Path Loss Exponent

Related terms:

Receiver, Coverage Area, Path Loss Model, Receiver Sensitivity, Signal Frequency, Transmission Power

View all Topics

Learn more about Path Loss Exponent

High-Level Requirements and Link Budget Analysis

Tony J. Rouphael, in RF and Digital Signal Processing for Software-Defined Radio, 2009

4.2.1 Path Loss

Path loss is intimately related to the environment where the transmitter and receiver are located. Path loss models are developed using a combination of numerical methods and empirical approximations of measured data collected in channel sounding experiments. In general, propagation path loss increases with frequency as well as distance:

(4.1)

where P_l is the average propagation path loss, d is the distance between the transmitter and receiver, n is the path loss exponent which varies between 2 for free space to 6 for obstructed in building propagation [7], and is the free space wavelength defined as the ratio of the speed of light in meters per second to the carrier frequency in Hz

(4.2)

Example 4-1: Path Loss

What is the path loss of a UWB MBOA signal transmitted at the carrier frequencies of 3432, 3960, and 4488 MHz as the range varies between 1 and 10 meters? Assume the path loss exponent for in-building line of sight to be 2.7.

First, let's compute the path loss at 4 meters for the carrier frequency of 3960 MHz using the relations in (4.1) and (4.2):(4.3)

For the complete range of 1 to 10 meters, the path loss is shown in Figure 4.6.

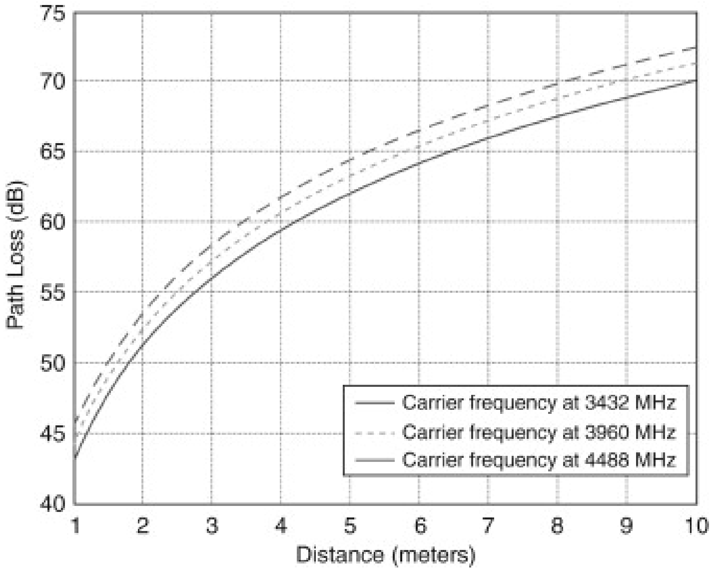

Figure 4.6. Path loss for a UWB MBOA signal with path loss exponent of 1.7 at the various carrier frequencies

> Read full chapter

Signal Parameter Estimation for the Localization Problem

David Munoz, ... Rogerio Enriquez, in Position Location Techniques and Applications, 2009

2.6.1 Log-Normal Propagation Model

The log-normal path-loss model may be considered as a generalization of the free-space Friis equation [32] where the power is allowed to decrease at a rate of $(1/d)^n$ (where d denotes distance or range), and where a random variable is added in order to account for shadowing (large–scale fading) effects. The model may be expressed as

(2.99)

where log stands for base 10 logarithm, $p_r(d)_{dB}$ is the received power at a distance d meters from the transmitter, n is the path-loss exponent that defines the rate of decay of power with respect to distance, and is a Gaussian random variable with zero mean and variance that is defined in dBs. Further, denotes the ensemble average over all possible received power values for a given reference distance denoted as d_0 meters. This reference power is usually measured a priori or calculated with the free space Friis equation. In all cases, d_0 should be as small as possible while being in the far–field region from the transmitter. In our discussion, we will consider that power measurements exclude small-scale fading since they were obtained by the averaging techniques presented in Section 2.6. The log-normal model basically states that the received power is not uniform when measured at different locations while maintaining the same distance separation between the transmitter and receiver. Note that a Gaussian random variable defined in decibels becomes a log-normal random variable when transformed into the linear domain. The log–normal Equation (2.99) is clearly a line with slope 10n when plotted versus distance values given in decibels. Typical values of path–loss exponents range between 1.5 and 5.

> Read full chapter

RF Propagation, Antennas, and Regulatory Requirements

Shahin Farahani, in ZigBee Wireless Networks and Transceivers, 2008

5.8 Range Estimation

From the concepts covered in previous sections, it is easy to come up with an equation to estimate the range associated with two wireless nodes. In Figure 5.9, if node A is a transmitter and node B is a receiver and the following values are known:

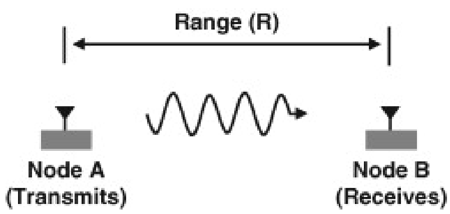

Figure 5.9. Range Estimation

- Node A transmit power (including the antenna gain, if any) = P_o(dBm)
- Node B receiver sensitivity = P_r(dBm)
- Path-loss exponent (use Table 5.1) = n

- Fade margin (see Section 5.5.3) = F_m(dB)
- Signal frequency = f(MHz)

Then the estimated range (*R*, in meters) can be calculated from the following equation:

(5.9)

For example, if node A transmits a 2450 MHz signal with 0 dBm output power and receiver sensitivity is –92 dBm in an environment with a path-loss exponent of 2.8 and a fade margin of 10 dB, the estimated range is 24 meters (79 feet). This book's companion Website has a simple calculator for range estimation.

5.8.1 Range Improvement Techniques

There are at least three ways to increase the range of wireless nodes. Equation 5.9 suggests that increasing the output power and/or improving sensitivity will extend the range. Alternatively, the range can be extended using a node-hopping (mesh networking) technique.

5.8.1.1 Range Improvement Using External PA and/or LNA

The output power of a typical IEEE 802.15.4 transmitter is normally adjustable and the range can be increased by increasing the transmitter output power. But if the maximum output power of the transmitter, which is typically around +3dBm, is not sufficient for the application, an external power amplifier (PA) can be added to boost the output power and consequently improve the range. Figure 5.10a shows a simplified diagram of how an external PA can be added in the transmit path. Transceiver ICs normally have a pin (called a T/R pin) that indicates whether the transceiver is in transmit or receive mode. This T/R pin can be used to put the external PA in the transmit path during the transmit mode of operation and remove it from the path when the transceiver is in receive mode.

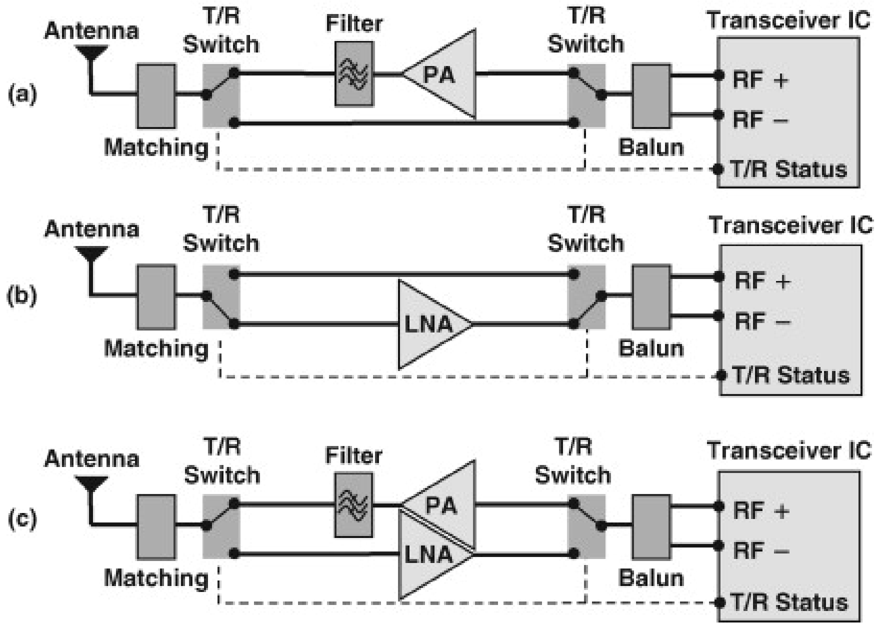

Figure 5.10. Range Extension Using External LNA and/or PA

The drawbacks of this method are additional cost of an external PA and higher current consumption of the node due to an extra PA. An extra PA can significantly reduce the battery life of a wireless node. Also, an additional PA not only amplifies the desired output signal, it also magnifies the unwanted out-of-band emissions. Therefore, it is recommended that you reexamine the conformity of the wireless node to regulatory requirements such as FCC and ETSI after an extra PA is added to a node. In most cases, adding an additional filter after the external PA to suppress the out-of-band emissions might be necessary. Two additional T/R switches in the receive path can degrade the receiver sensitivity around 1 dB to 2 dB, depending on the type of T/R switches used. In some of the existing transceiver ICs, the balun in Figure 5.10a is inside the transceiver IC package.

As an example, using Equation 5.9, by adding an external PA to node A in Figure 5.9 we increase the output power to +10 dBm from 0 dBm and consider a 2450 MHz signal, whereas node B receiver sensitivity is –92 dBm. In an environment with a path-loss exponent of 2.8 and a fade margin of 10 dB, the estimated range will increase to 54.7 m (179 feet) from 24 m (79 feet).

The concept of Signal to Noise Ratio (SNR) was discussed in chapter 4. The SNR is an indication of the signal quality and increasing SNR will improve receiver Packet Error Rate (PER). The RF and analog blocks in a receiver degrade the SNR of the received signal as they pass the signal to the receiver digital block. Noise Figure (NF) is a measure of SNR degradation caused by a block. Noise Figure of a block is the ratio of the SNR at the input of the block to the SNR at the output of the block. In an ideal case, with no SNR degradation, this ratio is equal to 1. The Noise Figure is generally provided in dB scale. If the NF is provided as a straight ratio instead of decibel, it is referred to as Noise Factor.

Another method for increasing the range is improving the sensitivity using an additional low-noise amplifier (LNA) in the receiver path (Figure 5.10.b). An LNA can improve overall NF of a receiver and consequently, improve the receiver sensitivity and range. Both the gain and the noise figure (NF) of this additional LNA impact the overall receiver sensitivity. To calculate the sensitivity improvement, you need to have the following:

- Original receiver sensitivity (without the additional LNA) = P_r(dBm)
- Original receiver NF (without the additional LNA) = NF_r(dB)
- External LNA NF = NF_{LNA}(dB)
- External LNA gain = G_{LNA}(dB)
- Total loss due to each T/R switch = $L_{T/R}$(dB)

Normally, the receiver NF is not quoted in the transceiver IC datasheet, but the manufacturer may provide the receiver NF value on request, or you can estimate the NF using the receiver sensitivity value. The improved sensitivity level of the receiver (P_{r+LNA}) can be calculated from the equation:

(5.10)

After calculating the new receiver sensitivity level, Equation 5.9 will provide an estimate of the improved range. For example, if in Figure 5.10b the original receiver sensitivity is –92 dBm with a noise figure of 17 dB, the addition of two T/R switches with total loss of 1 dB and an external LNA with gain of 15 dB and a noise figure of 1.5 dB will improve the overall sensitivity to –103 dB:

If a transmitter is emitting a 2450 MHz signal with 0 dBm output power in an environment with a path-loss exponent of 2.8 and fade margin of 10 dB, the estimated range will improve to 60.7 m (199 feet) from 24 m (79 feet).

Intuitively it is possible to use both external LNA and PA to maximize the range (Figure 5.10c). This book's companion Website has a simple calculator for range extension methods discussed in this section.

5.8.1.2 Mesh Networking to Improve Range

This is one of the important advantages of establishing a wireless network using ZigBee. In a mesh network, nodes are interconnected with other nodes so that at least two pathways connect each node to the rest of the network. As illustrated in Figure 5.11, node A uses nodes B, C, and D as routers to send the message to a faraway node E, which is outside the direct reach of node A. The details of mesh networking were covered in Chapter 4.

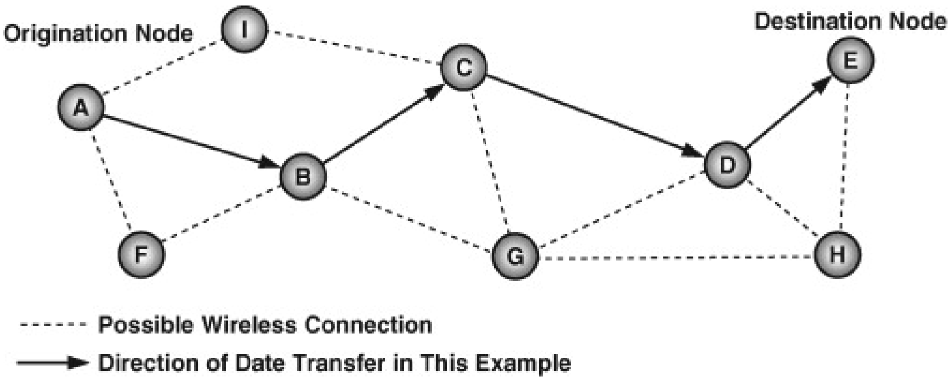

Figure 5.11. Mesh Networking Can be Used to Extend the Range

> Read full chapter

Location Estimation Methods

Shahin Farahani, in ZigBee Wireless Networks and Transceivers, 2008

7.2.1 RSSI-Based Location Estimation Using Trilateration

In an open environment with high probability of line-of-sight (LOS) and low-multipath effects, it might be possible to use a simple RSSI-based location-estimation algorithm if coarse accuracy is acceptable. Figure 7.2a shows an ideal location-estimation scenario where there are three nodes (nodes 1, 2, and 3) with known fixed locations. The fourth node is mobile, and the goal is to determine the estimated two-dimensional location of node 4. Two-dimensional (2D) means only *X* and *Y* coordinates of the node will be estimated. But the same concept can be extended to three-dimensional (3D) space as well. The location estimation in Figure 7.2a begins with node 4 transmitting a signal with a predefined output power. Assuming that all nodes in Figure 7.2a have omnidirectional antennas, each one of the fixed nodes 1–3 can estimate the distance (*r*) between its location and the location of node 4 using the following equation from Chapter 5:

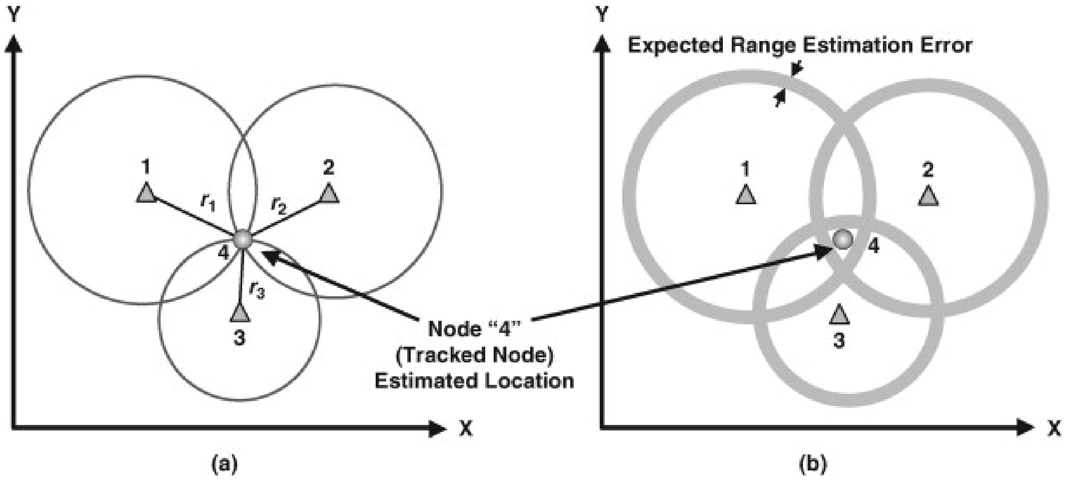

Figure 7.2. Location Estimation using Trilateration (a) Ideal Case and (b) with Range Estimation Error

(7.1)

where P_T is the transmitted power (in dBm) by node 4, P_R is the RSS at the fixed node location, f is the transmitted signal frequency in MHz, n is the path-loss exponent, and r is the distance in meters.

Node 1, for example, can estimate the distance (r_1) between its location and the location of node 4 using RSS. From the single measurement done by node 1, the only conclusion that can be made is that node 4 is located on the perimeter of a circle with radius of r_1 centered at node 1. Using the Euclidian distance, we can write the following simple equation:

or:

(7.2)

where (X_1, Y_1) and (X_4, Y_4) are coordinates for node 1 and node 4, respectively. Similar equations can be derived for node 2 coordinates (X_2, Y_2) and node 3 coordinates (X_3, Y_3). Therefore, to find the location of node 4, we need to find (X_4, Y_4) that satisfies the following equations:

(7.3)

In the ideal scenario shown in Figure 7.2a, there will be a pair of coordinates (X_4, Y_4) that satisfies this equation. This method of determining the relative location of nodes using the geometry of triangles is referred to as *trilateration*. However, in a practical implementation, due to measurement errors it might not be possible to make the right-hand side of Equation 7.3 a true zero vector for any value of (X_4, Y_4). The RSSI provided by the transceivers has limited accuracies, which directly impacts the estimated distance between the nodes. The path-loss exponent is determined experimentally and can be a source of major error. As shown in Figure 7.2b, the circles associated with each fixed node might not even have a common intercept point when the actual range estimation error is higher than the expected range estimation error.

Since it is not feasible to make the right-hand side of Equation 7.3 a true zero, we can define an error vector (*E*) instead:

(7.4)

where *abs*(.) is the absolute value function.

If the square error is defined as:

(7.5)

then the goal of location estimation becomes finding a pair of coordinates (X_4, Y_4) that minimizes the square error in Equation 7.4. This is a simple example of the classic optimization problem, where iterative or noniterative methods are used to minimize the value of an error function.

This simple RSSI-based location estimation can also be used when there are more than three fixed nodes with known locations. In this way the signal transmitted by the node with unknown location will be received by several nodes instead of only three nodes. The number of rows in Equation 7.4 is proportional to the number of fixed nodes participating in location estimation. Increasing the number of fixed nodes may improve the location-estimation accuracy in some applications. It is also possible to engage only the nearby nodes in location estimation. The RSSI value of the packet received by each anchor node indicates the distance between the nodes. If an anchor node receives a packet from the tracked node as part of the location-estimation process, the anchor node only participates in the location estimation if the RSSI of the received packet is above a certain limit. By modifying the RSSI limit, you increase or decrease the number of anchor nodes participating in the location estimation.

This simplified method was used in this subsection to describe the basic concept of location estimation using RSSI. This method requires further improvements to become a practical method of locationing. The error in RSSI measurement can result in an unacceptable level of inaccuracy in the estimated location. This method is not suitable for high multipath environments such as an office or an industrial warehouse. Alternative methods, which are more applicable to *indoor* location estimation, are discussed in Sections 7.2.3 and 7.2.4.

> Read full chapter

Aerial Platforms for Public Safety Networks and Performance Optimization

Akram Al-Hourani, ... Abbas Jamalipour, in Wireless Public Safety Networks 3, 2017

7.2.1 The nature of the air-to-ground radio channel

Radio propagation in an AtG radio channel largely differs from legacy terrestrial propagation in cellular networks. For instance, in terrestrial propagation radio models, the mean path loss is usually captured in a log-distance relation, i.e. the path loss is proportional to the logarithm of the distance, having a certain propagation exponent. Namely, a typical terrestrial path loss has the form of [AL 14a]:

where ▯ is called the path -> path-loss exponent, L_{ref} is the reference path loss and d is the distance between the transmitter and the receiver, note that the loss here is measured in dB.

The main reasoning behind the wide adoption of the log-distance model in terrestrial communications is firstly due to its mathematical simplicity, and secondly to its level of accuracy. This level of accuracy is due to the near-homogenous propagation environment at a large scale. On the contrary, radio waves in an AtG channel travel freely without obstacles for a large distance before reaching the urban layer of the man-made structures. The latter layer causes the signal to scatter and diffract, leading to an excessive amount of losses on top of the free-space path loss incurred between the aerial platform and the terrestrial user. The nature of the channel between an aerial platform and a terrestrial user is shown in Figure 7.3.

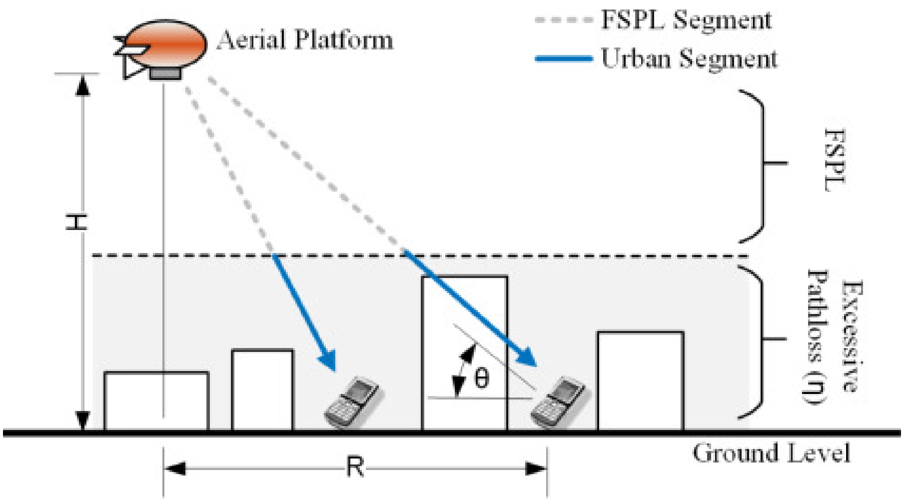

Figure 7.3. Air-to-Ground propagation between an aerial platform and a terrestrial user

Accordingly, a better way to model an AtG channel is considering the losses as composed of two parts [AL 14b]: the first part is the free-space path loss (FSPL) and the second part includes the additional losses caused by the urban environment, called the *excessive path loss* and denoted by . The AtG channel is shown in Figure 7.3, which indicates the two distinct propagation segments, namely the FSPL segment and the urban environment segment. The AtG channel loss can be expressed in dB as follows [AL 14b]:

[7.1]

where L_o represents the free-space path loss between the aerial platform and the terrestrial user, which is given by:

where f_c is the system frequency in Hz, c is the speed of light in meters per second and refers to the *propagation group*. A propagation group represents the set of receivers which share a similar statistical behavior, namely a certain statistical distribution of fading and excessive path loss. It has been empirically noticed that an AtG channel adheres to two main propagation groups that correspond to (1) good propagation conditions group, and (2) *bad propagation conditions group*. These distinct performance groups are due to shadowing caused by urban structures, for example they can be referred to as a line-of-sight (LoS) group and a non-line-of-sight (NLoS) group. The probability that a receiver belongs to a certain group depends on the elevation angle (i.e. the angle at which the aerial platform is seen from the ground receiver), we call this probability the *group occurrence probability* and we denote it by p. The excessive path loss is captured in the random variable n, where its statistics is also dependent on the propagation group.

In the upcoming sections, the reader will learn more about these concepts using practical examples and illustrations.

> Read full chapter

Wireless Communication: Concepts, Techniques, Models

Anurag Kumar, ... Joy Kuri, in Wireless Networking, 2008

A Characterization of the Power Attenuation Process

It follows from the linear model with flat fading, shown in Equation 2.8, that the received sequence, Y_k, $k \geq 1$, is also a complex valued random process. The problem for the receiver, on receiving the sequence of complex numbers Y_k, $k \geq 1$, is to carry out a *detection* of which symbols X_k, $k \geq 1$, were sent and hence which user bits were sent. This problem is particularly challenging in mobile wireless systems since the channel is randomly changing with time. The analysis and design of modulation schemes often is based on the analysis of received signal power to noise power ratios. Hence, it is important to have an effective but simple model of the channel power attenuation process, H_k.

The process $\{H_k\}$ is characterized by writing it in terms of three multiplicative components, that is,

(2.10)

Let us write the marginal terms of the stationary random processes in this expression by dropping the symbol index *k*. We will now discuss each of these terms.

The term is the path loss factor. Here, *d* is the distance between the transmitter and the receiver when the *k*-th symbol is being received, d_0 is the "far field" *reference distance* beyond which this model is applicable, and □ is the path loss exponent, which is typically in the range 2 to 5. The value of d_0 relates to the antenna dimensions and the propagation environment. For distances less than d_0, a different path loss exponent may be used, or, when d_0 is very small, we may assume no path loss.

If the attenuation is measured at various points at a distance *d* from the transmitter, then the attenuation will be found to be random, owing to variations in the terrain, and in the media through which the signal may have passed. Empirical studies have shown that this randomness is captured well if the second factor *S*, in (2.10), has the form , with □ being a Gaussian random variable with mean 0 and variance □2. This is called the *shadowing* component of the attenuation, and, since $\log_{10}$ of this term has a Gaussian (or normal) distribution, it is called *log-normal shadowing*. It is often convenient to express values of power and power ratios in the *decibel* (*dB*) unit which is obtained by taking 10 $\log_{10}$ of the value. Hence the shadowing attenuation in signal power is 10 $\log_{10} S = -$□ dB, which is zero mean Gaussian with variance □2. A typical value of □ is 8 dB. Considering two standard deviations above and below the mean, this value means that, with a high probability, shadowing can result in a variation of channel gain of times to times the mean path loss.

Shadow fading is spatially varying, and hence if there is relative movement between the transmitter and the receiver then shadow fading will vary. The correlation in the shadow fading in dB between two points separated by a distance *D* is given by , where D_0 is a parameter that depends on the terrain. Some measurements have given D_0 = 500 m for suburban terrains, and D_0 = 50 m for urban terrains. Hence if the distance is varying by a few meters per second (note that 36 Kmph = 10 meters/second) then the shadowing will vary over seconds, which means that the variations will occur over hundreds of thousands of symbols.

We now turn to the third factor, R_2, in the expression for attenuation in (2.10). Typical carrier frequencies used in mobile wireless networks are 900 MHz, 1.8 GHz (e.g., these two frequency bands are used in cellular wireless telephony systems), or 2.4 GHz (e.g., used in IEEE 802.11 wireless LAN systems). Hence, the carrier wave periods are a few picoseconds. Thus, when the transmitted signal arrives over several paths then very small differences in the path lengths (a few centimeters) can cause large differences in the phases of the carriers that are being superimposed. Thus, although these time delays may not result in ISI, the superposition of the delayed carriers results in constructive and destructive carrier interference, leading to variations in signal strength. This phenomenon is called multipath fading. This is a random attenuation that has strong autocorrelation over a time duration called the *coherence time,* T_c; that is, the attenuations at two time instants separated by more than the coherence time are weakly correlated. The coherence time is related to the Doppler frequency, f_d, which is related to the carrier frequency, f_c, the speed of movement, ν, and the speed of light, c, by . Roughly, the coherence time is the inverse of the Doppler frequency. For example, if the carrier frequency is 900 MHz, and ν = 20 meters/sec, then f_d = 60 Hz, leading to a coherence time of 10s of milliseconds. In the indoor office or home environment, the Doppler frequency could be just a few Hz (e.g., 3 Hz), with coherence times of 100 s of milliseconds. The marginal distribution of R_2 depends on whether all the signals arriving at the receiver are scattered signals, or if there is a *line-of-sight* signal as well. In the former case, assuming uniformly distributed arrival of the signal from all directions, the distribution of R_2 is exponential with mean $E(R_2)$, that is,

The distribution of the amplitude attenuation (i.e., R) is Rayleigh; hence this is also called Rayleigh fading. On the other hand if there is a line-of-sight component so that a fraction of the signal arrives directly, and the remaining signal arrives uniformly over all directions, then

where

This is called the Ricean distribution.

With this characterization of the attenuation in the received signal power we can now write the received SNR (denoted by Γ_{rcv}) in terms of the ratio of the transmitted signal power to the received noise power (denoted by Γ_{xmt}). We have

(2.11)

Then, in dB, we can write the received SNR as

(2.12)

> Read full chapter

Defending Against Identity-Based Attacks in Wireless Networks

Yingying Chen, Jie Yang, in Handbook on Securing Cyber-Physical Critical Infrastructure, 2012

8.6.2 Theoretical Analysis of the Spatial Correlation of RSS

Although affected by random noise, environmental bias, and multipath effects, the RSS measured at a set of landmarks (i.e., reference points with known locations) is closely related to the transmitter's physical location and is governed by the distance to the landmarks [43]. The RSS readings at the same physical location are similar, whereas the RSS readings at different locations in physical space are distinctive. Thus, the RSS readings present strong spatial correlation characteristics.

We define the RSS value vector as where n is the number of landmarks/access points(APs) that are monitoring the RSS of the wireless nodes and know their locations. Generally, the RSS at a landmark from a wireless node is lognormally distributed [44]:

Equation 8-21

where i is the ith landmark, represents the transmitting power of the node at the reference distance d_0, dj is the distance between the wireless node j and the ith landmark, and is the path loss exponent, Xi is the shadow fading which follows zero mean Gaussian distribution with standard deviation [44, 45]. For simplicity, we assume the wireless nodes have the same transmission power and are static in the network. Research work [41] discussed the issue of using different transmission power levels and research work [53] proposed to detect identity-based spoofing attacks in mobile wireless environments. Given two wireless nodes in the physical space, the RSS distance between two nodes in signal space at the ith landmark is given by

Equation 8-22

where ΔX follows zero mean with standard deviation. According to Eq. (8-22), when the two wireless nodes are at the same location (i.e.,), the RSS distance in signal space at the ith landmark follows a normal distribution with zero mean and standard deviation. Whereas the distance follows a normal distribution with mean and standard deviation if these two nodes are at different locations.

The square of RSS distance in *n*-dimensional signal space (i.e., at *n* landmarks) is then determined by

Equation 8-23

where with is the RSS distance at *i*th landmark and is given by Eq. (8-22).

Based on Eqs (8-22) and (8-23), we know that, when these two wireless nodes are at the same location, the distance in *n* dimension signal space follows a *central chi-square distribution* with *n* degree of freedom [46]. When two wireless nodes are at the same location, the probability density functions (PDF) of the random variable , which is the square distance in *n*-dimensional signal space, can be represented as:

Equation 8-24

where $x \geq 0$ and denotes the Gamma function, which has closed-form values at the half-integers.

However, when these two wireless nodes are at different locations, becomes a *noncentral chi-square distribution* with *n* degree of freedom and a noncentrality parameter , where

Equation 8-25

and *dij*, with $i = 1,2,\ldots n.$, $j = 1,2.$, is the distance from *j*th wireless nodes to the *i*th landmark. The PDF of the random variable $X = \Delta D_2$ when two wireless nodes are at different locations can be represented as:

Equation 8-26

where $I(z)$ is a modified Bessel function of the first kind [46].

Given the threshold , the probability that we can determine the two nodes are at different locations in a two-dimensional physical space with *n* landmarks (i.e., detection rate *DR*) is given by

Equation 8-27

and the corresponding false positive rate is

Equation 8-28

where is the CDF of the random variable *X*.

From Eqs (8-27) and (8-28), for a specified detection rate *DR*, the threshold of test can be obtained as

Equation 8-29

and the false positive rate can be represented in terms of the detection rate

Equation 8-30

From Eq. (8-27), we can see that the detection rate *DR* increases with □, which can be represented by the distance between two wireless nodes together with the landmarks. Moreover, for a specified detection rate *DR*, Eq. (8-30) shows that the false positive rate *FPR* increases with the standard deviation of shadowing □.

We next study the detection power of the proposed approach by using the RSS-based spatial correlation. Figure 8-4 presents the numerical results of receiver operating characteristic (ROC) curves based on Eqs (8-27) and (8-28) when randomly placing two wireless devices in a 100 by 100 square-foot area. There are four landmarks deployed at the four corners of the square area. The physical distance between two wireless devices is 16, 20, and 25 ft, respectively. The path loss exponent □ is set to 2.5 and the standard deviation of shadowing is 2 dB. From the figure, we observed that the ROC curves shift to the upper left when increasing the distance between two devices. This indicates that the farther away the two nodes are separated, the better detection performance the proposed method can achieve. This is because the detection performance is proportional to the noncentrality parameter □, which is represented by the distance between two wireless nodes together with the landmarks.

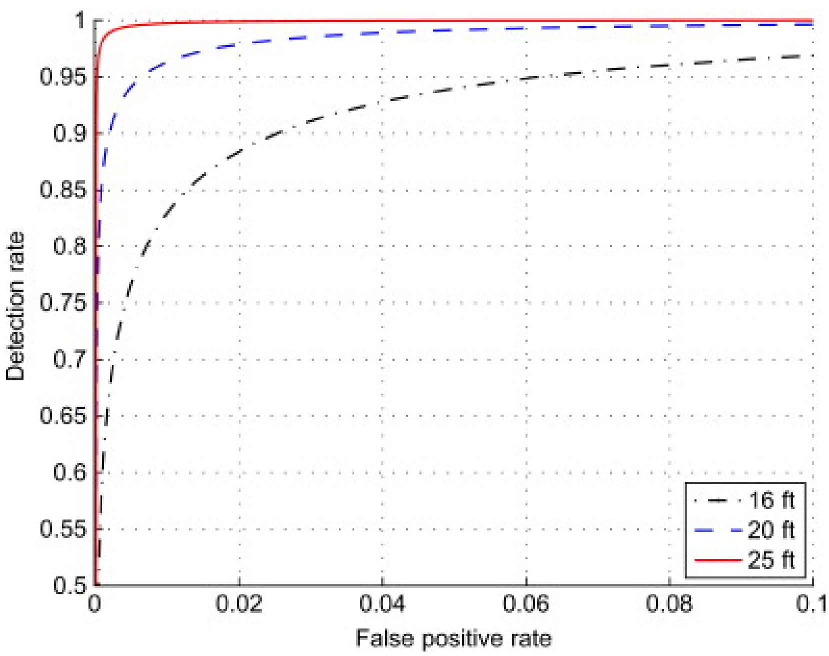

Figure 8-4. The ROC curves when the distance between two wireless devices is 16, 20, and 25 ft, respectively. The standard deviation of shadowing is 2 dB. The path loss exponent is 2.5.

We further investigate the detection performance of the proposed approach under RSS variations. In this study, we fixed the distance between two wireless devices as 25 ft. The obtained ROC curves when the standard deviation of shadowing is set to 2 dB, 3 dB, and 4 dB, respectively, are shown in Figure 8-5. From the figure, it can be seen that we can obtain better detection performance with lower standard deviation of shadowing . A larger standard deviation of shadowing causes the two distributions, that is, noncentral chi-square and central chi-square to get closer to one anotherThe above analysis provides the theoretical support of using the spatial correlation in RSS inherited from wireless nodes to perform attack detection. It also showed that the RSS readings from a wireless node over time fluctuate under different. Consequently, the smaller standard deviation of shadowing results in a better detection performance.

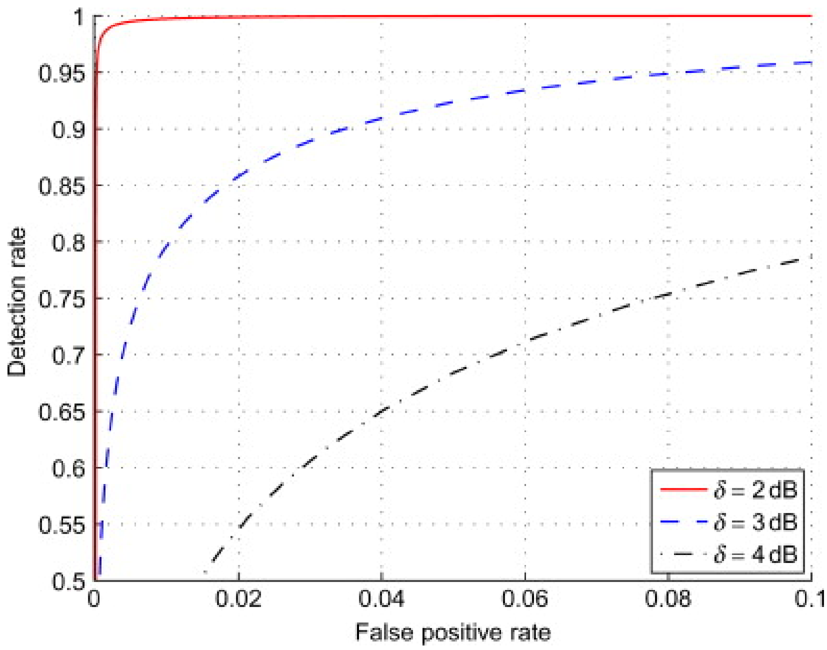

Figure 8-5. The ROC curves when the standard deviation of shadowing is 2 dB, 3 dB, and 4 dB, respectively. The distance between two devices is 25 ft.

> Read full chapter

Cellular FDM-TDMA

Anurag Kumar, ... Joy Kuri, in Wireless Networking, 2008

4.6.1 Analysis of Signal Strength Based Handovers

We consider an MS located on the line joining two BSs, BS 0 and BS 1, as shown in Figure 4.17. Let $S_i(x)$ = Received signal power from BS *i*, *i* {0, 1}, when the *MS* is at the distance *x* from BS 0. Then, recalling the path loss and shadowing model from Section 2.1.4, we have

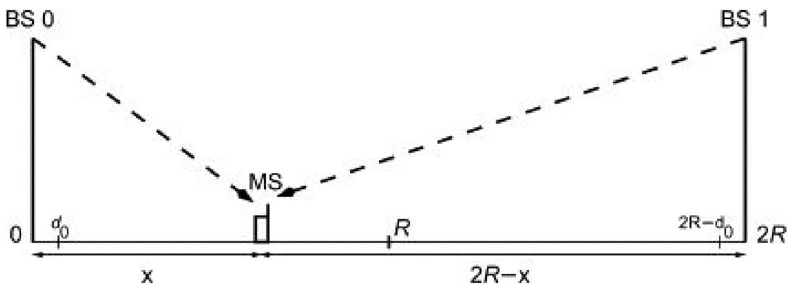

Figure 4.17. Handover: An MS located on the line joining two neighboring BSs that are at the distance *2R*. The MS is located at distance *x* from BS 0. The MS provides signal strength measurements from each of the BSs.

where □ is the path loss exponent, and the shadow fading, □*i*, *i* □ {0, 1}, is normally distributed with mean 0, and variance □2. Also,

Let us assume that $S_0(d_0) = S_1(d_0)$. Then, for $d_0 \le x \le 2R - d_0$,

where □1–□0 is normally distributed with 0 mean and variance 2□2. In Figure 4.18, we show the variation of $[S_0(x) – S_1(x)]_{dB}$ as the MS moves from BS 0 to BS 1. The solid curve shows the mean 10□ log . The two dashed curves above and below the solid curve represent the variability due to shadowing, and can be viewed as the bounds within which $[S_0(x) – S_1(x)]_{dB}$ stays, with a high probability. The half-width of the curved strip defined by the two dashed curves is proportional to .

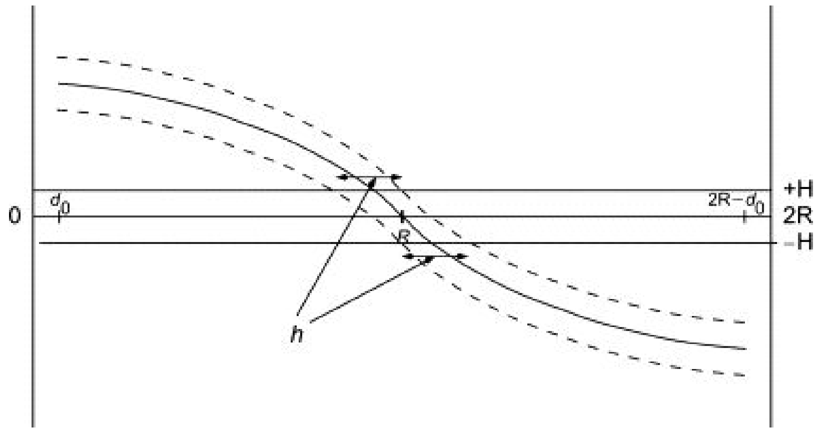

Figure 4.18. Handover: The difference in signal strengths, $S_0(x) – S_1(x)$ (in dB) at an MS that is at position *x* on the line joining BS 0 and BS 1. For an explanation of *h*, see the text.

Suppose the MS is being served by BS 0. A simple handover approach is to hand over the MS to BS 1 when $[S_0(x) – S_1(x)]_{dB} < –H$, where *H* is a design parameter; that is, handover occurs when the signal strength from BS 0 is sufficiently lower than the signal strength from BS 1. For the *H* shown in Figure 4.18, and with the preceding interpretation of the dashed curves, we can see that that handover will occur with a positive probability when the MS is at a distance greater than *R* from BS 0 (notice that the lower dashed curve falls below the horizontal line for $-H$, when $x > R$). There are two issues here:

a. If the coverage of either cell extends only up to a distance R, then once the MS is beyond R, the handover should occur with a high probability.
b. With this design, if the MS is moving about in the region around the middle of the line joining the two BSs, then it will be repeatedly handed over between the two BSs, thus increasing the chance of the call being dropped, and also increasing the load on the call management processors.

These two issues can be addressed by extending the coverage of each BS beyond R, to an additional distance, say, h. Suppose h is chosen so that

or

where is the half width of the dashed strip, and a is chosen from the standard normal tables so that the tail probability of the random variable $\square_1 - \square_0$) beyond is small. This choice of h is shown in Figure 4.18, since at $x = R+h$ the upper dashed curve falls below $-H$. Now, when deciding to hand over from BS 0 to BS 1, we check if both of the following tests are true:

for a suitably chosen $S_{threshold}$. Both these tests will succeed beyond $R+h$ with a high probability, and, thus, the handover will take place with a high probability. Further, the reverse handover will take place with a very small probability. Thus, this handover strategy has a *hysteresis* built into it.

Although this design solves the problem of repeated handovers from one cell to the other, the extension of the cell coverage into the neighboring cell impacts the earlier SIR analysis. Let

so that

Thus, in the cochannel interference calculations, we now need to use

It follows that a larger value will need to be used for a given SIR constraint, thus requiring a larger value of N_{reuse}, and lowering the spectrum efficiency. It is thus important to design handover schemes that can reduce the cell expansion factor b.

> Read full chapter

Mesh Networks: Fundamental Limits

Anurag Kumar, ... Joy Kuri, in Wireless Networking, 2008

9.1.1 Random Graph Models for Wireless Networks

In most wireless mesh networks, the nodes use omnidirectional antennas and multiple access protocols to access the channel. Hence, the channel becomes a broadcast channel. Transmission energy from a node reaches all nodes in the network. The received power at these other nodes will depend on the distance and the characteristics of the radio channel between the transmitter and the receiver. If node i is located at x_i and is transmitting with power P_i, then the SINR at a receiver at location x_j is given by

(9.1)

where $L(x, y)$ is the path loss function when the transmitter is at x and the receiver is at y, N_0 is the thermal noise spectral density at the receiver, W is the channel bandwidth, and □, 0 ≤ □ ≤ 1, is the orthogonality factor between the transmissions. If the transmissions are all perfectly orthogonal, or if no other node is transmitting when Node i is transmitting, then □ = 0. Typically, $L(x, y)$ takes the form $L(d)$ where d is the distance between the points x and y. The standard model for path loss is the far field model where $L(x, y) = L(||x - y||)^{-□}$, $||x - y||$ is the Euclidean distance between x and y, and □ > 0 is called the path loss exponent. See Section 2.1.4 for a more detailed discussion on the signal attenuation and delay phenomena on the path from the transmitter to the receiver.

To see how to construct the network graph from the path loss model, assume that all nodes transmit with the same power, say P, and that the minimum SNR required at a receiver is □. Define

When no other node in the network is transmitting, all nodes at a distance less than r_1 from the transmitter, say Node 0, can decode the transmission with acceptable error probability; the decode region is a circle of radius r_1 centered at the location of Node 0. r_1 is called the *SNR-cutoff*.

Exercise 9.1

Obtain the SNR-cutoff for $L(d) = d^{-□}$.

Now consider the situation when Node 0 is transmitting. When there are other nodes transmitting in the same slot as Node 0, the extent of the decode region of Node 0 depends on the location of these other transmitters. This is because the signal from them will be the interference at the receivers of Node 0, and this reduces their SINR. This is evident from (9.1). We will first simplify the effect of this interference and assume that a transmission from node i, located at x_i, can be decoded at node j, located at x_j, if

(9.2)

$||\cdot||$ denotes the Euclidean distance. r is called the transmission range or the *cutoff*, and essentially captures the effect of interferences. The network graph $G = (V, E)$ is constructed as follows. The vertex set corresponds to the n nodes in the network. The edge set in the network graph, representing the links in the network, is given by

Figure 9.1 illustrates the edges obtained using (9.2) at two sample nodes in a network. Note that (9.2) only fixes the links in the network. The set of links that can simultaneously be active can now be derived from graph-based constraints like those described in Chapter 7.

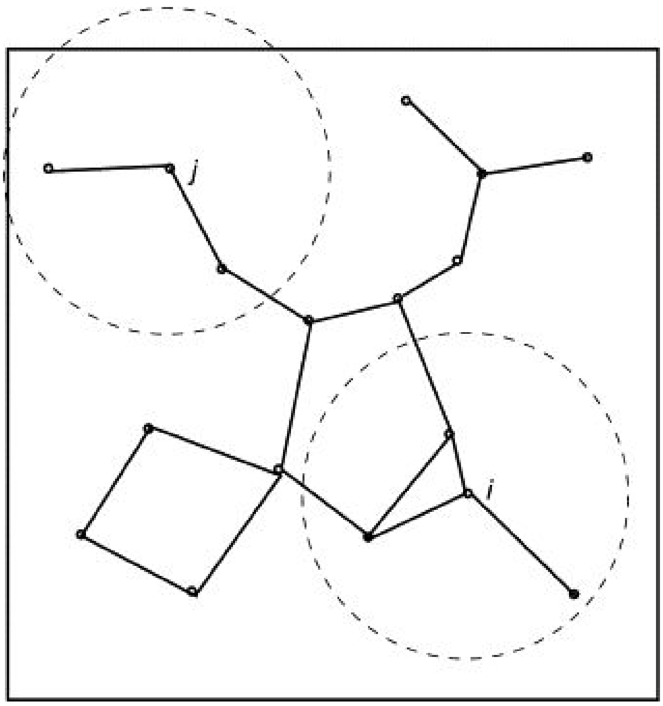

Figure 9.1. A sample realization of a random geometric graph. The cutoff region around nodes i and j is shown as dashed circles.

When the node locations are random, let X_i denote the random location of node i and let $\mathrm{X}n{:} = (X_i, i$ □{1, …, n}) denote the set of node locations in the n-node network. Since X_n is random, the network graph is a random graph. Let $G_n(r_n)$ denote a realization of a network graph obtained when there are n nodes in the network and the cutoff i s r_n. $G_n(r_n)$ is a graph-valued random variable.

Random graphs have been studied for more than 50 years, primarily as Erdös-Renyi random graphs. In this classical random graph model, in an n-node graph, edge (i, j), $1 \le i < j \le n$, occurs with probability p_n, $0 < p_n < 1$, independently of all the other edges. Observe that in the graph $G_n(r_n)$, the edges are not independent. To see this, observe that if (i, j) and (j, k) are edges in G, then we have more information on the relative locations of node pair (i, k). Therefore, we cannot say that the existence of edge (i, k) is independent of the existence of edges (i, j) and (j, k). Random graphs of the type $G_n(r_n)$ are called random geometric graphs. From the method used in determining the edges in the network, this is also called the Boolean model.

A network graph that captures the effect of interferences in more detail can also be constructed. Assume that signals transmitted by the different nodes are orthogonal, albeit with some imperfections, like in CDMA networks. Let each node randomly choose a color represented by an integer in [1, T]. Divide time into frames of T slots each. In slot t of each frame, $1 \leq t \leq T$, nodes that chose color t will transmit. Let S_t denote the set of nodes that transmit in slot t. Thus each node gets a chance to transmit once in every T slot. We will assume that even if a node, say Node j, is transmitting in a slot, it can decode a signal transmitted by another node, say Node i, if the SINR threshold is met. We will also assume that there is sufficient orthogonality between the transmissions so that Node j can decode all the simultaneous transmissions for which the SINR is above the threshold. Now consider Node i, transmitting with power P_i and a possible receiver Node j. Let SINRi,j be the SINR at Node j for the transmission from Node i. To obtain SINRi,j, observe that the received power from Node i is the signal and the received power from all other transmitters in the slot is the interference. Thus

Let Et denote the set of directed edges for slot t, those along which packets could be exchanged in slot t. E_t is obtained as follows:

Let E□T: = □Tt=1 where E□T is the set of all directed edges along which communication could take place in at least one of the T slots of a frame. Let E_T be the set of bidirectional edges in E□T,

ET is the set of undirected edges and the network graph, $G_T = (N, E_T)$, is called a signal-to-interference-ratio graph (STIRG). It is easier to study STIRGs by assuming that the nodes are distributed according to a Poisson spatial process in □2. There are many parameters that define the STIRG—the node density □, the transmission power P, the receiver noise power WN_0, the SINR threshold □, the orthogonality factor □, and the path loss function $L(\cdot)$. We will study the properties of STIRGs as a function of □ and □.

The properties of random graphs that are of interest to us will essentially be events in a probability space. Many a time, we will be analyzing the asymptotic behavior of the properties of the network graph. In this we will be using the order notation (see Section A.3 in Appendix A) with some modifications.

Consider a sequence of random experiments indexed by n. Let X be an event of interest and let $p_N(X)$ be the probability that X occurs in the n-th experiment of the sequence. If $p_n(X) \to 1$ as $n \to \infty$, we say that X occurs with high probability (w.h.p.). This means that for any □ > 0, we can find an $n^*(□)$ such that $\Pr(X_n) > 1 - □$ for all $n \geq n^*(□)$. We now apply this notion to the order notation. Let $X(n)$ be a random sequence. For example, $X(n)$ could be the random variable representing the execution time of a protocol in a random network; the protocol itself could be a randomized protocol. If

for some positive constant a, then $X(n) = O(g(n))$ w.h.p. and we write $X(n) = \tilde{O}(g(n))$. Similarly, we say if $X(n) = \Omega(g(n))$ w.h.p. And $X(n) = \tilde{o}(g(n))$ implies that $X(n) = o(g(n))$ w.h.p.

> Read full chapter