ABSTRACT

In Part I of this tutorial, the major elements that contribute to fading and their effects in a communication channel were characterized. Here, in Part II, these phenomena are briefly summarized, and emphasis is then placed on methods to cope with these degradation effects. Two particular mitigation techniques are examined: the Viterbi equalizer implemented in the Global System for Mobile Communication (GSM), and the Rake receiver used in CDMA systems built to meet Interim Standard 95 (IS-95).

Rayleigh Fading Channels in Mobile Digital Communication Systems Part II: Mitigation

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We repeat Fig. 1, from Part I of the article, where it served as a table of contents for fading channel manifestations. In Part I, two types of fading, large-scale and small-scale, were described. Figure 1 emphasizes the small-scale fading phenomena and its two manifestations, time spreading of the signal or signal dispersion, and time variance of the channel or fading rapidity due to motion between the transmitter and receiver. These are listed in blocks 4, 5, and 6. Examining these manifestations involved two views, time and frequency, as indicated in blocks 7, 10, 13, and 16. Two degradation categories were defined for dispersion, frequency-selective fading and flat- fading, as listed in blocks 8, 9, 11, and 12. Two degradation categories were defined for fading rapidity, fast-fading and slow-fading, as listed in blocks 14, 15, 17, and 18.

In Part I, a model of the fading channel consisting of four functions was described. These functions are shown in Fig. 2 (which appeared in Part I as Fig. 7). A *multipath-intensity profile*, $S(\tau)$, is plotted in Fig. 2a versus time delay τ. Knowledge of $S(\tau)$ helps answer the question "For a transmitted impulse, how does the average received power vary as a function of time delay, τ?" For a single transmitted impulse, the time, T_m, between the first and last received components represents the *maximum excess delay* during which the multipath signal power falls to some threshold level below that of the strongest component. Figure 2b shows the function $|R(\Delta f)|$, designated a *spaced-frequency* correlation function; it is the Fourier transform of $S(\tau)$. $R(\Delta f)$ represents the correlation between the channel's response to two signals as a function of the frequency difference between the two signals. Knowledge of $R(\Delta f)$ helps answer the question "What is the correlation between received signals that are spaced in frequency $\Delta f = f_1 - f_2$?" The *coherence bandwidth*, f_0, is a statistical measure of the range of frequencies over which the channel passes all spectral components with approximately equal gain and linear phase. Thus, the coherence bandwidth represents a frequency range over which frequency components have a strong potential for amplitude correlation. Note that f_0 and T_m are reciprocally related (within a multiplicative constant). As an approximation, it is possible to say that

$$f_0 \approx 1/Tm \qquad (1)$$

A more useful measurement of delay spread is most often characterized in terms of the root mean squared (rms) delay spread, σ_τ [1]. A popular approximation of f_0 corresponding to a bandwidth interval having a correlation of at least 0.5 is

$$f_0 \approx 1/5\sigma_\tau \qquad (2)$$

Figure 2c shows the function $R(\Delta t)$, designated the *spaced-time* correlation function; it is the autocorrelation function of the channel's response to a sinusoid. This function specifies to what extent there is correlation between the channel's response to a sinusoid sent at time t_1 and the response to a similar sinusoid sent at time t_2, where $\Delta t = t_2 - t_1$. The *coherence time*, T_0, is a measure of the expected time duration over which the channel's response is essentially invariant.

Figure 2d shows a *Doppler power spectral density*, $S(v)$, plotted as a function of Doppler-frequency shift, v; it is the Fourier transform of $R(\Delta t)$. The sharpness and steepness of the boundaries of the Doppler spectrum are due to the sharp upper limit on the Doppler shift produced by a vehicular antenna traveling among a dense population of stationary scatterers. The largest magnitude of $S(v)$ occurs when the scatterer is directly ahead of or directly behind the moving antenna platform. The width of the Doppler power spectrum is referred to as the *spectral broadening* or *Doppler spread*, denoted f_d and sometimes called the *fading bandwidth* of the channel. Note that the Doppler spread, f_d, and the coherence time, T_0, are reciprocally related (within a multiplicative constant). In Part I of this tutorial, it was shown that the time (approximately the coherence time) required to traverse a distance $\lambda/2$ when traveling at a constant velocity, V, is

$$T_0 \approx \frac{\lambda/2}{V} = \frac{0.5}{f_d} \qquad (3)$$

DEGRADATION CATEGORIES IN BRIEF

The degradation categories described in Part I are reviewed here in the context of Fig. 3, which summarizes small-scale fading mechanisms, degradation categories, and their effects. (This figure appeared in Part I as Fig. 6.) When viewed in the time-delay domain, a channel is said to exhibit *frequency-selective* fading if $T_m > T_s$ (the delay time is greater than the symbol time). This condition occurs whenever the received multipath components

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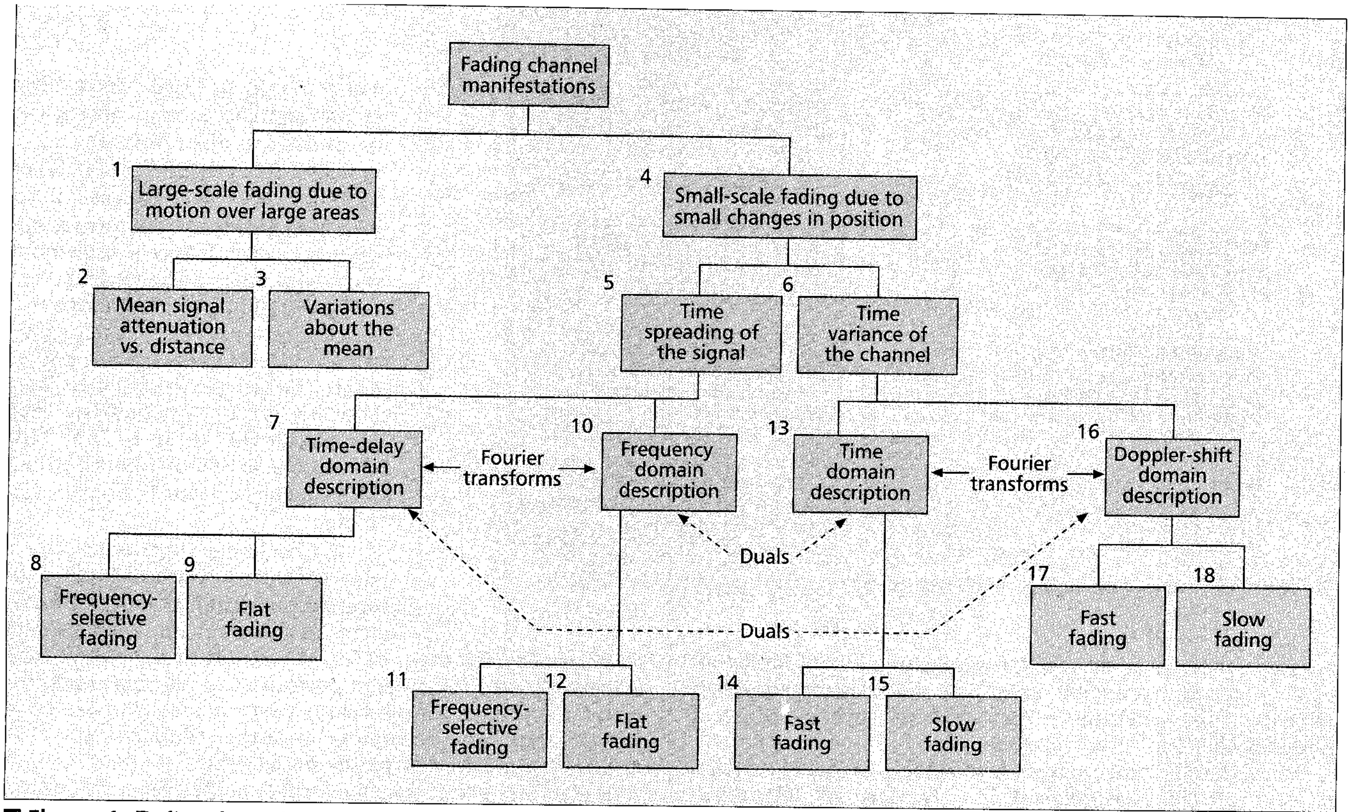


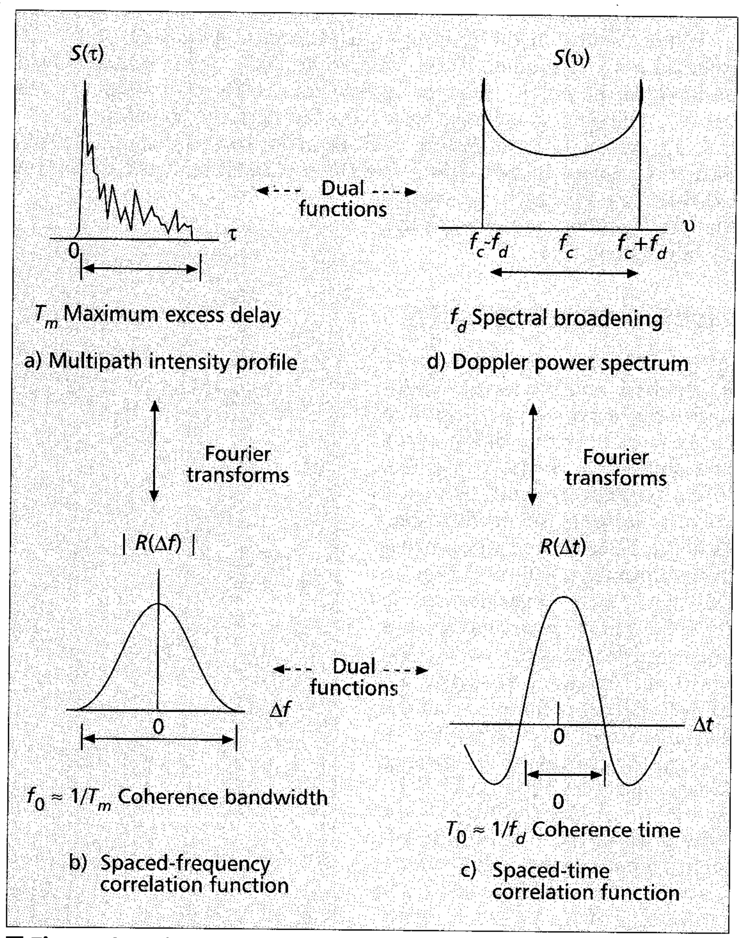
■ **Figure 1.** *Fading channel manifestations.*

of a symbol extend beyond the symbol's time duration, thus causing channel-induced intersymbol interference (ISI).

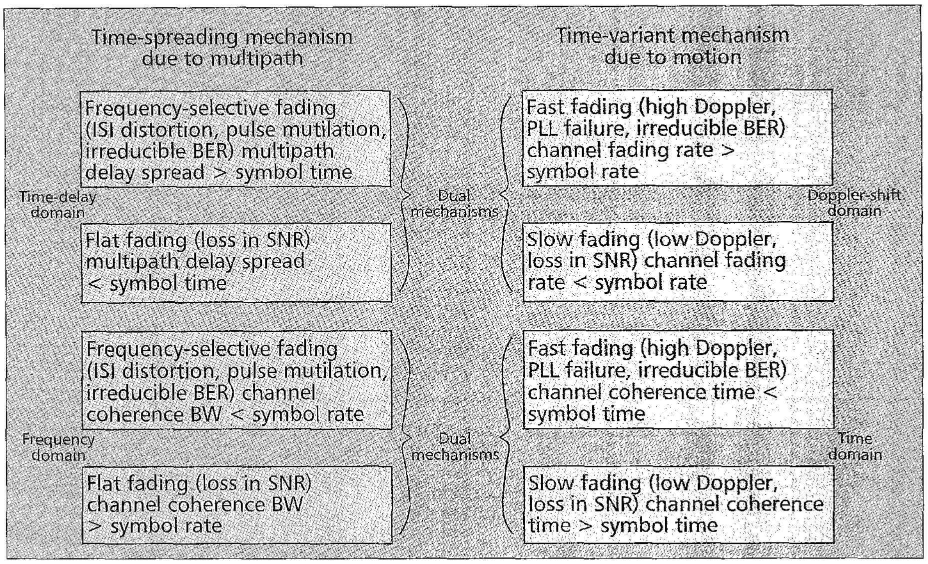
Viewed in the time-delay domain, a channel is said to exhibit *frequency-nonselective* or *flat fading* if $T_m < T_s$. In this case, all of the received multipath components of a symbol arrive within the symbol time duration; hence, the components are not resolvable. Here, there is no channel-induced ISI distortion, since the signal time spreading does not result in significant overlap among neighboring received symbols. There is still performance degradation since the unresolvable phasor components can add up destructively to yield a substantial reduction in signal-to-noise ratio (SNR).

When viewed in the frequency domain, a channel is referred to as frequency-selective if $f_0 < 1/T_s \approx W$, where the symbol rate, $1/T_s$ is nominally taken to be equal to the signal bandwidth W. Flat fading degradation occurs whenever $f_0 > W$. Here, all of the signal's spectral components will be affected by the channel in a similar manner (e.g., fading or no fading). In order to avoid ISI distortion caused by frequency-selective fading, the channel must be made to exhibit flat fading by ensuring that the coherence bandwidth exceeds the signaling rate.

When viewed in the time domain, a channel is referred to as fast fading whenever $T_0 < T_s$, where T_0 is the channel coherence time and T_s is the symbol time. Fast fading describes a condition where the time duration for which the channel behaves in a correlated manner is short compared to the time duration of a symbol. Therefore, it can be expected that the fading character of the channel will change several times during the time a symbol is propagating. This leads to distortion of the baseband pulse shape, because the received signal's components are not all highly correlated throughout time. Hence, fast fading can cause the



■ **Figure 2.** *Relationships among the channel correlation functions and power density functions.*



■ Figure 3. *Small-scale fading: mechanisms, degradation categories, and effects.*

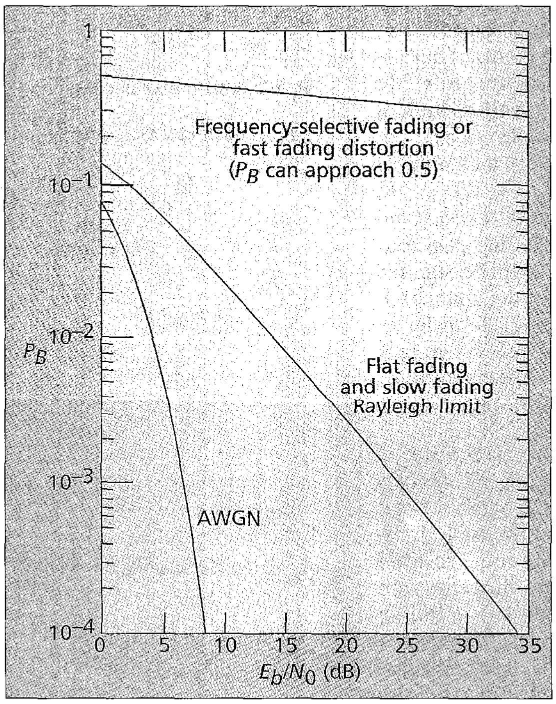
baseband pulse to be distorted, resulting in a loss of SNR that often yields an irreducible error rate. Such distorted pulses typically cause synchronization problems, such as failure of phase-locked-loop (PLL) receivers.

Viewed in the time domain, a channel is generally referred to as introducing *slow fading* if $T_0 > T_s$. Here, the time duration for which the channel behaves in a correlated manner is long compared to the symbol time. Thus, one can expect the channel state to remain virtually unchanged during the time a symbol is transmitted.

When viewed in the Doppler shift domain, a channel is referred to as fast fading if the symbol rate, $1/T_s$, or the signal bandwidth, W, is less than the fading rate, $1/T_0$ or f_d. Conversely, a channel is referred to as slow fading if the signaling rate is greater than the fading rate. In order to avoid signal distortion caused by fast fading, the channel must be made to exhibit slow fading by ensuring that the signaling rate exceeds the channel fading rate.

Mitigation Methods

Figure 4, subtitled "the good, the bad, and the awful," highlights three major performance categories in terms of bit error probability, P_B, versus E_b/N_0. The leftmost exponentially shaped curve represents the performance that can be expected when using any nominal modulation type in additive white Gaussian noise (AWGN). Observe that with a reasonable amount of E_b/N_0, good performance results. The middle curve, referred to as the *Rayleigh limit*, shows the performance degradation resulting from a loss in SNR that is characteristic of flat fading or slow fading when there is no line-of-sight signal component present. The curve is a function of the reciprocal of E_b/N_0 (an inverse-linear function), so for reasonable values of SNR, performance will generally be "bad." In the case of Rayleigh fading, parameters with overbars are often introduced to indicate that a mean is being taken over the "ups" and "downs" of the fading experience. Therefore, one often sees such bit error probability plots with mean parameters denoted by $\overline{P_B}$ and $\overline{E_b}/N_0$. The curve that reaches an irreducible level, sometimes called an *error floor*, represents "awful" performance, where the bit error probability can approach the value of 0.5. This shows the severe distorting effects of frequency-selective fading or fast fading.



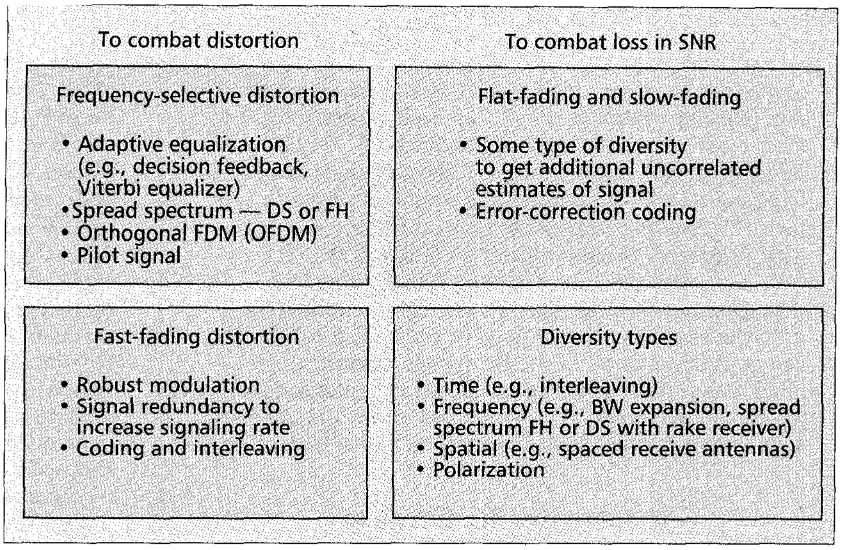
■ Figure 4. *Error performance: the good, the bad, and the awful.*

If the channel introduces signal distortion as a result of fading, the system performance can exhibit an irreducible error rate; when larger than the desired error rate, no amount of E_b/N_0 will help achieve the desired level of performance. In such cases, the general approach for improving performance is to use some form of mitigation to remove or reduce the distortion. The mitigation method depends on whether the distortion is caused by frequency-selective or fast fading. Once the distortion has been mitigated, the P_B versus E_b/N_0 performance should have transitioned from the "awful" bottoming out curve to the merely "bad" Rayleigh limit curve. Next, we can further ameliorate the effects of fading and strive to approach AWGN performance by using some form of diversity to provide the receiver with a collection of uncorrelated samples of the signal, and by using a powerful error correction code.

In Fig. 5, several mitigation techniques for combating the effects of both signal distortion and loss in SNR are listed. Just as Fig. 1 serves as a guide for characterizing fading phenomena and their effects, Fig. 5 can similarly serve to describe mitigation methods that can be used to ameliorate the effects of fading. The mitigation approach to be used should follow two basic steps: first, provide distortion mitigation; next, provide diversity.

Mitigation to Combat Frequency-Selective Distortion

- Equalization can compensate for the channel-induced ISI that is seen in frequency-selective fading. That is, it can help move the operating point from the error-performance curve that is "awful" in Fig. 4 to the one that is "bad." The process of equalizing the ISI involves some method of gathering the dispersed symbol energy back together into its original time interval. In effect, equalization involves insertion of a



■ **Figure 5.** *Basic mitigation types.*

filter to make the combination of channel and filter yield a flat response with linear phase. The phase linearity is achieved by making the equalizer filter the complex conjugate of the time reverse of the dispersed pulse [1]. Because in a mobile system the channel response varies with time, the equalizer filter must also change or adapt to the time-varying channel. Such equalizer filters are therefore called *adaptive equalizers*. An equalizer accomplishes more than distortion mitigation; it also provides diversity. Since distortion mitigation is achieved by gathering the dispersed symbol's energy back into the symbol's original time interval so that it doesn't hamper the detection of other symbols, the equalizer is simultaneously providing each received symbol with energy that would otherwise be lost.

- The decision feedback equalizer (DFE) has a feedforward section that is a linear transversal filter [1] whose length and tap weights are selected to coherently combine virtually all of the current symbol's energy. The DFE also has a feedback section which removes energy that remains from previously detected symbols [1–4]. The basic idea behind the DFE is that once an information symbol has been detected, the ISI it induces on future symbols can be estimated and subtracted before the detection of subsequent symbols.
- The maximum likelihood sequence estimation (MLSE) equalizer tests all possible data sequences (rather than decoding each received symbol by itself) and chooses the data sequence that is the most probable of the candidates. The MLSE equalizer was first proposed by Forney [5] when he implemented the equalizer using the Viterbi decoding algorithm [6]. The MLSE is optimal in the sense that it minimizes the probability of a sequence error. Because the Viterbi decoding algorithm is the way in which the MLSE equalizer is typically implemented, the equalizer is often referred to as the *Viterbi equalizer*. Later in this article we illustrate the adaptive equalization performed in the Global System for Mobile Communications (GSM) using the Viterbi equalizer.
- Spread-spectrum techniques can be used to mitigate frequency-selective ISI distortion because the hallmark of any spread-spectrum system is its capability to reject interference, and ISI is a type of interference. Consider a direct-sequence spread-spectrum (DS/SS) binary phase shift keying (PSK) communication channel comprising one direct path and one reflected path. Assume that the propagation from transmitter to receiver results in a multipath wave that is delayed by τ_k compared to the direct wave. If the receiver is synchronized to the waveform arriving via the direct path, the received signal, $r(t)$, neglecting noise, can be expressed as

$$r(t) = Ax(t)g(t)\cos(2\pi f_c t) + \alpha Ax(t - \tau_k)g(t - \tau_k)\cos(2\pi f_c t + \theta), \tag{4}$$

where x(t) is the data signal, g(t) is the pseudonoise (PN) spreading code, and.tk is the differential time delay between the two paths. The angle q is a random phase, assumed to be uniformly distributed in the range (0, 2p), and a is the attenuation of the multipath signal relative to the direct path signal. The receiver multiplies the incoming r(t) by the code g(t). If the receiver is synchronized to the direct path signal, multiplication by the code signal yields

$$Ax(t)g^2(t)\cos(2\pi f_c t) + \alpha Ax(t - \tau_k)g(t)g(t - \tau_k)\cos(2\pi f_c t + \theta) \tag{5}$$

where $g^2(t) = 1$, and if τ_k is greater than the chip duration,

$$\left|\int g^*(t)g(t - \tau_k)dt\right| \ll \int g^*(t)g(t)dt \tag{6}$$

over some appropriate interval of integration (correlation), where * indicates complex conjugate, and τ_k is equal to or larger than the PN chip duration. Thus, the spread-spectrum system effectively eliminates the multipath interference by virtue of its code-correlation receiver. Even though channel- induced ISI is typically transparent to DS/SS systems, such systems suffer from the loss in energy contained in all the multipath components not seen by the receiver. The need to gather up this lost energy belonging to the received chip was the motivation for developing the Rake receiver [7–9]. The Rake receiver dedicates a separate correlator to each multipath component (finger). It is able to coherently add the energy from each finger by selectively delaying them (the earliest component gets the longest delay) so that they can all be coherently combined.

- In Part I of this article, we described a channel that could be classified as flat fading, but occasionally exhibits frequency-selective distortion when the null of the channel's frequency transfer function occurs at the center of the signal band. The use of DS/SS is a good way to mitigate such distortion because the wideband SS signal would span many lobes of the selectively faded frequency response. Hence, a great deal of pulse energy would then be passed by the scatterer medium, in contrast to the nulling effect on a relatively narrowband signal (see Part I, Fig. 8c) [10].
- Frequency-hopping spread spectrum (FH/SS) can be used to mitigate the distortion due to frequency-selective fading, provided the hopping rate is at least equal to the symbol rate. Compared to DS/SS, mitigation takes place through a different mechanism. FH receivers avoid multipath losses by rapid changes in the transmitter frequency band, thus avoiding the interference by changing the receiver band position before the arrival of the multipath signal.
- Orthogonal frequency-division multiplexing (OFDM) can be used in frequency-selective fading channels to avoid the use of an equalizer by lengthening the symbol duration. The signal band is partitioned into multiple subbands, each exhibiting a lower symbol rate than the original band. The subbands are then transmitted on multiple

orthogonal carriers. The goal is to reduce the symbol rate (signaling rate), $W \approx 1/T_s$, on each carrier to be less than the channel's coherence bandwidth f_0. OFDM was originally referred to as *Kineplex*. The technique has been implemented in the United States in mobile radio systems [11], and has been chosen by the European community under the name coded OFDM (COFDM), for high-definition television (HDTV) broadcasting [12].
- *Pilot signal* is the name given to a signal intended to facilitate the coherent detection of waveforms. Pilot signals can be implemented in the frequency domain as an in-band tone [13], or in the time domain as a pilot sequence which can also provide information about the channel state and thus improve performance in fading [14].

Mitigation to Combat Fast-Fading Distortion

- For fast fading distortion, use a robust modulation (noncoherent or differentially coherent) that does not require phase tracking, and reduce the detector integration time [15].
- Increase the symbol rate, $W \approx 1/T_s$, to be greater than the fading rate, $f_d \approx 1/T_0$, by adding signal redundancy.
- Error-correction coding and interleaving can provide mitigation, because instead of providing more signal energy, a code reduces the required E_b/N_0. For a given E_b/N_0, with coding present, the error floor will be lowered compared to the uncoded case.
- An interesting filtering technique can provide mitigation in the event of fast-fading distortion and frequency-selective distortion occurring simultaneously. The frequency-selective distortion can be mitigated by the use of an OFDM signal set. Fast fading, however, will typically degrade conventional OFDM because the Doppler spreading corrupts the orthogonality of the OFDM subcarriers. A polyphase filtering technique [16] is used to provide time-domain shaping and duration extension to reduce the spectral sidelobes of the signal set, and thus help preserve its orthogonality. The process introduces known ISI and adjacent channel interference (ACI), which are then removed by a post-processing equalizer and canceling filter [17].

Mitigation to Combat Loss in SNR

After implementing some form of mitigation to combat the possible distortion (frequency-selective or fast fading), the next step is to use some form of diversity to move the operating point from the error-performance curve that is "bad" in Fig. 4 to a curve that approaches AWGN performance. The term "diversity" is used to denote the various methods available for providing the receiver with uncorrelated renditions of the signal. Uncorrelated is the important feature here, since it would not help the receiver to have additional copies of the signal if the copies were all equally poor. Listed below are some of the ways in which diversity can be implemented:

- Time diversity — Transmit the signal on L different time slots with time separation of at least T_0. Interleaving, often used with error correction coding, is a form of time diversity.
- Frequency diversity — Transmit the signal on L different carriers with frequency separation of at least f_0. Bandwidth expansion is a form of frequency diversity. The signal bandwidth, W, is expanded to be greater than f_0, thus providing the receiver with several independently fading signal replicas. This achieves frequency diversity of the order $L = W/f_0$. Whenever W is made larger than f_0, there is the potential for frequency-selective distortion unless we further provide some mitigation such as equalization. Thus, an expanded bandwidth can improve system performance (via diversity) only if the frequency-selective distortion the diversity may have introduced is mitigated.
- Spread spectrum is a form of bandwidth expansion that excels at rejecting interfering signals. In the case of direct-sequence spread spectrum (DS/SS), it was shown earlier that multipath components are rejected if they are delayed by more than one chip duration. However, in order to approach AWGN performance, it is necessary to compensate for the loss in energy contained in those rejected components. The Rake receiver (described later) makes it possible to coherently combine the energy from each of the multipath components arriving along different paths. Thus, used with a Rake receiver, DS/SS modulation can be said to achieve path diversity. The Rake receiver is needed in phase-coherent reception, but in differentially coherent bit detection a simple delay line (one bit long) with complex conjugation will do the trick [18].
- FH/SS is sometimes used as a diversity mechanism. The GSM system uses slow FH (217 hops/s) to compensate for those cases where the mobile user is moving very slowly (or not at all) and happens to be in a spectral null.
- Spatial diversity is usually accomplished through the use of multiple receive antennas separated by a distance of at least 10 wavelengths for a base station (much less for a mobile station). Signal processing must be employed to choose the best antenna output or to coherently combine all the outputs. Systems have also been implemented with multiple spaced transmitters; an example is the Global Positioning System (GPS).
- Polarization diversity [19] is yet another way to achieve additional uncorrelated samples of the signal.
- Any diversity scheme may be viewed as a trivial form of repetition coding in space or time. However, there exist techniques for improving the loss in SNR in a fading channel that are more efficient and more powerful than repetition coding. Error correction coding represents a unique mitigation technique, because instead of providing more signal energy it reduces the required E_b/N_0 in order to accomplish the desired error performance. Error correction coding coupled with interleaving [15, 20–25] is probably the most prevalent of the mitigation schemes used to provide improved performance in a fading environment.

Summary of the Key Parameters Characterizing Fading Channels

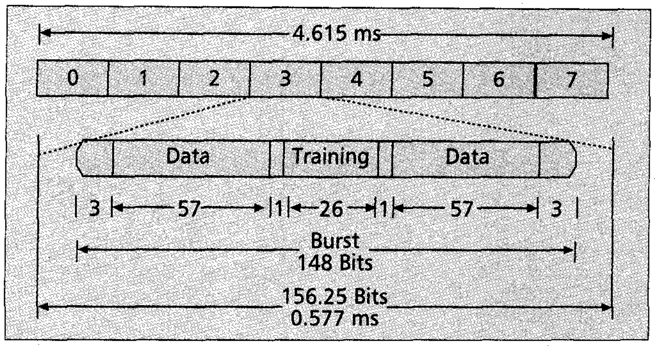
We summarize the conditions that must be met so that the channel does not introduce frequency-selective and fast-fading distortion. Combining the inequalities of Eq. 14 and 23 from Part I of this article, we obtain

$$f_0 > W > f_d \tag{7a}$$

or

$$T_m < T_s < T_0. \tag{7b}$$

In other words, we want the channel coherence bandwidth to exceed our signaling rate, which in turn should exceed the fading rate of the channel. Recall from Part I that without distortion mitigation, f_0 sets an upper limit on the signaling rate, and f_d sets a lower limit on it.



■ **Figure 6.** *The GSM TDMA frame and time slot containing a normal burst.*

FAST-FADING DISTORTION: EXAMPLE 1

If the inequalities of Eq. 7 are not met and distortion mitigation is not provided, distortion will result. Consider the fast-fading case where the signaling rate is less than the channel fading rate, that is,

$$f_0 > W < f_d. \tag{8}$$

Mitigation consists of using one or more of the following methods (Fig. 5):

- Choose the modulation/demodulation technique that is most robust under fast-fading conditions. This means, for example, avoiding carrier recovery with PLLs since fast fading could keep a PLL from achieving lock conditions.
- Incorporate sufficient redundancy that the transmission symbol rate exceeds the channel fading rate. As long as the transmission symbol rate does not exceed the coherence bandwidth, the channel can be classified as flat fading. However, even flat-fading channels will experience frequency-selective distortion whenever a channel null appears at the band center. Since this happens only occasionally, mitigation might be accomplished by adequate error correction coding and interleaving.
- The above two mitigation approaches should result in the demodulator operating at the Rayleigh limit [15] (Fig. 4). However, there may be an irreducible floor in the error-performance versus E_b/N_0 curve due to the frequency modulated (FM) noise that results from the random Doppler spreading (see Part I). The use of an in-band pilot tone and a frequency-control loop can lower this irreducible performance level.
- To avoid this error floor caused by random Doppler spreading, increase the signaling rate above the fading rate still further (100–200 x fading rate) [26]. This is one architectural motive behind time-division multiple access (TDMA) mobile systems.
- Incorporate error correction coding and interleaving to lower the floor and approach AWGN performance.

FREQUENCY-SELECTIVE FADING DISTORTION: EXAMPLE 2

Consider the frequency-selective case where the coherence bandwidth is less than the symbol rate; that is,

$$f_0 < W > f_d. \tag{9}$$

Mitigation consists of using one or more of the following methods (Fig. 5):

- Since the transmission symbol rate exceeds the channel fading rate, there is no fast-fading distortion. Mitigation of frequency-selective effects is necessary. One or more of the following techniques may be considered.
- Adaptive equalization, spread spectrum (DS or FH), OFDM, pilot signal. The European GSM system uses a midamble training sequence in each transmission time slot so that the receiver can learn the impulse response of the channel. It then uses a Viterbi equalizer (explained later) for mitigating the frequency-selective distortion.
- Once the distortion effects have been reduced, introduce some form of diversity and error correction coding and interleaving in order to approach AWGN performance. For direct-sequence spread-spectrum (DS/SS) signaling, a Rake receiver (explained later) may be used for providing diversity by coherently combining multipath components that would otherwise be lost.

FAST-FADING AND FREQUENCY-SELECTIVE FADING DISTORTION: EXAMPLE 3

Consider the case where the coherence bandwidth is less than the signaling rate, which in turn is less than the fading rate. The channel exhibits both fast-fading and frequency-selective fading, which is expressed as

$$f_0 < W < f_d \tag{10a}$$

or

$$f_0 < f_d. \tag{10b}$$

Recalling from Eq. 7 that f_0 sets an upper limit on the signaling rate and f_d sets a lower limit on it, this is a difficult design problem because, unless distortion mitigation is provided, the maximum allowable signaling rate is (in the strict terms of the above discussion) less than the minimum allowable signaling rate. Mitigation in this case is similar to the initial approach outlined in example 1.

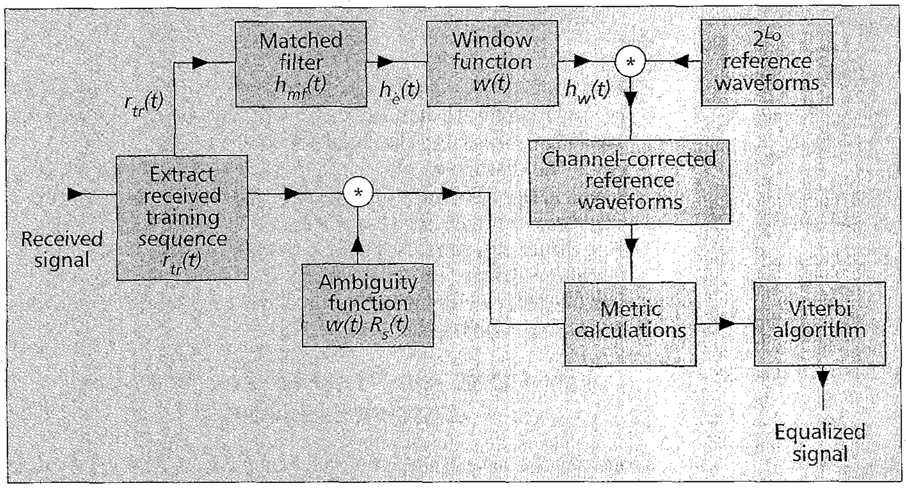
- Choose the modulation/demodulation technique that is most robust under fast-fading conditions.
- Use transmission redundancy in order to increase the transmitted symbol rate.
- Provide some form of frequency-selective mitigation in a manner similar to that outlined in example 2.
- Once the distortion effects have been reduced, introduce some form of diversity and error correction coding and interleaving in order to approach AWGN performance.

THE VITERBI EQUALIZER AS APPLIED TO GSM

Figure 6 shows the GSM time-division multiple access (TDMA) frame, with a duration of 4.615 ms and comprising eight slots, one assigned to each active mobile user. A normal transmission burst, occupying one slot of time, contains 57 message bits on each side of a 26-bit midamble called a *training* or *sounding sequence*. The slot-time duration is 0.577 ms (or the slot rate is 1733 slots/s). The purpose of the midamble is to assist the receiver in estimating the impulse response of the channel in an adaptive way (during the time duration of each 0.577 ms slot). In order for the technique to be effective, the fading behavior of the channel should not change appreciably during the time interval of one slot. In other words, there should not be any fast-fading degradation during a slot time when the receiver is using knowledge from the midamble to compensate for the channel's fading behavior. Consider the example of a GSM receiver used aboard a high-speed train, traveling at a constant velocity of 200 km/hr (55.56 m/s). Assume the carrier frequency to be 900 MHz, (the wavelength is $\lambda = 0.33$ m). From Eq. 3, we can calculate that a half-wavelength is traversed in approximately the time (coherence time)

$$T_0 \approx \frac{\lambda/2}{V} \approx 3\text{ms} \tag{11}$$

Therefore, the channel coherence time is over five times



■ **Figure 7.** *The Viterbi Equalizer as applied to GSM.*

greater than the slot time of 0.577 ms. The time needed for a significant change in fading behavior is relatively long compared to the time duration of one slot. Note that the choices made in the design of the GSM TDMA slot time and midamble were undoubtedly influenced by the need to preclude fast fading with respect to a slot-time duration, as in this example.

The GSM symbol rate (or bit rate, since the modulation is binary) is 271 ksymbols/s and the bandwidth is W = 200 kHz. If we consider that the typical rms delay spread in an urban environment is in the order of σ_τ = 2 µs, then using Eq. 2 the resulting coherence bandwidth is $f_0 \approx$ 100 kHz. It should therefore be apparent that since $f_0 < W$, the GSM receiver must utilize some form of mitigation to combat frequency-selective distortion. To accomplish this goal, the Viterbi equalizer is typically implemented.

Figure 7 illustrates the basic functional blocks used in a GSM receiver for estimating the channel impulse response, which is then used to provide the detector with channel-corrected reference waveforms [27]. In the final step, the Viterbi algorithm is used to compute the MLSE of the message. A received signal can be described in terms of the transmitted signal convolved with the impulse response of the channel. We show this below, using the notation of a received training sequence, $r_{tr}(t)$, and the transmitted training sequence, $s_{tr}(t)$, as follows:

$$r_{tr}(t) = s_{tr}(t) * h_c(t) \qquad (12)$$

where * denotes convolution. At the receiver, $r_{tr}(t)$ is extracted from the normal burst and sent to a filter having impulse response, $h_{mf}(t)$, that is matched to $s_{tr}(t)$. This matched filter yields at its output, an estimate of $h_c(t)$, denoted $h_e(t)$, developed from Eq. 12 as follows:

$$\begin{aligned} h_e(t) &= r_{tr}(t) * h_{mf}(t) \\ &= s_{tr}(t) * h_c(t) * h_{mf}(t) \qquad (13) \\ &= R_s(t) * h_c(t) \end{aligned}$$

where $R_s(t)$ is the autocorrelation function of $s_{tr}(t)$. If $R_s(t)$ is a highly peaked (impulse-like) function, then $h_e(t) \approx h_c(t)$.

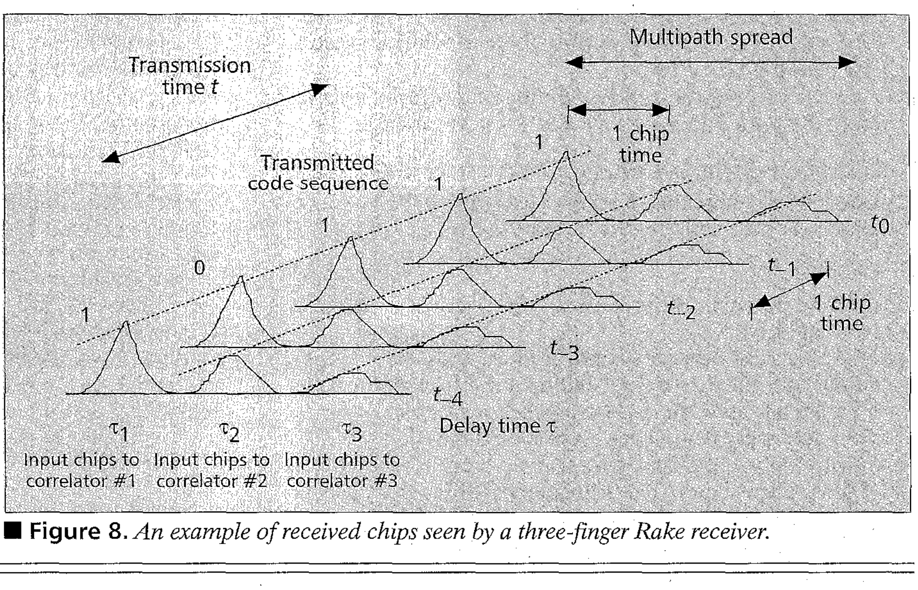
Next, using a windowing function, $w(t)$, we truncate $h_e(t)$ to form a computationally affordable function, $h_w(t)$. The window length must be large enough to compensate for the effect of typical channel-induced ISI. The required observation interval L_o for the window, can be expressed as the sum of two contributions. The interval of length L_{CISI} is due to the controlled ISI caused by Gaussian filtering of the baseband pulses, which are then minimum shift keying (MSK) modulated. The interval of length L_C is due to the channel-induced ISI caused by multipath propagation; therefore, L_o can be written as

$$L_o + L_{CISI} + L_C. \qquad (14)$$

The GSM system is required to provide mitigation for distortion due to signal dispersions of approximately 15–20 µs. The bit duration is 3.69 µs. Thus, the Viterbi equalizer used in GSM has a memory of 4–6 bit intervals. For each L_o-bit interval in the message, the function of the Viterbi equalizer is to find the most likely L_o-bit sequence out of the 2^{L_o} possible sequences that might have been transmitted. Determining the most likely L_o-bit sequence requires that 2^{L_o} meaningful reference waveforms be created by modifying (or disturbing) the 2^{L_o} ideal waveforms in the same way the channel has disturbed the transmitted message. Therefore, the 2^{L_o} reference waveforms are convolved with the windowed estimate of the channel impulse response, $h_w(t)$, in order to derive the disturbed or channel-corrected reference waveforms. Next, the channel-corrected reference waveforms are compared against the received data waveforms to yield metric calculations. However, before the comparison takes place, the received data waveforms are convolved with the known windowed autocorrelation function $w(t)R_s(t)$, transforming them in a manner comparable to that applied to the reference waveforms. This filtered message signal is compared to all possible 2^{L_o} channel-corrected reference signals, and metrics are computed as required by the Viterbi decoding algorithm (VDA). The VDA yields the maximum likelihood estimate of the transmitted sequence [6].

The Rake Receiver Applied to Direct-Sequence Spread-Spectrum Systems

Interim Specification 95 (IS-95) describes a DS/SS cellular system that uses a Rake receiver [7–9] to provide path diversity. In Fig. 8, five instances of chip transmissions correspond-



■ **Figure 8.** *An example of received chips seen by a three-finger Rake receiver.*

ing to the code sequence 1 0 1 1 1 are shown, with the transmission or observation times labeled t_{-4} for the earliest transmission and t_0 for the latest. Each abscissa shows three "fingers" of a signal that arrive at the receiver with delay times τ_1, τ_2, and τ_3. Assume that the intervals between the t_i transmission times and the intervals between the τ_i delay times are each one chip long. From this, one can conclude that the finger arriving at the receiver at time t_{-4}, with delay τ_3, is time-coincident with two other fingers, namely the fingers arriving at times t_{-3} and t_{-2} with delays τ_2 and τ_1, respectively. Since, in this example, the delayed components are separated by exactly one chip time, they are *just* resolvable. At the receiver, there must be a sounding device that is dedicated to estimating the τ_i delay times. Note that for a terrestrial mobile radio system, the fading rate is relatively slow (milliseconds) or the channel coherence time large compared to the chip time ($T_0 > T_{ch}$). Hence, the changes in τ_i occur slowly enough that the receiver can readily adapt to them.

Once the τ_i delays are estimated, a separate correlator is dedicated to processing each finger. In this example, there would be three such dedicated correlators, each processing a delayed version of the same chip sequence, 1 0 1 1 1. In Fig. 8, each correlator receives chips with power profiles represented by the sequence of fingers shown along a diagonal line. Each correlator attempts to match these arriving chips with the same PN code, similarly delayed in time. At the end of a symbol interval (typically there may be hundreds or thousands of chips per symbol), the outputs of the correlators are coherently combined, and a symbol detection is made. At the chip level, the Rake receiver resembles an equalizer, but its real function is to provide diversity.

The interference-suppression nature of DS/SS systems stems from the fact that a code sequence arriving at the receiver merely one chip time late, will be approximately orthogonal to the particular PN code with which the sequence is correlated. Therefore, any code chips that are delayed by one or more chip times will be suppressed by the correlator. The delayed chips only contribute to raising the noise floor (correlation sidelobes). The mitigation provided by the Rake receiver can be termed path diversity, since it allows the energy of a chip that arrives via multiple paths to be combined coherently. Without the Rake receiver, this energy would be transparent and therefore lost to the DS/SS system. In Fig. 8, looking vertically above point τ_3, it is clear that there is interchip interference due to different fingers arriving simultaneously. The spread-spectrum processing gain allows the system to endure such interference at the chip level. No other equalization is deemed necessary in IS-95.

SUMMARY

In Part II of this article, the mathematical model and the major elements that contribute to fading in a communication channel have been briefly reviewed (from Part I of the tutorial). Next, mitigation techniques for ameliorating the effects of each degradation category were treated, and summarized in Fig. 5. Finally, mitigation methods that have been implemented in two system types, GSM and CDMA systems meeting IS-95, were described.

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