1

Introduction

This chapter offers basic and high-level introductions to terminology, definitions, and concepts concerning jitter, noise, signal integrity, bit error rate, and working mechanisms for communication link systems. Sources and root causes of jitter, noise, and signal integrity then are discussed, followed by statistical and system views on jitter, noise, and signal integrity. Then we give a historical overview of the evolution of and advancement path for jitter, noise, and signal integrity. This chapter ends by discussing this book's organization and flow.

1.1 JITTER, NOISE, AND COMMUNICATION SYSTEM BASICS

The essence of communication is about transmitting and receiving a signal through a medium or channel. An early mathematical model for communication may be tracked back to Claude Shannon's famous 1948 paper.[1] Depending on what kind of medium is used to transmit and receive a signal, communication systems are grouped into three basic categories: fiber, copper, and wireless (or free space) (see Figure 1.1). The bandwidths typically are a few THz for fiber and a few GHZ for copper media. Considering the constraints of bandwidth, attenuation, and cost, fiber-based communication is often used for long-distance (> 1 km), high-data-rate (up to > 100 Gb/s per channel) communication. Copper-based communication is used for medium-distance (< 1 km) and medium-high data rates (1 Mb/s to a few Gb/s per channel). Wireless is used for medium distance (~ km) and medium data rates (up to ~100 Mb/s). The choice of a communication medium is largely determined by cost and application requirements. Clearly, fiber has the highest intrinsic bandwidth, so it can deliver the highest data rate possible for a single channel.

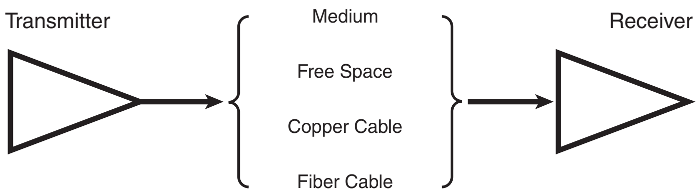

Figure 1.1 A simple communication system, including three basic building blocks: transmitter, medium, and receiver.

1.1.1 What Are Jitter, Noise, and Signal Integrity?

When a signal is transmitted and received, a physical process called noise is always associated with it. Noise is basically any undesired signals added to the ideal signal. In the context of digital communication, the information is encoded in logical bits of 1 and 0. An ideal signal may be represented by a trapezoid wave with a finite 0 to 1 rise time or 1 to 0 fall time. In the presence of noise, it is the sum of ideal signal, with the noise giving rise to the "net" or actual signal waveform. If no noise is added, the actual signal is identical to the ideal signal waveform. If the noise is added, the actual signal is deviated from the ideal signal, as shown in Figure 1.2.

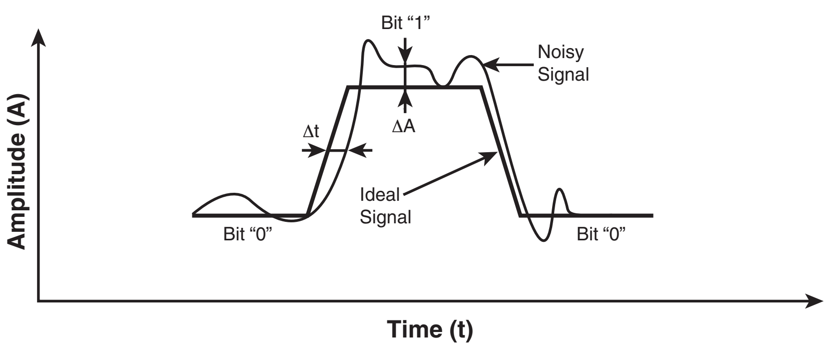

Figure 1.2 An ideal signal versus a noisy signal for a digital waveform.

The deviation of a noisy signal from its ideal can be viewed from two aspects: timing deviation and amplitude deviation. The amplitude of the digital signal for a copper-based system is the voltage, and for a fiber-based or radio frequency (RF) wireless system it is the power. The deviation of the signal amplitude (ΔA) is defined as the amplitude noise (or simply noise), and the deviation of time (Δt) is defined as the timing jitter (or simply jitter). Those definitions will be used throughout this book. The impacts of timing jitter and amplitude noise are not symmetrical, though. Amplitude noise is a constant function and can affect system performance all the time. Timing jitter affects system performance only when an edge transition exists.

Signal integrity generally is defined as any deviation from ideal waveform.[2] As such, signal integrity contains both amplitude noise and timing jitter in a broad sense. However, certain signal integrity signatures such as overshoot, undershoot, and ringing (see Figure 1.3) may not be well covered by either noise or jitter alone.

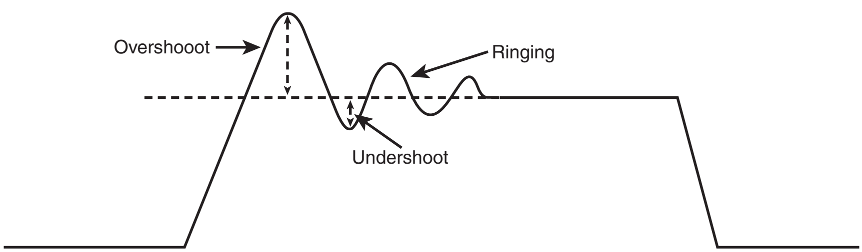

Figure 1.3 Some signal integrity key signatures.

1.1.2 How Do Jitter and Noise Impact the Performance of a Communication System?

There is no doubt that jitter, noise, and signal integrity all impact the quality of a communication system. The following sections discuss and illustrate how jitter and noise cause a bit error and under what conditions this bit error occurs. Then the metric that is commonly used to quantify the bit error rate in a communication system is discussed.

1.1.2.1 Bit Error Mechanisms

The impacts of timing jitter and amplitude noise can best be understood from the perspective of a receiver for a communication system.[3] A receiver samples the incoming logical 1 pulse data at a sampling time of t_s and threshold voltage of v_s, as shown in Figure 1.4. For a jitter- and noise-free digital pulse, an ideal receiver samples the data at the center of the incoming pulse. In this context, clearly there is no need to talk about signal integrity, because its effects are covered by jitter and noise. Under such conditions, threshold crossing times for rising and falling edges satisfying the conditions of $t_r < t_s < t_f$ and $V_1 > v_s$ result in a logical 1 being detected, and the data bit is received correctly (see part (a) of Figure 1.4). In the presence of jitter and noise, the rising and falling edges can move along the time axis, and the voltage level can move along the amplitude axis. As such, the correct bit detection conditions for sampling time and voltage may not be satisfied, resulting in a bit error due to bit 1 being received/detected as bit 0. The violations of those sampling conditions can occur in three scenarios:

- The crossing time of the rising edge lags behind the sampling time, or $t_r > t_s$.
- The crossing time of the falling edge is ahead of the sampling time, or $t_f < t_s$.
- The logical 1 voltage is below the sampling voltage v_s, or $V_1 < v_s$.

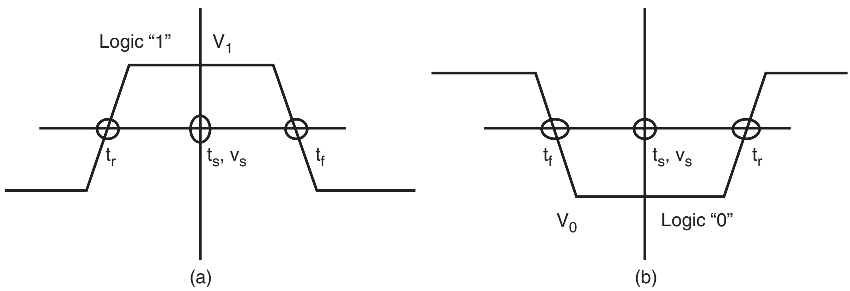

Figure 1.4 A receiver sampling an incoming data bit 1 (a) and 0 (b), where t_r and t_f are the timings for the 50% crossing (or zero crossing timings) for the rising and falling edges, respectively, and t_s and v_s are the sampling time and voltage, respectively.

For a zero pulse or bit "0"detection, in the case of part (b) of Figure 1.4, the correct detection condition becomes $t_r < t_s < t_f$ and $V_0 < v_s$. Similarly, the violation of correct sampling condition causes a bit error because bit 0 is received as bit 1. The violation scenarios for timing are similar to those of bit 1 pulse (part (a) of Figure 1.4). However, the violation condition for voltage becomes $V_0 > v_s$.

1.1.2.2 Bit Error Rate (BER)

We have demonstrated how jitter and noise cause a digital system bit error with a simple example. Because a digital system transmits and receives many bits for a given time, the system's overall performance can best be described by the rate of bit failure—namely, the ratio of the total failed bits N_f to the total bits received N. This ratio is called the bit error rate (BER) or bit error ratio. Bit error ratio is a more precise definition because BER = N_f/N and no normalization of time is as involved as most of the rate definition otherwise required.

BER is the bottom-line metric for determining a good communication system. At multiple Gb/s rates, the BER requirement for most communication standards such as Fibre Channel (FC), Gigabit Ethernet, SONET, and PCI Express is 10^{-12} or smaller. Larger BER degrades network or link efficiency and, worse, system latency. A simple implication of BER = 10^{-12} is that with 10^{12} bits being transmitted/received, only one bit error is allowed. Clearly BER depends on data rate, jitter, and noise in the communication system. The definition of BER implies that BER is a counting statistic so that Poisson statistics may apply.

1.2 SOURCES OF TIMING JITTER, AMPLITUDE NOISE, AND SIGNAL INTEGRITY

Jitter and noise are deviations from an ideal signal. Jitter and noise can have many causes. The physical nature of various noise and jitter sources for a communication system can be classified into two major classes: intrinsic and nonintrinsic. The intrinsic type has to do with the physical properties of electrons and "holes" in electrical or semiconductor devices. The nonintrinsic type are design-related and may be eliminated. These types are discussed in detail in the following sections.

1.2.1 Intrinsic Noise and Jitter

Intrinsic noise is fundamentally caused by the randomness and fluctuation of electrons and "holes" existing in all the electronic/optical/semiconductor circuits/devices. Intrinsic noise can be minimized but cannot be completely removed from devices or systems. Therefore, this kind of noise puts a fundamental limit on device and system performance and dynamic range. Typical intrinsic noises in electrical-optical devices include thermal noise, shot noise, and flick noise.

1.2.1.1 Thermal Noise

Thermal noise is caused by the random motion of charge carriers under the thermal equilibrium condition. The kinetic energy of those randomly fluctuating charge carriers is proportional to their temperature, as well as to their mean-square velocity. The power spectrum density (PSD) of the thermal noise is white and apparently proportional to its temperature. Thermal noise places a fundamental limit on the signal-to-noise ratio (SNR) performance because it exists in all electric/optical/semiconductor devices having a nonzero absolute temperature. Johnson[4] first discovered that the noise in a conductor depends on temperature and resistor under the thermal equilibrium condition. Nyquist[5] shortly after developed a theory to explain Johnson's discovery based on the second law of thermodynamics. Because of their pioneering contributions, thermal noise is sometimes called Johnson noise or Nyquist noise.

1.2.1.2 Shot Noise

Shot noise is produced by individual quantized carrier flow (current) in a potential barrier with a random generation time or spatial distribution. In other words, shot noise is basically due to random flow fluctuation. Schottky[6] first studied shot noise in vacuum tube diodes and later it was also found in P-N junction in a semiconductor transistor. Shot noise is directly proportional to DC bias current, as well as the charge of the carrier. Shot noise is typically larger than thermal noise in semiconductor devices.

1.2.1.3 Flick Noise

Flick noise is a phenomenon that is found to have a noise power spectrum inversely proportional to the frequency over a wide range of frequencies. Johnson was the first to observe flick noise in an electronic system.[7] Flick noise can be found in all active devices, and some passive devices such as carbon resistors. DC current is necessary to produce flick noise. No universally accepted theory explains the cause and mechanism of flick noise, unlike the causes of thermal and shot noise. As a result, the quantitative study of flick noise is mostly empirical. It has been found that the PSD of flick noise is proportional to $1/f^{\alpha}$, where α is around 1. Because of this reason, flick noise is also called 1/f noise. One common interpretation of flick noise is the "trap and release" theory. It is believed that the flow of carriers due to the DC current can be trapped due to contamination and defects in devices. However, the "trap and release" process is random, giving rise to the flick noise that is most significant at low frequencies.[8]

1.2.2 Translation of Noise to Timing Jitter

Noise is typically described using physical quantities or parameters. In communication, computer, and electronic systems, those quantities may include voltage, current, or power. We use the generic term of amplitude to represent those physical quantities. Assuming that the amplitude noise $\Delta A(t)$ is superimposed on the amplitude waveform of $A_0(t)$ so that the total waveform has the following form:

$$A(t) = A_o(t) + \Delta A(t)$$

Equation 1.1

the corresponding timing jitter can be estimated through the linear small-signal perturbation theory as the following:

$$\Delta t(t) = \Delta v \, / \, (\frac{dA_o(t)}{dt}) = \Delta t \, / \, k$$

Equation 1.2

where k = $(dA_0(t)/dt)$ is the slope or slew rate of the waveform.

This linear amplitude noise to timing jitter conversion is shown in Figure 1.5.

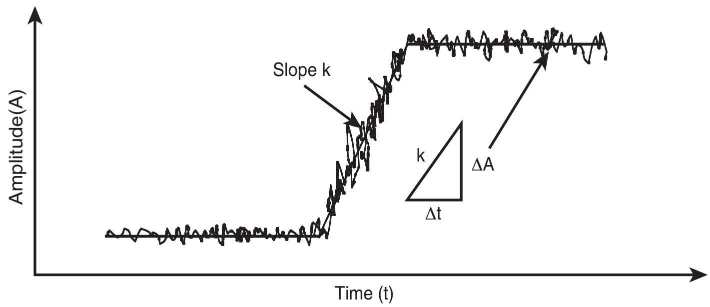

Figure 1.5 Amplitude noise to timing jitter conversion through the linear perturbation model.

You can see that for amplitude noise ΔA, the corresponding timing jitter decreases as the slope increases, and vice versa. To maintain a smaller timing jitter conversion, a large slope or fast slew rate is favored. In the context of a digital signal, this implies a small rise/fall time.

1.2.3 Nonintrinsic Noise and Jitter

Nonintrinsic jitter and noise are design-related deviations. In other words, those types of jitter and noise can be controlled or fixed with appropriate design improvements. Commonly encountered nonideal design-related noise and jitter include periodic modulation (phase, amplitude, and frequency), duty cycle distortion (DCD), intersymbol interference (ISI), crosstalk, undesired interference such as electromagnetic interference (EMI) due to radiation, and reflection caused by unmatched media. The following sections discuss these noise sources and their root causes.

1.2.3.1 Periodic Noise and Jitter

Periodic noise or jitter is a type of signal that repeats every time period. It can be described mathematically by the following general equation:

$$\Delta t_P = f(2\pi \frac{t}{T_0} + \phi_0)$$

Equation 1.3

where T_0 is the period, t is the time, and ϕ_0 is the phase of the periodic signal. The period T_0 and frequency f_0 satisfy the reciprocal relationship of $T_0 = 1/f_0$. Although the notation and discussion are based on timing jitter, the same type of discussion can be applied to amplitude noise. The frequency-domain periodic function can be obtained through Fourier Transformation (FT), a subject that is discussed in Chapter 2, "Statistical Signal and Linear Theory for Jitter, Noise, and Signal Integrity."

Periodic jitter can be caused by various modulation mechanisms, such as amplitude modulation (AM), frequency modulation (FM), and phase modulation (PM). Moreover, the modulation function can have various shapes. Typical modulation shapes include sinusoidal, triangular, and sawtooth. It is apparent that a periodic amplitude noise causes period timing jitter, with the amplitude proportional inversely to the slope or slew rate of the edge transition, as discussed in section 1.3.2. In the computer environment, period noise/jitter can be caused by switching power supply, spread-spectrum clock (SSC), and period EMI sources.

1.2.3.2 Duty Cycle Distortion (DCD)

DCD is defined as the deviation in duty cycle from its normal value. Mathematically, a duty cycle is the ratio of pulse width to its period for a clock signal, as shown in Figure 1.6.

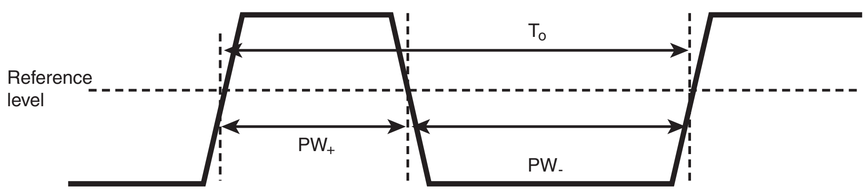

Figure 1.6 Illustration of period (T_0), pulse width PW_+/PW_- (either positive or negative), and reference level for a periodic signal.

Duty cycle is defined as follows:

$$\eta_+ = \frac{PW_+}{T_0}, \quad \eta_- = \frac{\mathrm{PW}_-}{\mathrm{T}_0}$$

Equation 1.4

Most clocks have a nominal duty cycle of 50%. So either shorter pulse width or longer pulse width causes DCD. DCD can be caused by pulse width deviation, period deviation, or both. Furthermore, pulse width deviation can be caused by the deviation of reference signal level. Another DCD-causing mechanism is propagation delay if the clock is formed from rising and falling edges of two half-rate clocks and those two half-rate clocks undergo different propagation delays. Because a clock can have many periods, DCD must be looked at from the distribution point of view with many samples considered, and the average period should be used for the overall DCD estimation.

1.2.3.3 Intersymbol Interference (ISI)

ISI is related to data signal, but a clock signal does not have ISI by definition. A data signal is a generic digital signal form that does not have to have an edge transition in every UI or bit period, like the clock signal. The data signal can be kept at the same amplitude level for many UIs without an edge transition, whereas a clock signal cannot be. The type of data pattern used in digital communication critically depends on the coding scheme of the communication architecture.[9] An important parameter for digital pattern is the run length, which is defined as the maximum length of consecutive 1s or 0s within a pattern. The run length determines the lowest frequency of the data pattern spectrum and therefore governs the frequency range for the test coverage. The long-haul fiber-optic communication standard SONET uses a scramble code scheme and can have a much longer run length (such as a run length of 23, 31) and therefore relatively low-frequency spectral content. A short-haul data communication standard such as Fibre Channel or Gigabit Ethernet uses block code (e.g, 8B10B coding) that has a shorter run length (e.g., a run length of 5) and relatively high-frequency spectral content.

In a lossy medium, the previous bits can cause both transition timing and amplitude level off the ideal values. In copper-based communication systems, this is due to the "memory" characteristics of the electronic devices used to switch bits between 1s and 0s. One example of this "memory" nature is the capacitive effect. Due to capacitive effect, each transition has a finite charge or discharge time. If the transition happens such that the next transition occurs before the previous transition reaches the designated level, deviation of both time and level occurs for the current bit. Such an effect can be cascaded. The ISI effect is shown in Figure 1.7.

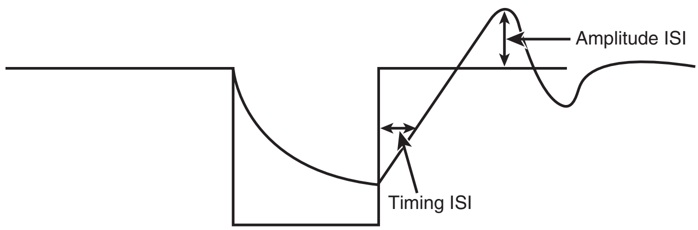

Figure 1.7 The ISI effect for both timing and amplitude.

Any pulse-width broadening or spreading effects cause ISI, and dispersion is a known physical phenomenon that causes a traveling pulse to be broadened or spread. As such, ISI is expected to occur in a fiber-based communication system too.[10] For multimode fiber, the spread mechanism is called mode dispersion (MD), where a number of electromagnetic waves can exist in the multimode fiber waveguide, and the number of wave modes depends on the physical parameters of the multimode fiber, such as refraction index and geometry. Those different modes have different propagation times. The spread of the propagation times in multimode fiber cause the pulse to spread at the other end of the fiber. For a single-mode fiber, the dominant spread mechanism is the dispersion effects, including chromatic dispersion (CD) and polarization mode dispersion (PMD). The physical reason for CD is that the refraction index of the fiber material is wavelength-dependent. Therefore, the group velocity of the wave propagation inside the fiber is wavelength-dependent. Both laser source and modulation waveform have some spread in their spectrum. The combined spread spectrum of the input optical waveform, coupled with the CD effect, causes the optical pulse train to spread in the time domain, resulting in both timing and amplitude ISI. PMD is due to the birefringence, in which the refraction indexes along the two orthogonal axes are different, causing different propagation velocities. Again, the two different velocities for the two orthogonal modes of PMD eventually cause pulse train at the other end of the fiber to spread, resulting in ISI. Figure 1.8 shows the dispersion effects on a pulse for an optical fiber.

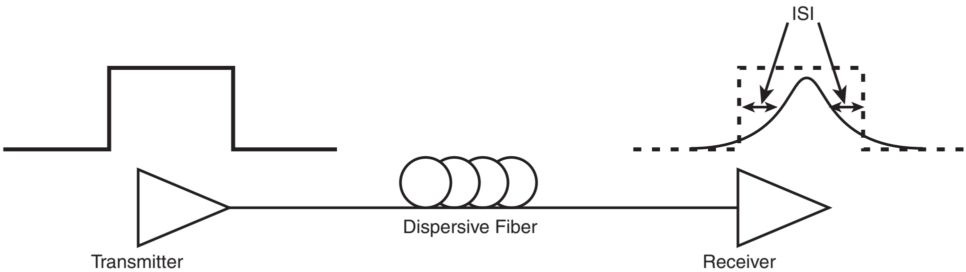

Figure 1.8 ISI effects in a fiber-based communication link.

1.2.3.4 Crosstalk

Two types of crosstalk are discussed here. One is associated with copper cables, and the other is associated with optical fibers.

1.2.3.4.1 Copper-Based Crosstalk

Crosstalk is basically an interference phenomenon. Crosstalk is generally involved in a parallel channel system in which signals are propagated concurrently and affect each other. For copper-based communication channels, crosstalk is caused by electromagnetic coupling. For integrated circuits (ICs) where the geometry and space between connects is relatively small, the capacitive coupling is the dominant mechanism.[11, 12] When a signal transition happens in one channel, some of its energy leaks to the neighboring or adjacent channel through charge flow due to capacitive coupling, causing the signal level in that channel to fluctuate. For board-level circuits where the geometry is relatively large, inductive and capacitive coupling are both important. Inductive coupling follows Lentz's Law, in which changing the magnetic field flux generates an electrical field, and that electrical field, coupled with electrical charge, causes voltage fluctuation. In general, the effect of crosstalk can be modeled primarily as the voltage fluctuation or noise. However, it can affect the timing jitter directly as well. When two transmission lines are coupled capacitively, and when digital transitions occur simultaneously on two lines from the same end (the near end), the slew rate of the signals at the other end (the far end) is larger if the two transitions at the near end are in phase (have the same polarity) or is smaller if the two transitions are out of phase (have the opposite polarity). Figure 1.9 shows the capacitive and inductive coupling mechanisms for crosstalk.

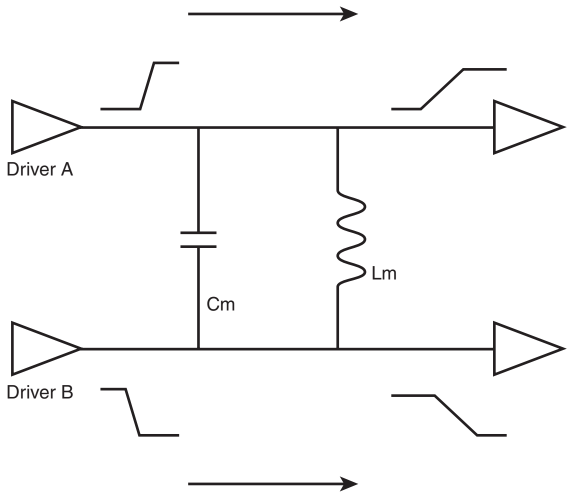

Figure 1.9 Schematic drawing of crosstalk caused by capacitive and inductive coupling. The crosstalk due to the simultaneous steps' response with opposite polarities slows down the slew rate of the step signals at the far end.

From the definitions of mutual capacitive and inductive constants C_m and L_m, the voltage noises due to capacitive and inductive coupling can be calculated according to the following equation:

$$V_{mc} = Z_v C_m \frac{dV_d}{dt}$$

Equation 1.5

where Z_v is the impendence of the impacted or victim line and dV_d/dt is the time derivative of the driving voltage. For inductive-induced voltage noise, we have

$$V_{mL} = L_m \frac{dI_d}{dt} = \frac{L_m}{Z_d} \frac{dV_d}{dt}$$

Equation 1.6

where Z_d is the impendence of the driving line, and dI_d/dt and dV_d/dt are driver current and voltage time derivatives or change rate, respectively.

You can see that crosstalk is proportional to the voltage or current slew rate. As the date rate or frequency keeps increasing, the rise time of the digital signal becomes smaller. Therefore, the slew rate and crosstalk-induced noise increase.

As mentioned in previous sections, timing jitter due to crosstalk can be estimated through division of appropriated far-end signal slew rate.

1.2.3.4.2 Fiber-Based Crosstalk

Crosstalk can also happen in optical fiber-based communication systems, particularly in multiple-channel systems such as wavelength division multiplexing (WDM) systems.[13] In a WDM or dense WDM (DWDM) system, crosstalk can happen through linear and/or nonlinear effects. Linear effects often refer to the leaking of photon energy from neighboring channels that have different wavelengths to the concerned channel in the optical filters or demultiplexers, causing the amplitude noise fluctuation. Nonlinear effects include the following:

- Stimulated Raman Scattering (SRS), in which short-wavelength channels can amplify long-wavelength channels over a wide wavelength range
- Stimulated Brillouin Scattering (SBS), in which short-wavelength channels can amplify long-wavelength channels over a narrow wavelength range
- Four-wave mixing (FWM), in which a new wave or signal, or the fourth wave, is generated when three wavelengths from three WDM channels satisfy a certain relationship

Like copper-based crosstalk, fiber-based crosstalk causes amplitude noise for the transmitting signal and subsequently causes timing jitter through the slew rate conversion, in turn degrading system performance.

1.3 SIGNAL AND STATISTICAL PERSPECTIVES ON JITTER AND NOISE

We will first talk about the limitations and drawbacks of peak-to-peak-based metrics for jitter. Then we will discuss why the jitter component method of quantifying jitter is better and more accurate and should be used to describe and quantify statistical processes such as jitter and noise.

1.3.1 Peak-to-Peak and Root-Mean-Square (RMS) Description

For many years, jitter was quantified by peak-to-peak value and/or standard deviation (1 σ or rms) of the entire jitter histogram or distribution. It is now widely realized that this can be very misleading. In the presence of random and unbounded jitter or noise (such as thermal noise or shot noise), expected peak-to-peak value is a monotonically increasing function of statistical sample size. Peak-to-peak value is a useful parameter for bounded jitter or noise but not for unbounded ones. Similar problems occur with the standard deviation calculation.

In the presence of bounded, non-Gaussian jitter or noise, the total jitter or noise histogram or distribution is not a Gaussian, and the statistical standard deviation or rms estimation does not equal the 1 σ of the true Gaussian distribution. Therefore, the latter is the correct quantity to describe a Gaussian process or distribution. Using standard deviation or rms based on the total jitter or noise histogram statistics "inflates" the true 1 σ value for Gaussian process.

To demonstrate the incorrect usage of statistical peak-to-peak in the presence of unbounded Gaussian jitter or noise, we start with a single Gaussian distribution via Monte Carlo method. We determine the peak-to-peak value for a given sample size N that is monotonically increasing and then plot the peak-to-peak value as a function of sample size. Figure 1.10 shows the results, clearly demonstrating the monotonicity trend.

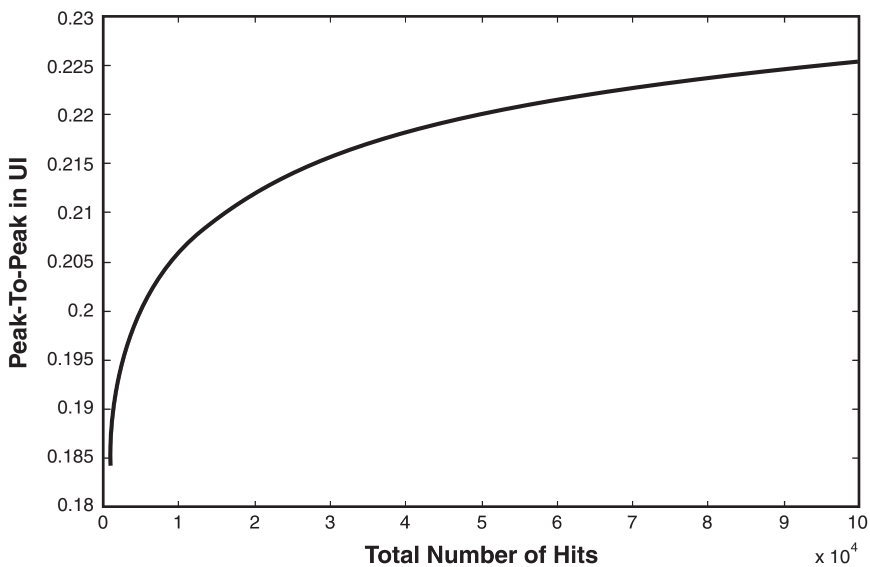

Figure 1.10 Peak-to-peak value plotted as a function of number of samples (N). The histogram distribution is a Gaussian, and the 1 σ of the Gaussian equals 0.03 UI (bit clock period).

To demonstrate how different a statistical standard deviation or rms and 1 σ of a Gaussian distribution can be, we assume that the histogram distribution has a bimodal distribution that is the superimposition of two identical Gaussians with different mean positions. Each peak corresponds to a single Gaussian mean position. Then standard deviation for such a bimodal distribution is 1.414 times (or 41.4% larger than) the true Gaussian 1 σ value when they are well separated (10 σ apart).

As the goal becomes to completely grasp the jitter or noise process, as well as to quantify the overall distribution and its associated components and root causes, the simple parameter-based approach to jitter or noise becomes insufficient and invalid. What is needed is the distribution function such as probability density function (PDF) and its associated component PDFs. Those PDFs not only give the overall description for jitter or noise statistical process, but also give the corresponding root causes.

1.3.2 Jitter or Noise PDF and Components Description

Jitter or noise is a complex statistical signal and therefore can have many components associated with it. We will focus on jitter, but the same concept applies well to noise. In general, jitter can be split into two components: deterministic jitter (DJ) and random jitter (RJ). The amplitude of DJ is bounded, and that of RJ is unbounded and Gaussian. This classification scheme is the first step in jitter separation.[14]

Jitter can be further separated after the first-layer splitting, as shown in Figure 1.11. Within deterministic jitter, jitter can be further classified into periodic jitter (PJ), data-dependent jitter (DDJ), and bounded uncorrelated jitter (BUJ). DDJ is the combination of DCD and ISI. BUJ can be caused by crosstalk. Within random jitter, jitter can be single-Gaussian (SG) or multiple-Gaussian (MG). Each jitter component has some specific corresponding root causes and characteristics. For example, the root cause of DJ can be a bandwidth-limited medium, reflection, crosstalk, EMI, ground bouncing, periodic modulations, or pattern dependency. The RJ source can be thermal noise, shot noise, flick noise, random modulation, or nonstationary interference.

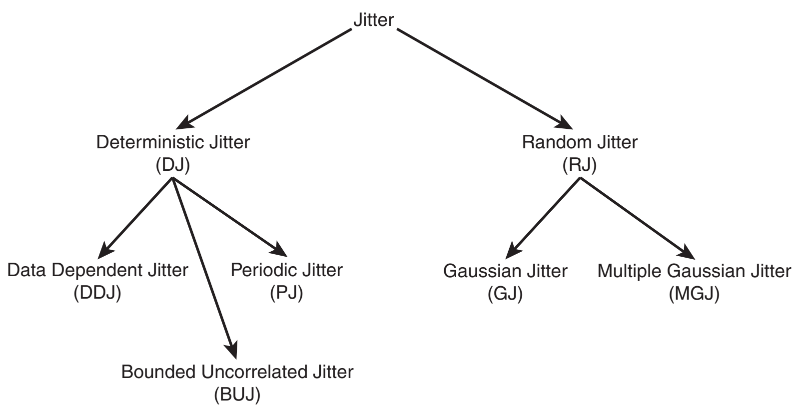

Figure 1.11 Jitter classification scheme from a signal statistical view.

A similar type of noise component tree classification can be developed, as shown in Figure 1.12.

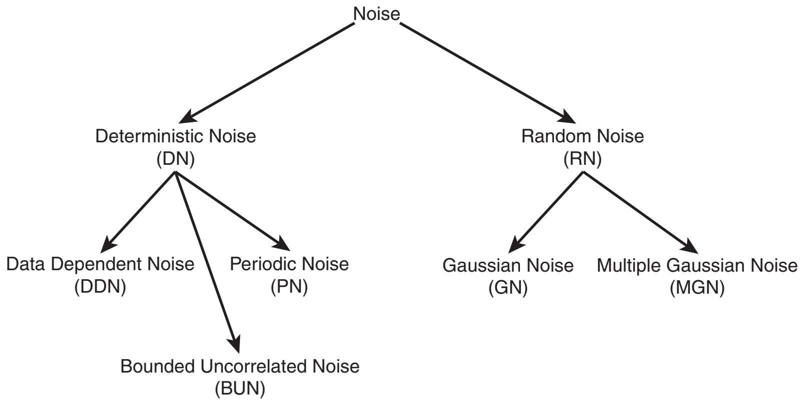

Figure 1.12 Noise classification scheme from a signal statistical view.

Most of the component concepts for jitter and noise are symmetrical, except DCD, which does not have a noise counterpart. Also, the same type of jitter and noise component may or may not be correlated.

1.4 SYSTEM PERSPECTIVE ON JITTER, NOISE, AND BER

This section briefly discusses jitter, noise, and BER within a high-speed linksystem. It also covers the role that clock recovery plays in providing the timing reference and in tracking low-frequency jitter, as well as jitter transfer functions.

1.4.1 The Importance of Reference

The beginning of this chapter defined jitter as any deviation from ideal timing. This definition is from the point of view of a “static timing reference” (see Figure 1.13). In other words, the ideal timing reference is a fixed timing point. This definition is very useful from concept and mathematical views, but it needs to be enhanced to be useful for the system application. Although it’s true in a wide sense that jitter is any deviation from the ideal, if the properties of the reference are considered, the resulting jitter can be quite different. For example, a data signal with a sinusoidal timing jitter referenced to an ideal clock with a perfect period (i.e., zero-jitter) has a larger peak value than when it is referenced to the same clock but modulated with the same kind of sinusoidal, because the reference clock moves "in phase" with the data signal in this case.

This is in analogy to Newton's law for motion. Whether or not an object moves critically depends on the reference. In parallel, we can fairly say that whether or not a signal has jitter depends on the reference signal used to determine the timing. For illustration purposes, we will focus on a timing reference signal in the context of serial data communication. However, the general concept applies to other systems.

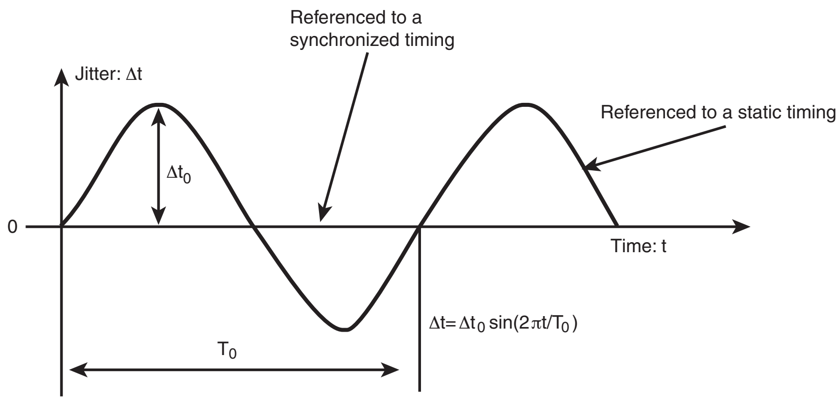

Figure 1.13 The same jitter source results in two different jitter estimations with two different jitter references. One is a static ideal timing, and the other is a synchronized timing giving rise to zero jitter estimation.

1.4.2 Jitter Transfer Function in Serial Data Communication

Serial data communication embeds the clock signal in its transmitting data bit stream. At the receiver side, this clock needs to be recovered through a clock recovery (CR) device where phase-locked loop (PLL) circuits are commonly used. It is well known that a PLL typically has certain frequency response characteristics. Therefore, when a receiver uses the recovered clock to time or retime the received data, the jitter seen by the receiver follows certain frequency response functions. Figure 1.14 shows a typical serial link system with a transmitter (Tx), medium or channel, and receiver (Rx).

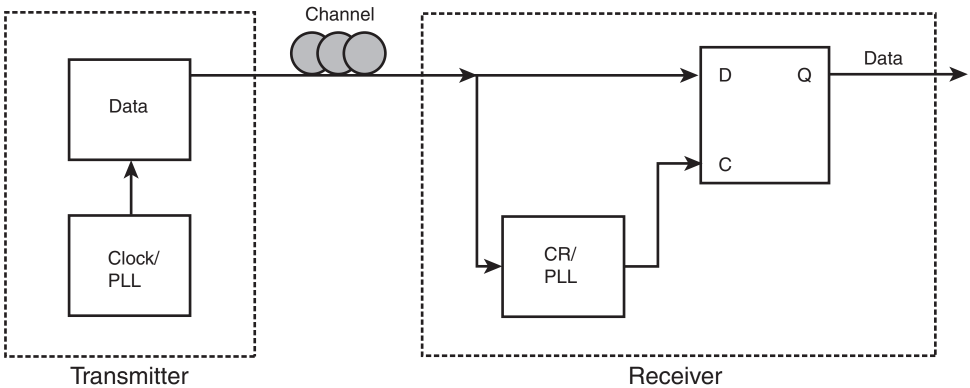

Figure 1.14 A schematic block diagram for a serial link composed of three key elements: transmitter (Tx), medium (or channel), and receiver (Rx). Clock for Tx data generation and clock recovery (CR)/PLL for receiver are also shown.

A PLL typically has a low-pass frequency response function $H_L(f)$, as shown in Figure 1.15.

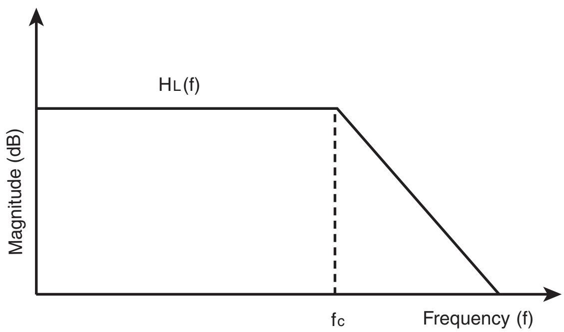

Figure 1.15 A typical PLL magnitude frequency response.

Any good estimation methodology should emulate the actual device behavior. In the case of receiver jitter, noise, and BER estimation/measurement, the model/measurement setup should estimate/measure the jitter as what a receiver sees. A receiver "sees" jitter on the data from its recovered clock.[15] Therefore, it is a difference function from clock to data, as shown in Figure 1.16.

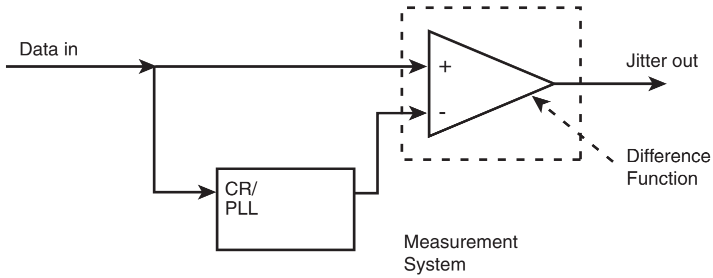

Figure 1.16 A jitter estimation/measurement system emulates jitter as seen by a serial data receiver. Note that the data latch function of "D" flip-flop in Figure 1.14 is replaced by the difference function to emulate the receiver jitter behavior.

Because the clock recovery (or PLL) device has a low-pass transfer function $H_L(f)$, the jitter output has a high-pass transfer function of $H_H(f)$, as shown in Figure 1.17. $H_L(s) + H_H(s) = 1$, where s is a complex frequency.

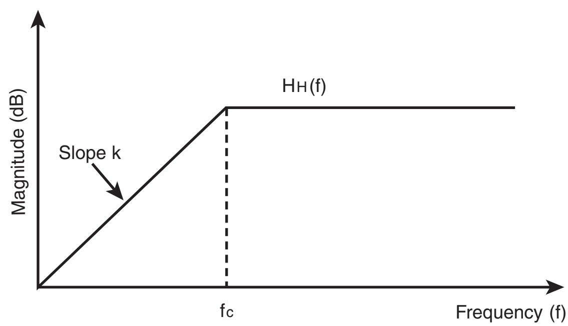

Figure 1.17 Jitter frequency response as seen by a serial receiver or as measured by a difference function.

The high-pass jitter transfer function shown in Figure 1.17 suggests that a receiver can track more low-frequency jitter at frequencies of $f < f_c$ than at higher frequencies of $f > f_c$. This implies that a receiver can tolerate more low-frequency jitter than high-frequency jitter, with a jitter tolerance function being the reciprocal of the jitter output function, as shown in Figure 1.17. Figure 1.18 shows the jitter tolerance mask corresponding to the jitter transfer function in Figure 1.17.

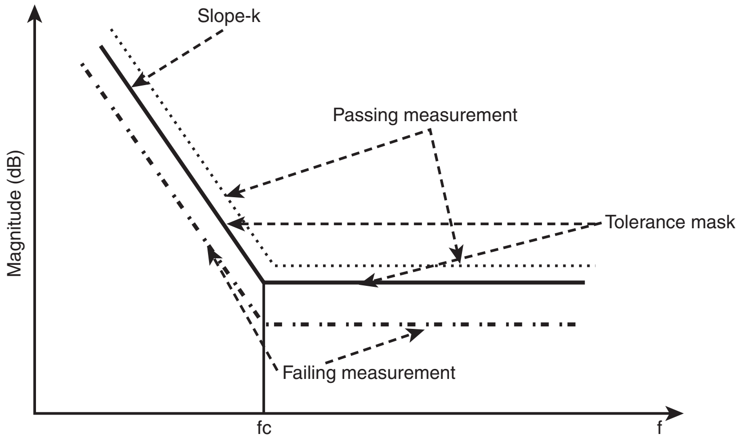

Figure 1.18 The receiver jitter tolerance mask corresponding to the jitter transfer function shown in Figure 1.17.

Notice the same magnitude but different polarity slopes in Figures 1.17 and 1.18 at frequencies at $f < f_c$. For a receiver tolerance test, a receiver should be able to tolerate more jitter than those defined in the Figure 1.18 mask. So the mask is a minimum jitter magnitude as a function of frequency that a receiver must satisfy. When the mask has a second-order slope—namely, –40dB/decade—a receiver with a first-order jitter transfer function with a slope of 20 dB/decade does not meet the tolerance requirement. A second-order jitter transfer function may meet the tolerance requirement, and a third-order jitter transfer function with a 60 dB/decade slope will exceed the tolerance requirement.

The jitter transfer function is a very important element in estimating the relevant jitter in a serial link. Without this building block, it is not possible to estimate the relevant jitter for the system and related BER performance in a rational way. We will give detailed discussions by using the jitter transfer function when we discuss specific communication link technologies in the upcoming chapters.

1.5 HISTORICAL OVERVIEW OF JITTER, NOISE, BER, AND SIGNAL INTEGRITY

During the last two decades, two books were published with significant space dedicated to jitter analysis.[16, 17] During that time, most communication architectures operated at a data rate of less than 1 Gb/s. Jitter was not as serious then as it is today, when most leading communication links are running at rates of 1 to 10 Gb/s.

The book by Trischitta & Varma[16] in 1989 was mostly focused on the accumulated jitter in a network system and jitter related to some specific components in an optical network at the time, including regenerators, retimers, and multiplexers. The jitter handling in this book was tightly coupled with the link architecture of almost 20 years ago, so many of the concepts and theories in this book do not apply well to the serial link architectures developed after the 1990s.

The book by Takasaki[17] in 1991 treated digital transmission design and jitter in the same context. This book is weighted more toward digital transmission, with only two chapters dedicated to jitter topics, covering jitter generation and accumulation. This book does discuss jitter classification in some way. The major point of Takasaki on jitter classification is that there are two types of jitter: random and systematic. However, it has no quantitative math model discussion on the jitter classification scheme. It also has no further discussion of the jitter component beyond random and systematic. The discussion of jitter accumulation is largely based on the repeater component in a network.

In the past 15 years, significant progress has been made in the field of understanding jitter, and related new theory, definition, analysis methods, and measurement tools. In particular, more rigorous definition of and theory about jitter and its associated jitter components have been developed (such as [15], [18], and [19], to name a few). Now jitter and noise component concepts have been widely accepted and adopted by many serial data communication standards. In fact, jitter and noise component concepts are required for determining the link jitter budget, for debugging and diagnostics for designing and testing most of the multiple Gb/s serial data links, and for setting standards. In addition, a generic jitter transfer function for a linear or quasi-linear system has recently been developed. Such a method can be applied to jitter analysis for most of the serial data communication links and standards [15], [20]. The combination of the statistical and system transfer function elements of a link system in estimating overall jitter, noise, BER (JNB), and signal integrity (SI) performance has put the research and application in those areas on a new historical plateau.

In light of that significant progress in JNB and SI, as well as the ever-increasing importance of their roles in > 1 Gb/s serial communication links in both network and PC applications, a new book summarizing that progress—with an emphasis on the latest definitions, theories, and applications, as well as simulation modeling, measurement, and analysis technologies—is apparently greatly needed.

1.6 OVERVIEW OF THIS BOOK

This book systematically presents the latest developments and advancements in jitter, noise, and BER (JNB) and SI. It guides you from the basic math, statistics, circuit, and system models to the final practical applications. It covers fundamental theory, to JNB and SI simulation/modeling, to JNB and SI diagnostics/debug and compliance testing, with an emphasis on two major applications: clock and serial data communications. It tries to keep a good balance and coupling between theory and practical applications.

As you have seen, this chapter is a high-level overview of JNB and SI basics. This chapter introduced the JNB component classification scheme, JNB interrelationship, root-cause mechanisms, JNB measurement references, clock recovery, and associated JNB transfer functions.

Chapter 2 reviews and introduces the basic theories on statistics, linear time-invariant (LTI) systems, and digital signal processing that you need to quantitatively understand and model JNB and its related components. Also introduced in this chapter is the statistical signal process theory that is used to quantify the JNB spectrum and power spectrum density (PSD).

Chapter 3 describes the jitter component in a quantitative manner. Detailed root causes and mathematical models and treatments for each jitter component are given. This chapter lays the necessary physical and mathematical foundation for jitter and noise to warrant the precision of the estimations. Jitter component math models can be applied to noise components similarly.

Chapter 4 discusses jitter, noise, and BER correlatively from the view of statistical signal processing. It first discusses the jitter total PDF and its relationship with the PDFs of its component. It then discusses the noise total PDF and its relationship with the PDFs of its component. It also discusses the joint PDFs, with both jitter and noise being considered. Finally, it discusses the BER cumulative distribution function (CDF) and its relationship to both corresponding jitter and noise PDFs. Applications of two-dimensional eye diagram and BER contour are introduced.

Chapter 5 focuses on jitter separation methods in the statistical distribution domain. Equipped with basic math knowledge, as well as the mechanism and physical nature of each jitter component, you will read about jitter separation. Jitter separation is an important step toward understanding and quantifying jitter components, because in reality the jitter that we observe or measure is a "compound" jitter with many components. Jitter separation methods based on jitter PDF and BER CDF are introduced. The "Tailfit" method in which random jitter PDF is quantified by a Gaussian distribution is first introduced for PDF-based jitter separation. Then the same method is applied to BER CDF-based distribution, in which the random jitter CDF is quantified by the integration of a Gaussian—namely, an error function.

Chapter 6 discusses jitter separation methods in the time and frequency domain. Both spectrum (first moment) and PSD (second moment)-based jitter separation methods are introduced, and associated advantages and disadvantages of each method are given. This chapter also compares statistical PDF/CDF-based methods with spectrum and PSD-based jitter separation methods.

Chapter 7 focuses on clock jitter because it is an important topic for all digital systems, so it deserves a dedicated treatment. New concepts of phase jitter, period jitter, cycle-to-cycle jitter, and their corresponding relationships are discussed. The mathematical relationship between phase jitter, period jitter, and cycle-to-cycle jitter is given in both time and frequency domain. Furthermore, phase jitter and its relationship to the conventional phase noise for quantifying the performance of clock or crystal oscillator in the frequency spectrum domain are also discussed.

Chapter 8 focuses on PLL jitter because it is widely used in clock generation and clock recovery and is an important metric for any high-speed PLL. Jitter at the PLL output and its relationship with PLL building elements such as phase detector (PD), low-pass filter (LPF), voltage control oscillator (VCO), and dividing/multiplying are discussed. Jitter as a function of PLL reference and internal noise sources and transfer functions is derived and applied to PLL and its jitter analysis for second-order, third-order, and general nth-order PLL implementations.

Chapter 9 is dedicated to jitter, noise, and SI mechanisms and root-cause sources or mechanisms for a high-speed link system. Those mechanisms are discussed within the context of link architecture, including its subsystems of transmitter, receiver, channel, and reference clock. For the transmitter, reference clock jitter and voltage driver noise are discussed. For the receiver, jitter from clock recovery circuits and data sampler are the focus. For the channel, various losses in both copper- and optical-based channels are covered. For the reference clock, jitter due to PLL or crystal oscillator, as well as spread-spectrum clocking (SSC), are presented. This chapter also discusses the link jitter budget method using the RJ root-sum-square (RSS) method to ensure the link's interoperability and overall BER performance.

Chapter 10 focuses on quantitative modeling and analysis for jitter, noise, and SI. Modeling and analysis methods are presented for link subsystems of transmitter, receiver, and channel within the framework of the LTI system theory. By using the cascading property of the LTI theorem, signal, jitter, and noise at the channel and receiver outputs are readily obtainable. Important subsystems of equalization and clock recovery are included in the modeling and analysis. For the equalization, both transmitter and receiver equalizations are considered. The modeling and analysis methods introduced in this chapter can give estimations for most advanced serial links today and are scalable to future link advancement given that they are LTI-based.

Chapter 11 is dedicated to the various testing aspects of jitter, noise, and SI. It describes test implications and requirements for the link architectures/topologies operating mechanisms, with a focus on clock recovery and equalization. Testing requirements and methods for links with clock recovery and equalization are presented, covering transmitter, channel or medium, reference clock, and PLL. System test methods such as loopback test also are discussed.

Chapter 12 is an executive summary and overview of the entire book. Future works and trends for JNB and SI at high speed as data rates keep increasing also are discussed.

ENDNOTES

1. C. E. Shannon, "A Mathematical Theory of Communication," *Bell System Technical Journal*, vol. 27, pp. 379–423, 623–656, July, October, 1948.
2. H. Johnson and M. Graham, *High-Speed Digital Design: A Handbook of Black Magic*, Prentice-Hall, 1993.
3. A. B. Carlson, *Communication Systems: An Introduction to Signals and Noise in Electrical Communication*, Third Edition, McGraw-Hill, 1986.
4. J. B. Johnson, *Phys. Rev.*, vol. 32, pp. 97–109, 1928.
5. H. Nyquist, *Phys. Rev.*, vol. 32, pp. 110–113, 1928.
6. W. Schottky, *Ann. Phys.*, 57, 541, 1918.
7. J. B. Johnson, "Electronic Noise: The First Two Decades," *IEEE Spectrum*, vol. 8, pp. 42–46, 1971.
8. A. Van Der Ziel, *Noise in Solid State Devices and Circuits*, Wiley InterScience, 1986.
9. S. Lin and D. J. Costello, Jr., *Error Control Coding: Fundamentals and Applications*, Prentice-Hall, 1983.
10. G. P. Agrawal, *Fiber Optic Communication Systems*, a Wiley InterScience Publication, John Wiley & Sons, Inc., Second Edition, 1997.
11. S. H. Hall, G. W. Hall, and J. A. McCall, *High-Speed Digital System Design: A Handbook of Interconnect Theory and Design Practices*, a Wiley InterScience Publication, John Wiley & Sons, Inc., 2000.
12. M. Li, *Design and Test for Multiple Gbps Communication Devices and Systems*, International Engineering Consortium (IEC), 2005.

13. National Committee for Information Technology Standardization (NCITS), Working Draft for "Fiber Channel—Methodologies for Jitter Specification," Rev. 10, 1999.
14. R. E. Best, *Phase-Locked Loops: Design, Simulation, and Applications*, Fourth Edition, McGraw-Hill, 1999.
15. M. Li and J. Wilstrup, "Paradigm Shift for Jitter and Noise in Design and Test > 1 Gb/s Communication Systems," an invited paper, IEEE International Conference on Computer Design (ICCD), 2003.
16. P. R. Trischitta and E. L. Varma, *Jitter in Digital Transmission Systems*, 1989, Artech House.
17. Y. Takasaki, *Digital Transmission Design and Jitter Analysis*, 1991, Artech House.
18. J. Wilstrup, "A Method of Serial Data Jitter Analysis Using One-Shot Time Interval Measurements," IEEE International Test Conference (ITC), 1998.
19. M. Li, J. Wilstrup, R. Jessen, and D. Petrich, "A New Method for Jitter Decomposition Through Its Distribution Tail Fitting," IEEE International Test Conference (ITC), 1999.
20. M. Li, "Statistical and System Approaches for Jitter, Noise and Bit Error Rate (BER) Tests for High Speed Serial Links and Devices," IEEE International Test Conference (ITC), 2005.