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Motivation and Context

The goal of this book is to present recent advances in space-time (ST) coded multi-antenna systems operating over broadband wireless mobile channels. To appreciate these advances, we start this chapter by tracing the evolution of wireless communications through past, present, and future-generation systems, which motivates the deployment of multiple antennas to meet today's and tomorrow's needs for high-performance multimedia services on the move. Since high-performance multi-antenna transceivers must account for the wireless interface, we proceed to overview briefly the characteristics and detrimental effects of fading that have to be mitigated at the design phase. Subsequently, we classify wireless fading channels according to the coherence and selectivity they exhibit in diversifying waveforms transmitted in the time, frequency, and space dimensions. This diversification can be exploited to combat fading effects; and depending on the channel type, diversity appears in the multipath, Doppler, and/or space domains.

Transmission and reception by multiple antennas model, respectively, the excitation and response of multi-input multi-output (MIMO) channels, which provide spatial diversity in the form of transmit-diversity or receive-diversity when multiple antennas are deployed, respectively, at the transmitter or receiver. Both forms of spatial diversity are particularly attractive because at the cost of deploying multiple antennas, they do not necessarily incur loss in spectral efficiency and power efficiency or a considerable increase in complexity. In this chapter we outline the benefits of receive-diversity and allude to the challenges emerging with transmit-diversity systems which motivate the need for judicious transmitted waveform designs through the use of ST coding. Based on intuitive arguments, we further describe two major innovations that ST coded multi-antenna systems demonstrate over their single-antenna counterparts. We close this chapter with a road map of the book's contents.

1.1 EVOLUTION OF WIRELESS COMMUNICATION SYSTEMS

The history of wireless communications is relatively short. In fact, the debut of wireless communications could be dated back to 1901, when the first telegraph was sent across the Atlantic Ocean from Cornwall to St. John's Newfoundland [192]. However, cellular wireless communications have experienced unprecedented growth over the past thirty years and have already revolutionized the way that we communicate and live. Thus far, cellular wireless communications have evolved through three generations.

- **1G systems**

  First-generation (1G) cellular systems were introduced between the late 1970s and the early 1980s. All 1G systems were analog and were designed for narrowband voice services. The multiple-access technology used in 1G systems is frequency division multiple access (FDMA). Examples of 1G cellular systems include the Advanced Mobile Phone System (AMPS) in North America, the European Total Access Communications System (ETACS), and the Nippon Telephone and Telegraph (NTT) system in Japan.

- **2G systems**

  Second-generation (2G) cellular systems were deployed in the early 1990s, and are still in service in most countries. 2G systems are all-digital and employ either time division multiple access (TDMA) or code division multiple access (CDMA). Relative to the 1G systems, 2G systems offer better spectral efficiency, enhanced system capacity, and improved quality-of-service. Examples of 2G cellular systems include the Global System for Mobile Communication (GSM) in Europe, the Personal Communication Service (PCS) IS-95 system in North America, and the Personal Digital Cellular (PDC) system in Japan.

- **3G systems**

  Although data services are offered by 2G systems, they are predominantly narrowband. Driven by the growing demand for broadband wireless services such as high-speed wireless Internet access, wireless television and mobile computing, cellular wireless communication systems are now evolving into their third generation (3G) under the names International Mobile Telecommunications 2000 (IMT-2000) and CDMA-2000. IMT-2000 employs wideband direct-sequence CDMA (DS-CDMA), and CDMA-2000 is based on multi-carrier CDMA (MC-CDMA). Compared to 2G systems, 3G systems are capable of supporting much higher transmission rates and user mobility. 3G systems are currently under development and are designed to support transmission rates up to several megabits per second at end-user speeds as high as a few hundred kilometers per hour. Compared to 1G and 2G systems, the emphasis in 3G is on high-quality broadband multimedia services. Fourth-generation (4G) systems are on the horizon and are envisioned to offer broadband services with even higher data rates and user mobility.

The ultimate goal of wireless communications is to provide "anywhere, anytime, anymedia" wireless access at a reasonably low price. How to achieve this ambitious goal with limited bandwidth and power resources at affordable complexity while adhering to various implementation constraints brings about tremendous challenges to researchers and developers. The contents of this book provide a step toward addressing these challenges through ST coding techniques for use by broadband wireless mobile systems with multiple antennas deployed at the transmitter and/or receiver side.

Before we move on to argue as to the potential benefits that multi-antenna systems have over their single-antenna counterparts, it is instructive to overview the main source of these challenges, which can be traced back to the only element the designer does not have full control over: the wireless propagation channel.

1.2 WIRELESS PROPAGATION EFFECTS

When designing high-performance wireless systems, it is important to understand the challenges posed by wireless propagation. To this end, we summarize in this section the basic factors influencing propagation over wireless point-to-point links and their effects on the transmitted waveforms.

- **Path loss**

  It is well known that as electromagnetic waves propagate in free space with the speed of light, their magnitude decays with distance. Specifically, a waveform transmitted with a certain power from one point is received at a distance d with power proportional to $1/d^n$, where the exponent n depends on terrain contours and the environment (urban versus rural and outdoor versus indoor). For typical wireless links, n ranges between 2 and 4, whereas values of 4 to 6 are encountered within stadiums, buildings and other indoor environments. The reason for this loss is simple. Even if propagation encounters no obstacle, the radius of the transmitted spherical wave grows with distance. And since the energy is fixed, the received power at any point on the sphere is reduced along with the square of the distance. Besides distance, the power decay also depends on the wavelength, the transmit-antenna gain, the propagation medium itself, and the receive-antenna gain. As far as path loss is concerned, the harshness of the propagation medium constitutes the major difference between wireline and wireless communications and explains why the exponent n exceeds 2 in wireless links [242]. The obvious practical implication is that designers of wireless communication systems have to pay more attention to power efficiency.

- **Shadowing, reflecting, diffracting, and scattering**

  Even though waveforms travel along straight lines in free space, wireless communications do not take place in free space. If, on the other hand, the straight line between transmitter and receiver is "obstacle-free," line-of-sight (LOS) wireless communication is possible. When obstacles are present, whether wireless transmissions can penetrate them depends on the operating carrier frequency; low-frequency transmissions are easier to penetrate. The higher the frequency, the closer the transmitted waveforms resemble light in their propagation characteristics. Thus, even small obstacles such as a wall or a tree may block wireless transmissions, as they block light. This severe form of attenuation is called the shadowing effect. When the size of obstacles is much larger than the transmitted wavelength, reflections occur. Since the obstacles absorb part of the incident power, the reflected signal has reduced power. If the size of an obstacle is on the order of the transmitted wavelength or less, then scattering takes place (i.e., the incoming signal is scattered into a bunch of weaker waveforms). At obstacle edges in particular, diffraction happens pretty much as with light. Reflected, scattered, and diffracted renditions of the transmitted waveform can cause severe fading of the wireless link; but when constructively combined, they can facilitate wireless communications, especially when LOS is not available.

- **Multipath propagation**

  Whether LOS is absent or present to allow for the direct transmitter-receiver path, a waveform propagating through the wireless interface may be reflected, scattered, or diffracted before reaching the receiver through various indirect paths. Traveling over the direct and all indirect paths constitutes what is known as multipath propagation. As the transmitted waveform travels through all these paths, multiple versions of it reach the receiver at different times, since the speed of light is finite and multiple paths have variable length. This is a direct manifestation of multipath propagation and causes what is known as time dispersion or delay spread of the transmitted waveform. The delay spread deforms a narrow pulse into a broader one. As a result, one symbol waveform may spill over adjacent symbol waveforms, inducing what is called inter-symbol interference (ISI). But even when waveforms propagating through different paths arrive at the receiver almost simultaneously, they may differ in amplitude, phase, or carrier frequency offset. These differences may add constructively or destructively to amplify or attenuate the received power.

- **Fading effects**

  While multipath induces delay spread, the situation becomes even worse if the propagation medium changes or if transmitter and receiver are in relative motion, which is common in wireless mobile communications. In these cases, the power of the received waveform changes considerably over time because the transmitted waveform experiences time-varying paths. Experimental measurements at the output of real-world wireless channels confirm that instantiations of the receive-power over time resemble realizations of a random process. As power fluctuates randomly, it may approach vanishing levels, in which case we say that the wireless channel is fading. The random nature of the wireless fading channel constitutes its major difference with the wireline channel, which is modeled as deterministic. In general, fading effects caused by wireless propagation can be roughly categorized in three scales: (i) large-scale path loss effects modeled through an envelope that decays with distance; (ii) large- to medium-scale slowly varying shadowing effects modeled by a random channel amplitude following log-normal distribution; and (iii) small-scale fast-varying effects modeled as a random channel amplitude adhering to a Rice distribution if LOS is present, to a Rayleigh distribution if LOS is absent, or more generally, to a Nakagami distribution which can approximate a Rayleigh or Rice distribution and also capture additional fading effects [236, 242].

The brief account of all these propagation characteristics testifies to the harshness of wireless point-to-point links and illuminates the challenges that wireless systems face relative to their wireline counterparts. Detailed treatment of fading effects encountered with single-antenna links can be found in standard monographs devoted to wireless communication systems and their performance analysis [214, 218, 236, 242].

The multi-antenna wireless systems considered in this book comprise a collection of point-to-point wireless links, each corresponding to one pair of transmit-receive antennas. The fading effects summarized in this section are thus also present in the multi-antenna channel [97, 263]. In addition, depending on how the distance between neighboring antenna elements compares with the wavelength, the multiple channels may exhibit variable degrees of correlation. The latter will also affect the performance of multi-antenna wireless systems [87, 136, 235].

But before we proceed to discuss the challenges and potential benefits brought by multi-antenna channels, it is instructive to highlight the key parameters that define even single-antenna wireless channels. Based on these parameters and the time-bandwidth extent of the transmission, we will be able to classify wireless channels as we summarize next.

1.3 PARAMETERS AND CLASSIFICATION OF WIRELESS CHANNELS

In this section, we introduce key parameters characterizing the wireless fading channel. These parameters affect two basic features of random channels: coherence and selectivity. On a per realization basis, coherence pertains to the extent over which the channel's impulse or frequency response remains approximately constant, whereas selectivity refers to variation in the channel's impulse or frequency response. As an example, Figure 1.1 depicts coherence segments and selectivity features in a realization of the wireless channel gain varying across time, frequency, or space.

Besides controlling coherence and selectivity, the four major parameters we specify next lead to a natural classification of wireless fading channels that we present subsequently.

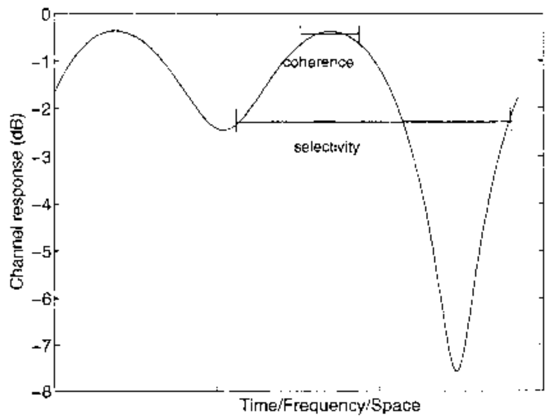

Figure 1.1 Coherence vs. selectivity

1.3.1 Delay Spread and Coherence Bandwidth

Let us consider a linear time-invariant random channel with impulse response $h(t;\tau) = h(\tau), \forall t$. With $\star$ denoting convolution, the channel output $y(t)$ is related to the channel input $x(t)$ via $y(t) = (h \star x)(t)$. If through this wireless channel we transmit a pulse of infinitesimal width (delta-like input) and what we receive at the channel output spreads over τ_d seconds, we say that the channel is time-dispersive with delay spread τ_d. In terms of the impulse response realization $h(\tau)$, the delay spread is defined as the lag τ_d for which $h(\tau) = 0, \forall \tau > \tau_d$. In the ensemble sense, we can let $\phi_h(\tau) := E[|h(\tau)|^2]$ denote the variance of the nonstationary channel process with respect to (w.r.t.) τ, which we assume to be zero-mean, and define the delay spread as the lag τ_d for which $\phi_h(\tau) = 0, \forall |\tau| > \tau_d$. For average error performance analysis, the ensemble definition is useful. However, the sample definition of τ_d is also useful since it is necessary to cope with the delay spread in each realization.

The finite nonzero support of $h(\tau)$ and $\phi_h(\tau)$ is only approximately true, in general, and is satisfied by the popular tap-delay line model with a finite number of taps [214]. We can also interpret τ_d through the clusters of rays arriving coherently to form each tap (path) in the multipath propagation model discussed in Section 1.2. In this case, τ_d represents the maximum or the root-mean-square (rms) of the cluster arrival times, which correspond to the relative delays among paths [218].

Clearly, if the pulse used to transmit every information-bearing symbol has duration T_p and the period with which symbols are transmitted is T_s, the duration of each received symbol waveform through the channel $h(\tau)$ with delay spread τ_d will be $T_p + \tau_d$, and overlap among received pulses will be avoided if $T_s \geq T_p + \tau_d$. If on the other hand, $T_s < T_p + \tau_d$, then received symbol pulses will overlap, giving rise to ISI.

It is also useful to consider the effect of ISI channels in the frequency domain. The Fourier transform of $\phi_h(\tau)$ defines the channel's power spectral density (PSD)

$\Phi_h(f)$, whose nonzero support (bandwidth B_h) is infinite since $\phi_h(\tau)$ has finite support. Of interest is the coherence bandwidth $B_{\text{coh}} < B_h$, which is defined as the frequency spread over which the channel PSD can be considered approximately flat. The uncertainty principle asserts that frequency spread and delay spread must be inversely proportional:

$$B_{\text{coh}} \approx \frac{1}{\tau_d}.$$

To appreciate where the term coherence comes from, it is useful to recall that two waveforms, $x(t)$ and $y(t)$, are called coherent if they are equal up to a (generally complex) scale; i.e., if $y(t) = \alpha x(t)$. Notice that if these two coherent waveforms are realizations of two stationary random processes, their cross-correlation has constant magnitude and their correlation coefficient (a.k.a. coherence) has magnitude one. Let us also recall that according to its Fourier expansion, a waveform consists of complex exponentials. Consider one such exponential with frequency f_0 as the channel input. If f_c denotes the carrier frequency and $f_0 \in [f_c - B_{\text{coh}}/2, f_c + B_{\text{coh}}/2]$, the channel output $y_0(t)$ is coherent with the channel input $x_0(t) = A_0 \exp(j2\pi f_0 t)$ because complex exponentials are eigenfunctions of linear time-invariant channels and thus $y_0(t) = H(f_0)x_0(t)$, where $H(f_0)$ is the channel's frequency response at f_0. If the transmitted waveform consists of complex exponentials with frequencies within the coherence bandwidth of the channel, all these exponentials will be affected by basically the same $H(f_0)$ since $H(f)$ is approximately flat over B_{coh}; and thus the channel output will be coherent with the channel input. This explains why B_{coh} is called the coherence bandwidth.

On the other hand, if the input contains frequencies outside B_{coh}, then the scale $H(f_0)$ will generally be different for different f_0's and the transmitted waveform will not be coherent with the received one. In this case, where the transmission bandwidth exceeds the channel's coherence bandwidth, we say that the channel exhibits frequency-selectivity since it affects selectively the various "frequency components" of the channel input.

In a nutshell, how time-dispersive a linear time-invariant wireless channel is can be assessed in the time(lag) domain by the delay spread τ_d of its impulse response; and depending on how the delay spread compares with the symbol period and the duration of the symbol pulse, decides the severity of the ISI. In the frequency domain, the amount of time dispersion and ISI correspond to the degree of frequency selectivity, which is quantified by the channel's coherence bandwidth B_{coh}, and depending on how the latter compares with the transmission bandwidth decides how pronounced the frequency selectivity is.

1.3.2 Doppler Spread and Coherence Time

Suppose now that the wireless channel is linear but time-varying and non-dispersive in time with impulse response $h(t;\tau) = h(t), \forall\tau$. With $x(t)$ and $y(t)$ denoting as before the channel input and output, such a channel obeys a multiplicative input-output relationship: $y(t) = h(t)x(t)$. Using as input a finite-duration sinusoidal waveform with infinitesimal bandwidth (delta-like in the frequency domain), the time-varying channel $h(t)$ will disperse the bandwidth of this input waveform by an amount B_d, pretty much as frequencies shift due to the Doppler effect when a source emitting harmonic signals is moving. Such a channel clearly causes frequency dispersion of its input waveforms by an amount B_d that we naturally call Doppler spread.

In a dual fashion to the time-dispersive channel, if $H(\nu)$ denotes the Fourier transform of a frequency-dispersive channel $h(t)$ over an infinite time horizon, the Doppler spread is defined as the frequency for which $H(\nu) = 0, \forall|\nu| > B_d$. As before, for a realization-independent definition, we can let $\phi_h(t) := E[h(t_1)h^*(t_1 + t)]$ denote the autocorrelation of the channel process (* denotes conjugation), which is assumed zero-mean stationary w.r.t. t_1, and define the Doppler spread as the frequency B_d for which the PSD $\Phi_h(\nu) = 0, \forall|\nu| > B_d$. In general, the finite nonzero support of $H(\nu)$ and $\Phi_h(\nu)$ is only approximately true and is satisfied by the finite-parameter basis expansion model introduced in Chapter 9.

If information symbols are transmitted on successive frequency bands each with bandwidth B_p and successive bands are B_s Hertz (Hz) far apart from each other, each received symbol waveform will have bandwidth $B_p + B_d$ and the overlap in the frequency domain will be avoided if $B_s \geq B_p + B_d$. But if $B_s < B_p + B_d$, the received symbol bands will overlap in the frequency domain giving rise to what one could term Doppler intersymbol interference (D-ISI).

As with ISI, it is useful to consider how D-ISI manifests itself in the time domain. Clearly, since $\Phi_h(\nu)$ is bandwidth-limited, $\phi_h(t)$ has infinite support in the time domain. However, it is of interest to define the finite time horizon $T_{\rm coh}$ over which $h(t)$ and thus $\phi_h(t)$ can be considered as approximately constant functions of time. This time interval is called coherence time and in accordance with the uncertainty principle, it is inversely proportional to the Doppler spread:

$$T_{\rm coh} \approx \frac{1}{B_d}.$$

Clearly, an input pulse $x_0(t)$ of infinitesimal duration centered at t_0 will result in a coherent channel output $y_0(t) = h(t_0)x_0(t)$. And if the symbol waveform consists of such pulses with $t_0 \in [0, T_{\rm coh}]$ the entire received symbol waveform will remain coherent with the transmitted waveform since $h(t)$ is flat over $T_{\rm coh}$ seconds. This explains why $T_{\rm coh}$ is called coherence time. Except for cases of high mobility, $T_{\rm coh}$ typically spans a number of symbol periods, allowing the channel to be considered as time-invariant over a time horizon of multiple symbols. Certainly, if the channel input $x(t)$ is considered over a duration exceeding $T_{\rm coh}$, the scale $h(t_0)$ will generally be different for different t_0's, and the transmitted waveform will not be coherent with the received one. In this case we say that the channel exhibits time selectivity since it affects selectively the various "time components" of the channel input.

To summarize: How frequency-dispersive a linear time-varying wireless channel is can be assessed in the frequency domain by the Doppler spread B_d of its frequency response; and depending on how the Doppler spread compares with the symbol rate and the bandwidth of the transmitted waveform, decides the severity of D-ISI. In the time domain, the amount of frequency dispersion and of D-ISI correspond to the degree of time selectivity, which is quantified by the channel's coherence time

$T_{\rm coh}$, and depending on how the latter compares with the duration of the transmitted waveform decides how pronounced the time selectivity is.

A wireless channel can generally be flat, individually time- or frequency-selective, but also jointly time- and frequency-selective. However, with the $B_{\rm coh}$ and $T_{\rm coh}$ of a channel given, which form of selectivity is present and to what degree depends on the transmission bandwidth and the time window over which the channel input (and thus the channel output) is considered. Based on the size of this time-bandwidth extent, the wireless channels considered in this book fall into four categories, depending on the amount of time selectivity and frequency selectivity they exhibit.

- **Quasi-static flat and block fading channels**

  The random channels in this class are linear time-invariant (i.e., not time-selective) and also not frequency-selective; i.e., both ISI and D-ISI are absent over the time interval and bandwidth considered. This occurs when (i) the transmission bandwidth is smaller than the coherence bandwidth, or equivalently, the symbol period is larger than the delay spread plus the duration of the transmitted symbol waveform; and (ii) the time horizon considered is smaller than the coherence time of the channel. Their impulse response is flat (a random constant complex-valued coefficient) in both the time and the lag variable; i.e., $h(t;\tau) = h$. Because a random constant is a non-ergodic[1] process, these channels are considered as time-invariant only over a finite time horizon. Thus, they are quasi-static in the sense that their impulse response is a random constant assuming a certain value over a coherence time interval of duration $T_{\rm coh}$, but may change from one coherence interval to the next. Quasi-static flat fading channels considered over multiple blocks of size exceeding $T_{\rm coh}$ are called block fading channels.

- **Time-invariant frequency-selective fading channels**

  In this class, the channels are linear time-invariant (i.e., not time-selective) over the time horizon considered, but they exhibit frequency-selectivity; i.e., ISI is present but D-ISI is absent. This happens when (i) the transmission bandwidth exceeds the coherence bandwidth, or equivalently, the symbol period is smaller than the delay spread plus the transmit pulse duration; and (ii) the time window considered is within the channel coherence time. The class of frequency-selective channels arises when the mobility is relatively low but the data rate is high, as in broadband fixed wireless applications. As we mentioned earlier, these channels are modeled using a tapped-delay line with a finite number of complex tap coefficients [214]. The values of these coefficients remain constant over intervals of size $T_{\rm coh}$, but they may change from one coherence interval

[1] Non-ergodic channels yield non-ergodic received processes. Since detection and estimation tasks at the receiver must be performed on a per realization basis, sample statistics can be used instead of ensemble statistics only if the received process is ergodic; for example, to obtain the autocorrelation sequence or the average capacity from a single realization, the received process must be ergodic.

to the next. These channels are special cases of their multivariate counterparts dealt with in Chapters 7, 8, and 11.

- **Time-varying time-selective fading channels**

  This class of channels is dual to the preceding one. They are time-selective over the time horizon under study but not frequency-selective over the bandwidth considered. As a result, they induce D-ISI but not ISI. This class arises when (i) the time interval under consideration exceeds the channel coherence time; and (ii) the transmission bandwidth is within the channel's coherence bandwidth, or equivalently, the symbol period (minus the pulse duration) is larger than the channel delay spread. The class of time-selective but frequency-flat channels appears when mobility is high but data rates are not very high. They are modeled using the basis expansion model, which is discussed in Chapter 9.

- **Doubly selective fading channels**

  In the most general and challenging case, wireless channels can be both time- and frequency-dispersive. Since these channels can exhibit both frequency- and time-selectivity, they are often referred to as doubly selective channels. Depending on the delay spread and Doppler spread, a doubly selective channel may exhibit variable degrees of ISI and D-ISI. Clearly, the preceding two classes are subsumed as special cases of the doubly selective class. Both forms of selectivity emerge when: (i) the time horizon under consideration exceeds the channel coherence time; and (ii) the transmission bandwidth exceeds the channel coherence bandwidth. Channels in this class appear in many applications since they encompass cases where both data rates and mobility are high. They can be modeled using a tapped-delay line with time-varying taps whose variations adhere to the basis expansion model detailed in Chapter 9.

- **Multi-input multi-output fading channels**

  Multi-input multi-output (MIMO) fading channels are multivariate generalizations of the single-input single-output (SISO) random channels considered so far. In fact, a MIMO channel is a collection of SISO channels, each of which can fall into any of the four classes we discussed earlier, depending on the selectivity it exhibits in the time and/or the lag variable of its impulse response. MIMO channels arise in various scenarios: with single-user single-antenna systems entailing block transmissions, with multi-user single-antenna systems, and with multi-antenna systems in point-to-point or multi-access communications. Since multi-antenna communication systems provide the context of this book, MIMO fading channels will be encountered throughout. Multi-antenna transmitter and receiver designs will be sought for operation over quasi-static flat fading as well as frequency- and time-selective MIMO channels. Besides time and frequency selectivity, the multiple antennas deployed at the transmitter and/or receiver ends can provide an additional form of selectivity: space selectivity. Furthermore, transmitting over multiple antennas induces extra interference at the receiver side.

The most general wireless channel is a doubly selective MIMO channel, where ISI, D-ISI, and interference in space can be present simultaneously. For these sources of interference to be removed prior to demodulation, the doubly selective MIMO fading channel must be estimated at the receiver. Channel estimation and demodulation can be challenging due to the "curse of dimensionality." But the presence of interference in lag, Doppler, and/or spatial dimensions can be exploited to the designer's advantage. This is possible because over space and over a time-bandwidth window, these sources of interference provide multiple renditions of the information symbol, which can be used to mitigate fading effects by combining these "diversified" transmissions coherently.

As we will see soon, this diversification can improve the performance of MIMO communications through techniques that enable (at the transmitter) and collect (at the receiver) the diversity that these channels are capable of providing. The different flavors of diversity and diversity techniques that can be used to turn the "curse" into a "blessing" are overviewed in the next section.

1.4 PROVIDING, ENABLING, AND COLLECTING DIVERSITY

We define and quantify diversity analytically in Chapter 2. Our goal in this section is to provide an intuitive explanation of diversity and outline techniques for enabling and collecting the diversity to improve rate and error performance of wireless communications over fading channels. Diversity is usually defined in the literature as the number of independent (or at least uncorrelated) copies of the information-bearing signal available at the receiver and is often attributed to operations such as channel coding, interleaving, or frequency hopping performed at the transmitter. This is at times confusing since diversity indeed amounts to the signal copies provided by the channel, but these copies do not have to be independent or uncorrelated. Furthermore, diversity is provided inherently by the channel. The transmission scheme can only enable the diversity provided by the channel (or part of it), while the receiver processing can collect the diversity (or part of it). We can enable only part of the diversity with a suboptimum transmitter and also collect part of it with a suboptimum receiver but can never achieve more diversity than what the channel can provide us with.

Keeping these remarks in mind, we consider next the flavors of diversity provided by the various classes of channels described in Section 1.3. For each class we outline transceiver paradigms to enable and collect the channel diversity.

1.4.1 Diversity Provided by Frequency-Selective Channels

When ISI is present, received symbol waveforms overlap in the time domain, which implies that each symbol is replicated multiple times (each scaled with a different channel tap) in any time window exceeding the channel delay spread in size. These symbol replicas can be combined at the receiver to improve the instantaneous signal-

Figure 1.2 Multipath (or frequency) diversity enabled via coding and frequency hopping

Figure 1.3 Multipath (or frequency) diversity enabled via coding and multi-carrier modulation

to-noise-ratio (SNR) and in turn, the average error performance. Intuitively speaking, this combination adds degrees of freedom to the probability density function (pdf), which renders the pdf of the combined SNR "less fading."[2] In the case of frequency-selective channels, these degrees of freedom are related directly to what we will henceforth call multipath diversity.

A synonymous term often used in place of multipath diversity is frequency diversity. We prefer the term multipath diversity because it captures better the cause behind the creation of symbol replicas. The term frequency diversity refers to the enabling transmission that can take advantage of the frequency selectivity and can be explained if we consider the frequency response $H(f)$ of an ISI channel $h(\tau)$. Suppose that we transmit the same symbol on multiple carriers that are pairwise separated in the frequency domain by at least as much as the coherence bandwidth of the channel. Each symbol replica on a different carrier will then experience a different fading coefficient and combining the replicas at the receiver, using, e.g., a maximum ratio combiner (MRC), will improve the SNR and thus the average error performance.

Transmitting the same symbol on multiple carriers can be implemented either by periodic interleaving or by frequency hopping with a hop interval at least as large as the coherence bandwidth. Furthermore, instead of repetition we can use more powerful channel codes prior to frequency hopping. A system combining channel coding with frequency hopping to enable the multipath (or frequency) diversity of an ISI channel is depicted in Figure 1.2. Channel coding can also be combined with multicarrier modulation as illustrated in Figure 1.3. Clearly, even if fading nulls a number of bits in each channel codeword, the remaining bits, which are likely to experience less severe fading, can lead to reliable recovery of the information-bearing symbol. Maximum likelihood (ML) decoding at the receiver can certainly collect the maximum available diversity. However, the available diversity (or part of it) can be collected even with suboptimum detectors.

[2] When replicas are received through, e.g., uncorrelated complex Gaussian fading channels, the pdf of the combined SNR is chi-square with as many degrees of freedom as the number of replicas. We will see in Section 2.6 that this combined pdf is "less fading" than that of individual pdfs because it is smoother around the origin.

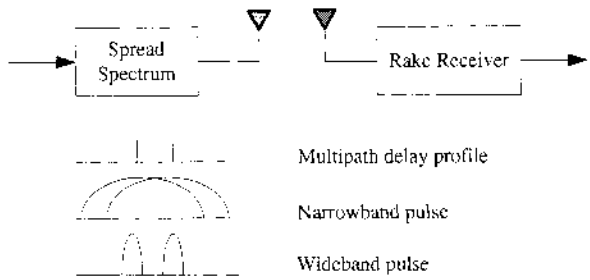

Figure 1.4 Multipath (or frequency) diversity enabled via SS transmission and collected using Rake reception

The amount of multipath diversity provided by an ISI channel depends on the degree of frequency selectivity it exhibits. The more pronounced that frequency-selectivity is, the higher multipath diversity it provides and this diversity can be enabled by shorter codewords, which in turn imply smaller decoding delay and lower complexity at the receiver. On the other hand, frequency selectivity depends on the transmission bandwidth. This means that even flat fading channels can be rendered frequency-selective if we increase the transmission bandwidth. Spread-spectrum (SS) transmissions aim at precisely this objective. Relying on spreading codes with good autocorrelation properties, a Rake receiver can effectively combine the multipath components and collect the available multipath diversity, as in the system depicted in Figure 1.4.

A recent approach to enabling multipath diversity is through the use of linear complex field coding (a.k.a. linear precoding). As we will see in Chapters 7, 8, and 10, this approach does not introduce redundancy and nicely complements the use of redundant channel coding in combating frequency-selective fading effects.

1.4.2 Diversity Provided by Time-Selective Channels

When D-ISI is present, symbol waveforms in the Doppler domain overlap, which means that over a range of Doppler frequencies a symbol can appear multiple times. For this reason, although this form of diversity is often referred to as time diversity, a more appropriate term which we use in this book, is Doppler diversity since signal copies appear in the Doppler domain. The potential benefits of time selectivity can also be appreciated from the input-output relationship in the time domain. If $h(t)$ varies over a symbol duration, it is likely that it will not stay in a deep fade over the duration of the entire symbol waveform, thus allowing for symbol recovery.

Recall that even if a channel is flat over a time window (e.g., of duration equal to one symbol period), it can become time-selective if we expand the time window under consideration. One simple approach to enabling the diversity provided in this case is by repeating the same symbol with period at least equal to the channel coherence time. Knowing the channel at the receiver end, we can employ an MRC

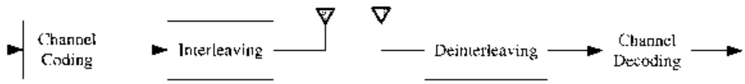

Figure 1.5 Doppler (or time) diversity enabled via channel coding and interleaving

receiver to combine the symbol replicas, enhance the SNR, and thus reduce the average probability of error due to fading. Clearly, periodic insertion of the same symbol can be implemented through an interleaving operation with depth equal to the period of insertion. Furthermore, since channel codes more powerful than repetition are available, it is possible to enable the diversity provided by a time-selective channel using channel coding and interleaving, as depicted in Figure 1.5. At the receiver end, the Doppler diversity can be collected if after deinterleaving, an ML detector is used for decoding; e.g., via Viterbi's algorithm if convolutional coding is employed. The intuitive reason why this approach can improve error performance is that even if fading nulls a number of bits in each codeword, the remaining bits can allow for reliable recovery of the information-bearing symbol.

Notice that coding followed by interleaving can improve error performance even when the channel is quasi-static (block fading), provided that the interleaver depth is larger than the channel coherence time. This is because a flat channel considered over an expanded time horizon can be rendered time-selective and thus provide Doppler diversity. Of course, the faster a channel varies, the higher Doppler diversity it provides, and the transmitter can enable it with an interleaver of smaller depth. This is important because as the interleaver depth increases, the decoding delay and complexity at the receiver side may become intolerable.

Decoding delay and complexity also depend on the amount of redundancy introduced by the channel code. It is possible, however, to eliminate this redundancy if instead of encoding over the Galois field, we encode symbols over the complex field. We will see in Chapter 9 that linear complex field coding offers an alternative means of enabling the Doppler diversity provided by time-selective channels. Galois field codes, on the other hand, offer higher coding gains than do complex field codes. These considerations suggest their joint use over fading channels as detailed in Chapter 10.

As the reader might have already guessed, a general doubly selective channel can provide both multipath and Doppler diversity. In fact, we will see in Chapter 9 that since the degrees of freedom in these general channels equal the product of those provided by time and frequency selectivity, the resulting multipath-Doppler diversity effects are multiplicative. To enable this large amount of diversity, one has to rely on channel coding and/or precoding applied both across frequency (or multipath components) and across time (or Doppler components). Furthermore, to make a quasi-static flat fading channel exhibit frequency and time selectivity, we should clearly increase the transmission bandwidth beyond the channel's coherence bandwidth and also expand the time window at the receiver to exceed the channel's coherence time. By expanding both time and bandwidth, complexity at the receiver will increase along with decoding delay while spectral efficiency will drop considerably, especially if Galois field coding is adopted.

One form of selectivity not requiring time or bandwidth expansion is the one provided by multi-antenna channels, which we consider next.

1.4.3 Diversity Provided by Multi-Antenna Channels

With multiple antennas deployed at the transmitter and/or at the receiver end, the resulting MIMO fading channel possesses degrees of freedom providing what we term space diversity. At the cost of deploying extra antennas even when the MIMO channel is flat, space diversity can improve error performance without increasing the transmission bandwidth to induce frequency selectivity or expanding the observation window to effect time selectivity. Certainly, if these forms of selectivity are already available by the MIMO channel, coded or precoded multi-antenna systems can provide the combined (multiplicative) form of multipath, Doppler, and space diversity.

Space diversity also comes in two flavors: transmit- and receive-antenna diversity. Receive-antenna diversity has been well studied for decades; transmit-antenna diversity has received increasing attention since the late 1990s.

1.4.3.1 Receive-Diversity SIMO Systems Figure 1.6 depicts a receive-diversity system with N_r receive-antennas and a single transmit-antenna giving rise to a single-input multiple-output (SIMO) channel. At each receive-antenna we obtain a copy of the transmitted symbol, s, denoted as $y_i = h_i s + v_i$, where h_i is the flat fading coefficient corresponding to the channel linking the transmit-antenna with the ith receive-antenna; and v_i denotes additive white Gaussian noise (AWGN). Even intuitively it is evident that with the same transmit power, the N_r receive-antennas collect N_r times more power than does a single receive-antenna. This clearly increases the instantaneous receive SNR at the output of a combiner that can be invoked to demodulate the symbol s using a weighted superposition of the received symbols (with corresponding weights w_i) as

$$\hat{s} = \sum_{i=1}^{N_r} w_i y_i = \sum_{i=1}^{N_r} w_i h_i s + \sum_{i=1}^{N_r} w_i v_i.$$

Commonly used combiners include the MRC, where $w_i = h_i^*$, the equal-gain combiner (EGC), where $w_i = h_i^*/|h_i|$, or the selective combiner (SC), where $w_j = 1, j = \arg\max_i |h_i|$, and $w_i = 0, \forall i \neq j$. These three combiners present different trade-offs among required channel knowledge, complexity, and average error probability achieved; but they all collect the maximum receive-antenna diversity even when the channels h_i are correlated and have different pdfs [289]. Again, the underlying reason is that the pdf of the combined SNR at each combiner's output possesses the sum of the degrees of freedom of each individual branch's pdf, which renders it smoother around the origin and thus more resilient to fading effects [289].

A receive-diversity system with multiple receive-antennas and a single transmit-antenna is well suited for the uplink since the access point (base station) can afford the deployment of multiple antennas. However, it may not be a good fit for the downlink since packing multiple antennas at mobile stations may not be cost-effective, and

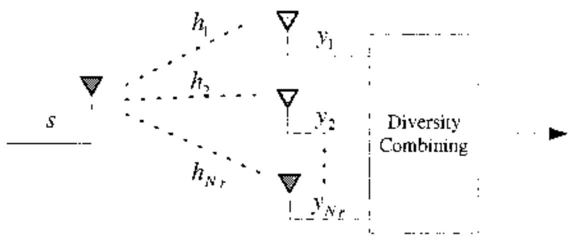

Figure 1.6 Space diversity provided by multiple receive antennas

certainly it is less feasible as the size of the handsets decreases. A more attractive solution for the downlink is to deploy multiple transmit-antennas to enable transmit-antenna space diversity.

1.4.3.2 Transmit-Diversity MISO Systems In a transmit-diversity system, the signal copies providing the diversity come from channels linking multiple (say, N_t) transmit-antennas to a single receive-antenna. However, different from receive-diversity systems, where the power available for demodulation can increase by a factor equal to the number of receive-antennas (N_r), the transmit-power in a transmit-diversity system has to be divided by the N_t transmit-antennas, while the receive-power is identical to that of a SISO system. Nonetheless, it will turn out that even a multi-input single-output (MISO) channel can provide degrees of freedom that can be used at the receiver to collect the available space diversity if the transmitted signals are also designed properly. Collecting the diversity provided by MISO channels is not as easy as in the SIMO case since the signals originating from the multiple transmit-antennas are superimposed at the receive-antenna. But even more challenging is the design of transmissions enabling the available space diversity.

To illustrate the importance of signal design in enabling the available transmit-diversity of MISO channels, let us consider the example system depicted in Figure 1.7, with two transmit-antennas and a single receive-antenna. Assuming flat fading channels where the two channel coefficients h_1 and h_2 are modeled as two zero-mean uncorrelated complex Gaussian random variables with unit variance, the received symbol can be expressed as

$$y = h_1 \frac{1}{\sqrt{2}} s_1 + h_2 \frac{1}{\sqrt{2}} s_2 + v,$$

where the $1/\sqrt{2}$ scale is present since the transmit-power is divided equally between the two antennas.

Consider now repeating the same symbol in both antennas (i.e., $s_1 = s_2$), which yields the input-output relationship

$$y = \left(\frac{h_1 + h_2}{\sqrt{2}} \right) s_1 + v.$$

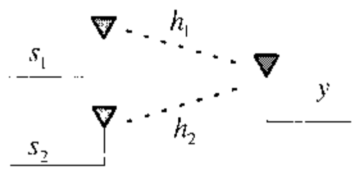

Figure 1.7 Space diversity provided by two transmit-antennas

Since h_1 and h_2 are uncorrelated Gaussian distributed, $(h_1 + h_2)/\sqrt{2}$ is distributed according to the same pdf as h_1. As a result, this multi-antenna system behaves like a SISO system, and no advantage can be enabled from the use of multiple transmit-antennas.

This simple example illustrates the need for smart signal designs in order to benefit from the transmit-diversity provided by a MISO fading channel. The main emphasis in this book is on signal designs that rely on what we will call ST coding techniques. ST codes presume the deployment of multiple transmit-antennas and will entail channel coding over the Galois field and possibly over the complex field in the form of spatial multiplexing across transmit-antennas.

1.4.3.3 Transmit/Receive-Diversity MIMO Systems For a more general system with multiple transmit-antennas and multiple receive-antennas, recent research has shown that ST coded transmissions over the resulting MIMO channel can effect diversity as high as the number of transmit-antennas times the number of receive-antennas ($N_t N_r$); see also Figure 1.8. As we will see in Chapter 3, this can provably enhance the average error performance of multi-antenna systems at sufficiently high SNR [6, 253, 289].

But furthermore, we show in Chapter 3 that the degrees of freedom that become available with a MIMO channel can boost the capacity (and thus data rates) of multi-antenna systems well beyond those available with single-antenna links over SISO channels [69, 256, 326]. This can be roughly guessed if we notice that with, for example, two transmit-antennas we can transmit two symbols simultaneously for every channel use. Critical to this enhanced rate performance is the use of ST coded transmissions over multiple transmit-antennas, which intuitively speaking, provide multiple "information pipes" for symbols to flow through the MIMO channel.

Besides error and rate performance, the design of ST coded multi-antenna transmissions must take into account a number of practical issues. One pertains to complexity, which becomes critical if ST coding is employed in the downlink, where the receiver can be a handset. Since a handset typically has limited power and size, one has to design ST coding schemes properly so that the corresponding receiver complexity is affordable. Another issue relates to data rate and mobility. As ST coded systems are envisioned to support high data rates possibly at high mobility, the MIMO fading channel is likely to be frequency- and/or time-selective. Hence, ST codes must be designed to account for these challenging propagation channels. Furthermore, ST

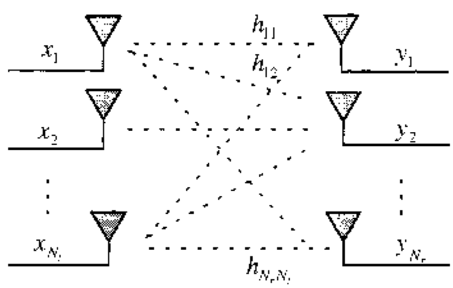

Figure 1.8 Space diversity provided by multiple transmit-receive antennas

coded systems should be robust, to handle various sources of interference in addition to fading and AWGN. These could include co-channel or intentional interference as well as multiuser interference in a multi-access setup. Finally, a major issue that must be taken into account when designing ST codes is how much knowledge of the channel state information (CSI) is available at which end of the link (receiver and possibly transmitter via, e.g., a feedback channel).

1.5 CHAPTER-BY-CHAPTER ORGANIZATION

Design of ST coded multi-antenna systems for operation over general broadband mobile wireless MIMO channels is the subject of this book. The ST designs we cover in the ensuing 14 chapters can be classified based on a variety of criteria, as we will see in Section 2.7. A common objective is to design systems flexible enough to reach desirable trade-offs among error performance, complexity, and spectral efficiency. As is often the case in practice, there is no universally "best ST code" and the designer's choice certainly depends on application-specific constraints. A road map of the contents in Chapters 2–15 is outlined next.

- *Fundamentals of ST Wireless Communications (Chapter 2)* In this chapter we describe fundamental capacity and error performance metrics pertaining to ST coded multi-antenna transmissions over quasi-static flat fading MIMO channels. Through these metrics it is established quantitatively that multi-antenna MIMO systems can improve considerably the capacity and error probability of SISO systems. After laying out a unifying system model for multi-antenna ST coded links, capacity and error performance bounds are presented to serve also as criteria for designing and analyzing ST coded systems — one based on the diversity gain and the other on the coding gain. Differences and similarities between channel coding and ST coding are presented, trade-offs between diversity and spatial multiplexing are delineated, and the definition of spectral efficiency used for comparing ST coded systems is specified.

- *Coherent ST Codes for Flat Fading Channels (Chapter 3)* In this chapter we present several classes of ST codes designed for quasi-static flat fading MIMO channels. The classes include ST trellis codes, orthogonal and quasi-orthogonal ST block codes, ST linear complex field codes, and unifying designs. They are coherent in the sense that they all require knowledge of the MIMO channel at their decoding stage. The designs aim to maximize the space diversity provided by the MIMO channel, but the resulting ST codes end up exhibiting different merits in rate, coding gain, and (de)coding complexity. Common to all these classes is that they improve error performance relative to their single-antenna counterparts, but in spectral efficiency they do not exceed rates achievable by SISO systems.

- *Layered ST Codes (Chapter 4)* The classes of coherent ST codes in this chapter aim to attain rates higher than those achievable by SISO systems, and they do so by ST coding and multiplexing groups of symbols which are called layers. These ST codes are capable of approaching the capacity gains that MIMO channels can provide. In this chapter, hybrid ST codes are further introduced to trade off diversity for spectral efficiency, thus bridging the rate-oriented layered ST codes with the classes of error performance-oriented ST codes of Chapter 3. Finally, ST codes are presented to achieve an interesting combination of full diversity at full spectral efficiency while being flexible enough to achieve desirable trade-offs in complexity, error, and rate performance.

- *Sphere Decoding and (Near-)Optimal MIMO Demodulation (Chapter 5)* This chapter deals with decoding of ST codes based on the so-called sphere decoding algorithm (SDA). SDA is useful for coherent demodulation of most ST codes designed in this book for flat, frequency-selective, time-selective, or doubly selective MIMO channels. Besides multi-antenna systems, SDA also finds applications in demodulation of single-antenna single-user block (coded or precoded) transmissions as well as in detection of multiuser transmissions. Zero-forcing, linear minimum mean-square error, decision feedback, and recent quasi-ML demodulators are also mentioned briefly in this chapter. But the emphasis on SDA is well justified because it outperforms these lower-complexity alternatives while being able to reach near-ML or exact-ML optimality for a number of SNR values and problem dimensions of practical interest. Surprisingly, SDA can achieve exact- or near-ML optimality at polynomial average complexity, which is approximately cubic in the problem dimension. (Recall that exact ML by enumeration incurs exponential complexity in the problem dimension.) In addition to hard decoding, this chapter shows how SDA can facilitate soft decoding of ST coded multi-antenna transmissions, which is necessary to approach the performance dictated by the enhanced capacity of MIMO channels.

- *Noncoherent and Differential ST Codes for Flat Fading Channels (Chapter 6)* In this chapter we present noncoherent and differential ST codes for quasi-static flat fading MIMO channels. As these two classes of ST codes do not require channel knowledge at their decoding stage, they are suitable for applications where channel estimation is impossible or cannot be afforded. They are also well motivated when the MIMO channel undergoes slow time variations. Non-coherent and differential ST codes have their own pros and cons, and one class could be preferred over the other, depending on application-specific trade-offs among error performance, complexity, and spectral efficiency. The theme of this chapter is to present the basic ideas behind these schemes, analyze their error performance, and compare them with their coherent counterparts.

- *ST Codes for Frequency-Selective Fading Channels: Single-Carrier Systems (Chapter 7)* The coherent ST codes designed in this chapter for single-carrier systems as well as those developed in Chapter 8 for multi-carrier systems are tailored for broadband wireless applications where frequency selectivity must be accounted for. Upon recognizing that the input-output relationship of a frequency-selective MIMO channel can be brought into an equivalent flat fading MIMO form, this chapter shows how to generalize the ST codes of Chapter 3 to enable the joint space-multipath diversity provided by frequency-selective MIMO channels. Besides diversity, these ST codes are compared on the basis of their outage capacity, and they all attain rates not exceeding those of a SISO channel.

- *ST Codes for Frequency-Selective Fading Channels: Multi-Carrier Systems (Chapter 8)* The ST codes in this chapter are also intended for frequency-selective MIMO channels but different from the ST codes in Chapter 7, they are designed for multi-carrier systems employing orthogonal frequency division multiplexing (OFDM) modulation. The classes considered include trellis codes, orthogonal block codes, and linear complex field codes based on digital phase sweeping (DPS) operations. All these classes of ST codes enable the maximum space-multipath diversity provided by the frequency-selective MIMO channel and relative error performance is assessed using their outage capacity. ST coded MIMO OFDM systems offer improved error performance relative to single-antenna OFDM systems at the same spectral efficiency. In addition, linear complex-field coded MIMO OFDM transmissions are designed in this chapter to enable maximum diversity at spectral efficiencies higher than those that SISO OFDM systems can attain. This feature, along with the lower-complexity that MIMO OFDM receivers can afford, favors multi-carrier multi-antenna systems over their single-carrier counterparts of Chapter 7, especially for MIMO channels with long delay spreads. The brief comparison between multi-carrier and single-carrier systems included at the end of this chapter gives an edge to single-carrier systems only when it comes to deciding on the basis of a low peak-to-average-power-ratio and insensitivity to carrier frequency offsets.

- *ST Codes for Time-Varying Channels (Chapter 9)* Mobility effects, drifts, and mismatch between transmit-receive oscillators all give rise to time-varying (TV) MIMO channels. To design ST codes for these channels, we rely on a parsimonious basis expansion model, based on which it becomes possible to establish a duality between time- and frequency-selective MIMO channels. Several classes of ST codes designed for frequency-selective channels are then mapped based on this duality, to obtain ST codes capable of enabling the joint space-Doppler diversity provided by time-selective MIMO channels. Furthermore, ST codes are designed in this chapter for doubly selective channels and are shown to enable the joint space-multipath-Doppler diversity. Besides coherent designs, differential ST codes are also developed.

- *Joint Galois-Field and Linear Complex-Field ST Codes (Chapter 10)* The reliability of MIMO fading links improves considerably with space-diversity but can at best reach the performance of uncoded transmissions over single-antenna AWGN channels, even if one can afford the complexity and cost of deploying multiple antennas. In this chapter we deal with concatenated Galois-field (GF) and LCF codes, which allow ST coded systems to approach the error performance dictated by the capacity of MIMO fading channels. In enabling the diversity, LCF codes turn out to be less complex and more spectrally efficient than GF codes; while GF codes bring larger coding gains than LCF codes, which is critical as the channel comes closer to an AWGN one. It is demonstrated that the combination of GF-LCF ST codes is powerful in dealing with MIMO fading channels when a turbo decoder with relatively low complexity is employed at the receiver. Besides flat fading MIMO channels, it is further shown that the GF-LCF ST system is applicable to time- and frequency-selective MIMO channels.

- *MIMO Channel Estimation and Synchronization (Chapter 11)* In this chapter we introduce carrier synchronization and MIMO channel estimation algorithms with universal applicability in decoding coherent ST coded transmissions. Along with preamble-based schemes, optimal training patterns are designed for channel estimation to optimize estimation performance jointly with transmitter resources (power and bandwidth). Estimators are derived for both single-carrier and multi-carrier transmissions over frequency-selective MIMO channels, but they specialize to flat MIMO channels, and through the duality established in Chapter 9, they also apply to time-selective and doubly selective MIMO channels. Decision-directed, Kalman-filtering based, and (semi-)blind alternatives are also outlined for joint channel estimation, tracking, and demodulation. Training patterns are further designed in this chapter for carrier frequency offset (CFO) estimation, ensuring full acquisition range. For acquisition ranges not exceeding half-subcarrier spacing, a low-complexity blind CFO estimator is also developed for MIMO OFDM systems.

- *ST Codes with Partial Channel Knowledge: Statistical CSI (Chapter 12)* Different from other chapters, in this chapter and Chapter 13 we consider ST systems where the multi-antenna transmitter has partial channel state information (CSI). Partial CSI here is modeled statistically as a multivariate complex Gaussian vector representing the rendition of the true channel as perceived by the transmitter. This form of statistical CSI can be either made a priori available to the transmitter through sounding experiments, or, it can reach the transmitter through feedback from the receiver. Either way, it is used in this chapter to design ST spread-spectrum systems, transmit-beamformers, and ST coder-beamformer hybrids operating with fixed or adaptive modulations. The designs are optimized to minimize appropriate error bounds and in certain cases to maximize the capacity of multi-antenna systems.

- *ST Codes with Partial Channel Knowledge: Finite-Rate CSI (Chapter 13)* In this chapter we present closed-loop ST designs where partial CSI becomes available to the transmitter only through feedback from the receiver in the form of a finite number of bits. Capitalizing on this finite-rate CSI, multi-antenna transmitters are designed based on beamforming with and without adaptive modulation, on unitary precoded ST multiplexing to further increase transmission rates, and on ST coder-precoder combinations. All designs are cast in a vector quantization framework which entails off-line construction of a codebook to specify a finite number of optimal transmission modes (beamformer or precoder codewords). One codeword for each feedback cycle is selected from this codebook in online operation to "best" adapt the ST transmitter to the quantized MIMO channel codeword conveyed by the feedback bits. Average error probability and in some cases capacity are adopted as criteria of optimality in both the codebook design and in the codeword selection rule.

- *ST Codes in the Presence of Interference (Chapter 14)* In this chapter we introduce spread-spectrum and transmit-beamforming-based schemes to enhance the robustness of ST coded multi-antenna systems in the presence of intentional or unintentional interference. The ST transmitter designs exploit the second-order spatial statistics of the MIMO channel and temporal statistics of the interference. Because ML decoding can be prohibitively complex in the presence of interference, low-complexity quasi-ML demodulators are devised. Since the latter require reliable channel estimators, optimal training sequences are also designed for channel estimation in the presence of interference by exploiting knowledge of the interference covariance matrix at the transmitter.

- *ST Codes for Orthogonal Multiple Access (Chapter 15)* Aiming at orthogonal ST multiple access, in this chapter we show how ST coding can be combined with signature spreading sequences to permeate single-user benefits to multiuser communications stemming from the use of multiple antennas. Spreading sequences are judiciously designed for quasi-synchronous system operation with either single-carrier or multi-carrier transmissions over frequency-selective MIMO channels. The single-carrier system relies on chip-interleaving and block-spreading operations to suppress multiuser interference deterministically. The multi-carrier system is based on MIMO OFDM, where non-overlapping subcarriers are assigned to individual users. With deterministic multiuser separation, ST (de)coding can be performed on a per user basis.