3

Noise- and Interference-Limited Systems

3.1 Introduction

This chapter explains the principles of link budgets, and the planning of wireless systems with one or multiple users. In Section 3.2, we set up link budgets for noise-limited systems and compute the minimum transmit power (or maximum range) that can be achieved in the absence of interference. Such computations give a first insight into the basic capabilities of wireless systems and also have practical applications. For example, Wireless Local Area Networks (WLANs) and cordless phones often operate in a noise-limited mode, if no other Base Station (BS) is in the vicinity. Even cellular systems sometimes operate in that mode if the user density is low (this happens, e.g., during the build-up phase of a network).

In Section 3.3, we discuss interference-limited systems. As we described in the first two chapters, the unregulated use of spectrum leads to interference that cannot be controlled by the user. When the spectrum is regulated, the network operator can determine the location of BSs, and thus impact the Signal-to-Interference Ratio (SIR). For either case, it is important to set up the link budgets that take the presence of interference into account; Section 3.3 describes these link budgets. In Chapter 17, we then see how these calculations are related to the cellular principle and the reuse of frequencies in different cells.

3.2 Noise-Limited Systems

Wireless systems are required to provide a certain minimum transmission quality (see Section 1.3). This transmission quality in turn requires a minimum *Signal-to-Noise Ratio* (SNR) at the receiver (RX). Consider now a situation where only a single BS transmits, and a Mobile Station (MS) receives; thus, the performance of the system is determined only by the strength of the (useful) signal and the noise. As the MS moves further away from the BS, the received signal power decreases, and at a certain distance, the SNR does not achieve the required threshold for reliable communications. Therefore, the range of the system is noise limited; equivalently, we can call it *signal power limited*. Depending on the interpretation, it is too much noise or too little signal power that leads to bad link quality.

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Let us assume for the moment that the received power decreases with d^2, the square of the distance between BS and MS. More precisely, let the received power P_{RX} be

$$P_{\mathrm{RX}} = P_{\mathrm{TX}} G_{\mathrm{RX}} G_{\mathrm{TX}} \left(\frac{\lambda}{4\pi d} \right)^2 \tag{3.1}$$

where G_{RX} and G_{TX} are the gains of the receive and transmit antennas, respectively,[1] λ is the wavelength, and P_{TX} is the transmit power (see Chapter 4 for a derivation of this equation and for more details).

The noise that disturbs the signal can consist of several components, as follows:

1. *Thermal noise*: The power spectral density of thermal noise depends on the environmental temperature T_{e} that the antenna "sees." The temperature of the Earth is around 300 K, while the temperature of the (cold) sky is approximately $T_{\mathrm{e}} \approx 4\,\mathrm{K}$ (the temperature in the direction of the Sun is of course much higher). As a first approximation, it is usually assumed that the environmental temperature is isotropically 300 K. Noise power spectral density is then

   $$N_0 = k_{\mathrm{B}} T_{\mathrm{e}} \tag{3.2}$$

   where k_{B} is Boltzmann's constant, $k_{\mathrm{B}} = 1.38 \cdot 10^{-23}$ J/K, and the noise power is

   $$P_{\mathrm{n}} = N_0 B \tag{3.3}$$

   where B is RX bandwidth (in units of Hz). It is common to write Eq. (3.2) using logarithmic units (power P expressed in units of dBm is $10 \log_{10} (P/1\,\mathrm{mW})$):

   $$N_0 = -174\,\mathrm{dBm/Hz} \tag{3.4}$$

   This means that the noise power contained in a 1-Hz bandwidth is -174 dBm. The noise power contained in bandwidth B is

   $$-174 + 10 \log_{10}(B)\,\mathrm{dBm} \tag{3.5}$$

   The logarithm of bandwidth B, specifically $10 \log_{10}(B)$, has the units dBHz.
2. *Man-made noise*: We can distinguish two types of man-made noise:
   (a) *Spurious emissions*: Many electrical appliances as well as radio transmitters (TXs) designed for other frequency bands have spurious emissions over a large bandwidth that includes the frequency range in which wireless communications systems operate. For urban outdoor environments, car ignitions and other impulse sources are especially significant sources of noise. In contrast to thermal noise, the noise created by impulse sources decreases with frequency (see Figure 3.1). At 150 MHz, it can be 20 dB stronger than thermal noise; at 900 MHz, it is typically 10 dB stronger. At Universal Mobile Telecommunications System (UMTS) frequencies, Neubauer et al. [2001] measured 5-dB noise enhancement by man-made noise in urban environments and about 1 dB in rural environments. Note that frequency regulators in most countries impose limits on "spurious" or "out-of-band" emissions for *all* electrical devices. Furthermore, for communications operating in licensed bands, such spurious emissions are the only source of man-made noise. It lies in the nature of the license (for which the license holder usually has paid) that no other intentional emitters are allowed to operate in this band. In contrast to thermal noise, man-made noise is not necessarily Gaussian distributed. However, as a matter of convenience, most system-planning tools, as well as theoretical designs, assume *Gaussianity* anyway.

[1] Roughly speaking, "receive antenna gain" is a measure of how much more power we can receive (from a certain direction) by using a specific antenna, compared with the use of an isotropic antenna; the definition for transmit antennas is similar. See Chapter 9 and/or Stutzman and Thiele [1997] for details.

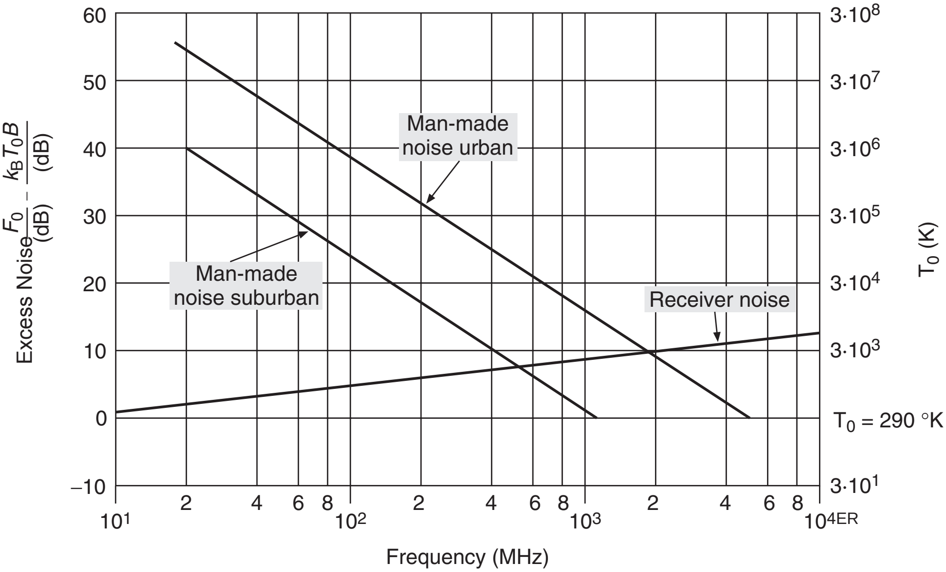

Figure 3.1 Noise as a function of frequency.

(b) *Other intentional emission sources*: Several wireless communications systems operate in unlicensed bands. In these bands, everybody is allowed to operate (emit electromagnetic radiation) as long as certain restrictions with respect to transmit power, etc. are fulfilled. The most important of these bands is the 2.45-GHz Industrial, Scientific, and Medical (ISM) band. The amount of interference in these bands can be considerable.

3. *Receiver noise*: The amplifiers and mixers in the RX are noisy, and thus increase the total noise power. This effect is described by the noise figure F, which is defined as the SNR at the RX input (typically after downconversion to baseband) divided by the SNR at the RX output. As the amplifiers have gain, noise added in the later stages does not have as much of an impact as noise added in the first stage of the RX. Mathematically, the total noise figure F_{eq} of a cascade of components is

$$F_{\mathrm{eq}} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \cdots \tag{3.6}$$

where F_i and G_i are noise figures and noise gains of the individual stages in absolute units (not in decibels (dB)). Note that for this equation, passive components, like attenuators with gain $m < 1$, can be interpreted as *either* having a noise figure of $F = 1/m$ and unit gain of $G = 1$, *or* unit noise figure $F = 1$, and gain $G = m$.

For a digital system, the transmission quality is often described in terms of the *Bit Error Rate* (BER) probability. Depending on the modulation scheme, coding, and a range of other factors (discussed in Part III of this book), there is a relationship between SNR and BER for each digital communications systems. A minimum transmission quality can thus be linked to the minimum SNR, SNR_{min}, by this mapping (see Figure 3.2). Thus, the planning methods of all analog and digital links in noise-limited environments are the same; the goal is to determine the minimum

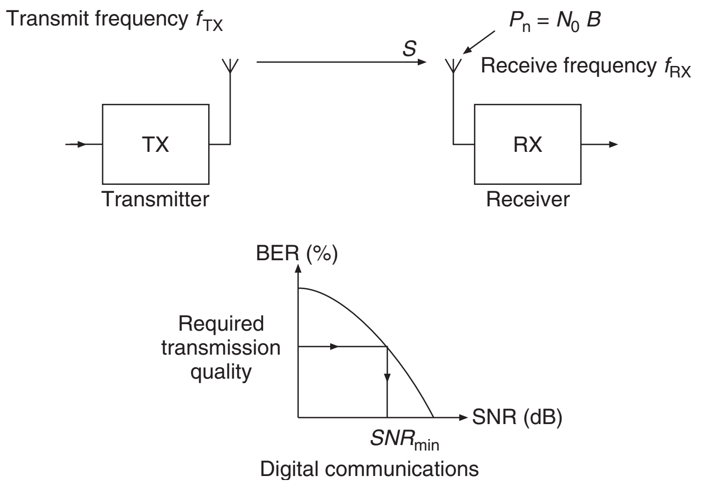

Figure 3.2 Noise-limited systems.
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signal power P_S:

$$P_S = SNR_{min} + P_n \tag{3.7}$$

where all quantities are in dB. However, note that the actual *values* will be different for different systems.

3.2.1 Link Budget

A link budget is the clearest and most intuitive way of computing the required TX power. It tabulates all equations that connect the TX power to the received SNR. As most factors influencing the SNR enter in a multiplicative way, it is convenient to write all the equations in a logarithmic form – specifically, in dB. It has to be noted, however, that the link budget gives only an approximation (often a worst case estimate) for the total SNR, because some interactions between different effects are not taken into account.

Before showing some examples, the following points should be stressed:

- Chapters 4 and 7 provide extensive discussions of path loss, i.e., the attenuation due to propagation effects, between TX and RX. For the purpose of this chapter, we use a simple model, the so-called "breakpoint" model. For distances $d < d_{break}$, the received power is proportional to d^{-2}, according to Eq. (3.1). Beyond that point, the power is proportional to d^{-n}, where n typically lies between 3.5 and 4.5. The received power is thus

$$P_{RX}(d) = P_{RX}(d_{break}) \left(\frac{d}{d_{break}} \right)^{-n} \text{ for } d > d_{break} \tag{3.8}$$

- Wireless systems, especially mobile systems, suffer from temporal and spatial variations of the transmission channel (*fading*) (see Section 2.1). In other words, even if the distance is approximately constant, the received power can change significantly with small movements of the

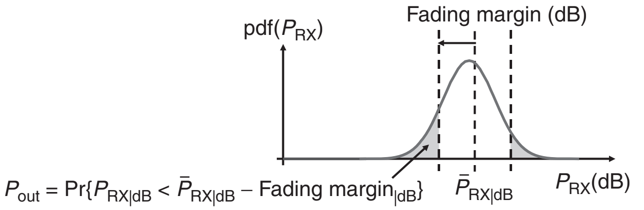

Figure 3.3 Fading margin to guarantee a certain outage probability.

TX and/or RX. The power computed from Eq. (3.8) is only a *mean* value; the ratio of the transmit power to this mean received power is also known as the *path loss* (inverse of the path gain).

If the mean received power is used as the basis for the link budget, then the transmission quality will be above the threshold only in approximately 50% of the times and locations.[2] This is completely unacceptable quality of service. Therefore, we have to add a *fading margin*, which makes sure that the minimum received power is exceeded in at least, e.g., 90% of all cases (see Figure 3.3). The value of the fading margin depends on the amplitude statistics of the fading and is discussed in more detail in Chapter 5.

- Uplink (MS to BS) and downlink (BS to MS) are reciprocal, in the sense that the voltage and currents at the antenna ports are reciprocal (as long as uplink and downlink use the same carrier frequency). However, the noise figures of BSs and MSs are typically quite different. As MSs have to be produced in quantity, it is desirable to use low-cost components, which typically have higher noise figures. Furthermore, battery lifetime considerations dictate that BSs can emit more power than MSs. Finally, BSs and MSs differ with respect to antenna diversity, how close they are to interferers, etc. Thus, the link budgets of uplinks and downlinks are different.

Example 3.1 *Link budget*

Consider the downlink of a GSM system (see also Chapter 24). The carrier frequency is 950 MHz and the RX sensitivity is (according to GSM specifications) −102 dBm. The output power of the TX amplifier is 30 W. The antenna gain of the TX antenna is 10 dB and the aggregate attenuation of connectors, combiners, etc. is 5 dB. The fading margin is 12 dB and the breakpoint d_{break} is at a distance of 100 m. What distance can be covered?

TX side:				
	TX power	P_{TX}	30 W	45 dBm
	Antenna gain	G_{TX}	10	10 dB
	Losses (combiner, connector, etc.)	L_f		−5 dB
	EIRP (Equivalent Isotropically Radiated Power)			50 dBm
RX side:				
	RX sensitivity	P_{min}		−102 dBm
	Fading margin			12 dB
	Minimum RX power (mean)			−90 dBm
	Admissible path loss (difference EIRP and min. RX power)			140 dB
	Path loss at $d_{break} = 100$ m	$[\lambda/(4\pi d)]^2$		72 dB
	Path loss beyond breakpoint	$\propto d^{-n}$		68 dB

[2] It would lie above the threshold in exactly 50% of the cases if Eq. (3.8) represented the *median* power.

Depending on the path loss exponent,

$$n = 1.5 \ldots 2.5 \text{ (line-of-sight)}^3$$

$$n = 3.5 \ldots 4.5 \text{ (non-line-of-sight)}$$

we obtain the coverage distance,

$$d_{\text{cov}} = 100 \cdot 10^{68}/^{(10n)}\text{m} \tag{3.9}$$

If, e.g., $n = 3.5$, then the coverage distance is 8.8 km.

This example was particularly easy, because RX sensitivity was prescribed by the system specifications. If it is not available, the computations at the RX become more complicated, as shown in the next example.

Example 3.2 *Link budget*

Consider a mobile radio system at 900-MHz carrier frequency, and with 25-kHz bandwidth, that is affected only by thermal noise (temperature of the environment $T_e = 300\,\text{K}$). Antenna gains at the TX and RX sides are 8 dB and −2 dB,[4] respectively. Losses in cables, combiners, etc. at the TX are 2 dB. The noise figure of the RX is 7 dB and the 3-dB bandwidth of the signal is 25 kHz. The required operating SNR is 18 dB and the desired range of coverage is 2 km. The breakpoint is at 10-m distance; beyond that point, the path loss exponent is 3.8, and the fading margin is 10 dB. What is the minimum TX power?

The way this problem is formulated makes working our way backward from the RX to the TX advantageous.

Noise spectral density	$k_B T_e$	−174 dBm/Hz
Bandwidth		44 dBHz
⋮		
Thermal noise power at the RX		−130 dBm
RX excess noise		7 dB
Required SNR		18 dB
⋮		
Required RX power		−105 dBm
Path loss from 10 m to 2-km distance	$(200^{3.8})$	87 dB
Path loss from TX to breakpoint at 10 m	$[\lambda/(4\pi d)]^2$	52 dB
Antenna gain at the MS G_{RX}	(2-dB loss)	−(−2) dB
Fading margin		10 dB
Required EIRP		46 dBm

[3] Note that the Line-Of-Sight (LOS) cannot exist beyond a certain distance even in environments that have no buildings or hills. The curvature of the earth cuts off the LOS at a distance that depends on the heights of the BS and the MS.

[4] In most link budgets, the antenna gain for the MS is assumed to be 0 dB. However, recent measurements have shown that absorption and reflection by the head and body of the user reduce the antenna gain, leading to losses up to 10 dB. This is discussed in more detail in Chapter 9.

TX antenna gain G_{TX}	(8-dB gain)	-8 dB
Losses in cables, combiners, etc. at TX	L_f	2 dB
Required TX power (amplifier output)		40 dBm

The required TX power is thus 40 dBm, or 10 W. The link budget is also represented in Figure 3.4.

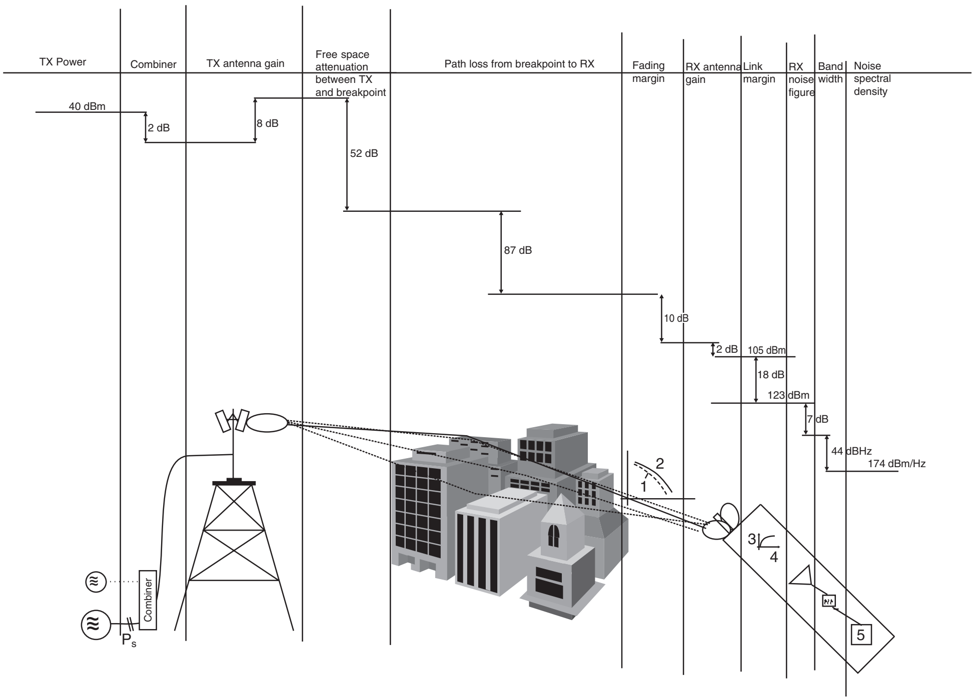

Figure 3.4 Link budget of Example 3.2. 1 = 10% decile; 2 = median; 3 = MOS; 4 = SNR; 5 = detector.

3.3 Interference-Limited Systems

Consider now the case that the interference is so strong that it completely dominates the performance, so that the noise can be neglected. Let a BS cover an area (cell) that is approximately described by a circle with radius R and center at the location of the BS. Furthermore, there is an interfering TX at distance D from the "desired" BS, which operates at the same frequency, and with the same transmit power. How large does D have to be in order to guarantee satisfactory transmission quality 90% of the time, assuming that the MS is at the cell boundary (worst case)? The computations follow the link budget computations of the previous section. As a first approximation, we treat the interference as Gaussian. This allows us to treat the interference as equivalent noise, and the minimum SIR, $SIR_{\min}$, takes on the same values as $SNR_{\min}$ in the noise-limited case.

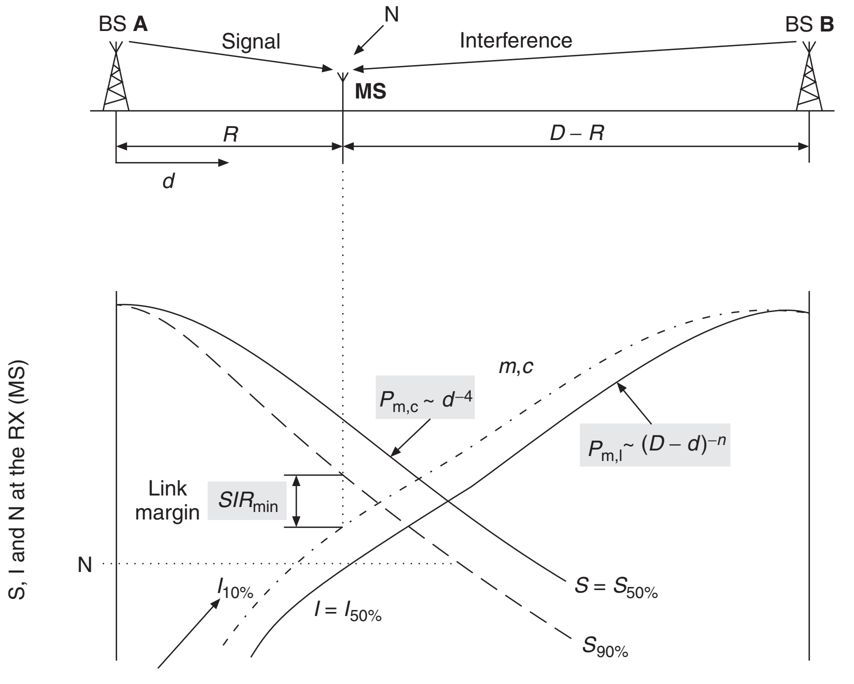

Figure 3.5 Relationship between cell radius and reuse distance. Solid lines: median values. Dashed lines: 90% decile of the desired signal. Dash–dotted lines: 10%-decile of the interfering signal.

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One difference between interference and noise lies in the fact that interference suffers from fading, while the noise power is typically constant (averaged over a short time interval). For determination of the fading margin, we thus have to account for the fact that (i) the desired signal is weaker than its median value during 50% of the time and (ii) the interfering signal is stronger than its median value 50% of the time. Mathematically speaking, the cumulative distribution function of the SIR is the probability that the ratio of two random variables is larger than a certain value in x% of all cases (where x is the percentage of locations in which transmission quality is satisfactory), see Chapter 5. As a first approximation, we can add the fading margin for the desired signal (i.e., the additional power we have to transmit to make sure that the desired signal level exceeds a certain value, x%, of the time, instead of 50%) and the fading margin of the interference –i.e., the power *reduction* to make sure that the interference exceeds a certain value only $(100 - x)$% of the time, instead of 50% of the time (see Figure 3.5). This results in an overestimation of the true fading margin. Therefore, if we use that value in system planning, we are on the safe side.

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