions is the single-slope path loss model (2.23) the same as the free-space path loss model (2.7)?

10. Consider a receiver with noise power –160 dBm within the signal bandwidth of interest. Assume a single slope path loss model with d0 = 1m, K obtained from the free-space path loss formula with isotropic antennas and fc = 1 GHz, and γ = 4. For a transmit power of Pt = 10 mW, find the maximum distance between the transmitter and receiver such that the received signal-to-noise power ratio is 20 dB.

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11. For the set of empirical measurements of Pr/Pt given in Table 2.2, find the path loss exponent γ and param eter K that minimizes the MSE between the single-slope model (2.24) in dB and the empirical dB power measurements, assuming that dr = 1m. This problem differs from Example 2.5.1 in that now the two parameters γ and K are jointly optimized to minimize MSE, rather than fixing the value of K based on free-space path loss or the measured attenuation at the reference distance dr. Find the received power at 150 m for this single-slope path loss model with transmit power of 1 mW (0 dBm) and compare with the result in Example 2.5.1. Does the better fitting of the data through this two-dimensional optimization lead to a larger or smaller value for this received power? Will this always be the case for any set of empirical measurements (why or why not)?

12. Find parameters for a multi-slope model with three segments to approximate the two-ray model path loss (2.13) over distances between 10 and 1000 meters, assuming ht = 10 m, hr =2m, and G0 = G1 = 1. Plot the path loss and the piecewise linear approximation using these parameters over this distance range.

13. This problem shows how different propagation models can lead to very different SNRs (and therefore differ ent link performance) for a given system design. Consider a linear cellular system using frequency division, as might operate along a highway or rural road (see Figure 2.13). Each cell is allocated a certain band of frequencies, and these frequencies are reused in cells spaced a distance d away. Assume the system has square cells, 2 km per side, and that all mobiles transmit at the same power P. For the following propagation models, determine the minimum distance that the cells operating in the same frequency band must be spaced so that uplink SNR (the ratio of the minimum received signal-to-interference or S/I power from mobiles to the base station) is greater than 20 dB. You can ignore all interferers except those from the two nearest cells operating at the same frequency.



Figure 2.13: Linear cellular system for Problem 2-14.

(a) Propagation for both signal and interference follow a free-space model.

(b) Propagation for both signal and interference follow the single-slope path loss model (2.23) with d0 = 100 m, K = 1, and γ = 3.

(c) Propagation for the signal follows the single-slope path loss model with d0 = 100 m, K = 1, and γ = 2, while propagation of the interfererence follows the same model but with γ = 4.

14. Show that the log-normal distribution for ψ given in (2.31) yields the Gaussian distribution for ψdB given in (2.34).

15. Table 2.7 lists a set of empirical path loss measurements. Assume a carrier frequency fc = 2 GHz.

(a) Find the parameters of a single-slope path loss model plus log-normal shadowing that best fit this data assuming K is calculated from free-space path loss at the reference distance dr = 1 m. (b) Find the path loss at 2 km based on this model.

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Table 2.7: path loss measurements for Problem 2-14

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Distance from transmitter Pr/Pt 

5 m –60 dB 

25 m –80 dB 

65 m –105 dB 

110 m –115 dB 

400 m –135 dB 

1000 m –150 dB 



(c) Find the outage probability at a distance d assuming the received power at d due to path loss alone is 10 dB above the required power for non-outage.

16. Consider a cellular system operating at 900 MHz where propagation follows free-space path loss with vari ations about this path loss due to log-normal shadowing with standard deviation σ = 6 dB. Suppose that for acceptable voice quality a signal-to-noise power ratio of 15 dB is required at the mobile. Assume the base station transmits at 1 W and that its antenna has a 3 dB gain. There is no antenna gain at the mobile, and the receiver noise in the bandwidth of interest is −60 dBm. Find the maximum cell size such that a mobile on the cell boundary will have acceptable voice quality 90% of the time.

17. In this problem we will simulate the log-normal fading process over distance based on the autocovariance model (2.39). As described in the text, the simulation first generates a white noise process and then passes it through a first-order filter with a pole at e−δ/Xc . Assume Xc = 20 m and plot the resulting log-normal fading process over a distance d ranging from 0 m to 200 m, sampling the process every meter. You should normalize your plot about 0 dB, since the mean of the log-normal shadowing is captured by path loss.

18. In this problem we will explore the impact of different log-normal shadowing parameters on outage prob ability. Consider a cellular system where the received signal power in dB has a Gaussian distribution with mean Pr dBm and standard deviation σ dB. Assume the received signal power must be above 10 dBm for acceptable performance.

(a) What is the outage probability when Pr = 15 dBm and σ = 8 dB?

(b) For σ = 4 dB, find the value of Pr required for the outage probability to be less than l% – a typical value for cellular systems.

(c) Repeat part (b) for σ = 12 dB.

(d) One proposed technique for reducing outage probability is to use macrodiversity, where a mobile unit’s signal is received by multiple base stations and then combined. This can only be done if multiple base stations are able to receive a given mobile’s signal. Explain why this might reduce outage probability.

19. Derive the formula for cell coverage percentage (2.49) by applying integration by parts to (2.47).

20. The cell coverage percentage is independent of the decorrelation distance Xc of the shadowing. Explain why. Now suppose it is known that a particular set of cell locations near the transmitter are in outage. Will the cell coverage percentage with this side information now depend on the decorrelation distance and, if so, how and why?

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21. Find the coverage area for a microcellular system where path loss follows the single-slope model (with γ = 3, d0 = 1 and K = 0 dB) and there is also log-normal shadowing with σ = 4 dB. Assume a cell radius of 100 m, a transmit power of 80 mW, and a minimum received power requirement of Pmin = −100 dBm.

22. Consider a cellular system where (a) path loss follows the single-slope model with γ = 6 and (b) there is also log-normal shadowing with σ = 8 dB. If the received power at the cell boundary due to path loss is 20 dB higher than the minimum required received power for non-outage, find the cell coverage area.

23. In microcells, path loss exponents usually range from 2 to 6 and shadowing standard deviation typically ranges from 4 to 12. Given a cellular system in which the received power due to path loss at the cell boundary equals the desired level for non-outage, find the path loss and shadowing parameters within these ranges that yield the best and worst coverage area. What is the coverage area when these parameters are in the middle of their typical ranges?



Figure 2.14: System with scattering for Problem 2-11.

24. What average power falloff with distance do you expect for the ten-ray model if G0 = G1 >> Gi for i = 2, ..., 9? Repeat for G0 = G1 = G2 >> Gi for i = 3, ..., 9. Explain your answers in both cases.

25. For the ten-ray model, assume that the transmitter and receiver are at the same height in the middle of a street of width 20 m. The transmitter-receiver separation is 500 m. Find the delay spread for this model.

26. Consider a system with a transmitter, receiver, and scatterer as shown in Figure 2.14. Assume the transmitter and receiver are both at heights ht = hr =4m and are separated by distance d, with the scatterer at distance .5d along both dimensions in a two-dimensional grid of the ground – that is, on such a grid the transmitter is located at (0, 0), the receiver at (0, d), and the scatterer at (.5d, .5d). Assume a radar cross-section of 20 dBm2, Gs = 1, and fc = 900 MHz. Find the path loss of the scattered signal for d = 1, 10, 100, and 1000 meters. Compare with the path loss at these distances if the signal is only reflected, with reflection coefficient R = −1.

27. Find the median path loss under the Hata model assuming fc = 900 MHz, ht = 20 m, hr = 5m, and d = 100 m for a large urban city, a small urban city, a suburb, and a rural area. Explain qualitatively the path loss differences for these four environments.

28. Consider a wireless link operating outdoors over 200m in the mmW freqency band. Assume the path loss follows the single-slope model with a path loss exponent γ = 2 and reference distance for antenna far field dr = 1 m. Without any additional attenuation caused by atmospheric and rain conditions, the model attenuation factor K = 1. Assume rain attenuation follows the simple approximation Krain = .95R.77 for R the rain rate in mm/hr. Assume the atmospheric absorption given in Fig. 2.11.

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(a) What is the received signal power due to path loss and oxygen absorption at carrier frequencies of 60 GHz and of 80 GHz assuming a transmit power of 1W?

(b) Consider now only the 80 GHz link. Assume a day where it is dry at 8 am, there is a heavy drizzle of 2.5mm/Hr at 12 pm and a heavy downpour of 50mm/Hr at 5 pm. What is the required transmit power at 8 am, 12 pm and 5 pm if we desire a received signal power of -50dBm at each of these time instances?

29. Using the indoor attentuation model, determine the required transmit power for a desired received power of –110 dBm for a signal transmitted over 100 m that goes through three floors with attenuation 15 dB, 10 dB, and 6 dB (respectively) as well as two double plaster-board walls. Assume a reference distance dr = 1, exponent γ = 4, and constant K = 0 dB.

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