1 CODING VS. SPREADING OVER BLOCK FADING CHANNELS

Ezio Biglieri, Giuseppe Caire, and Giorgio Taricco

Dipartimento di Elettronica
Politecnico di Torino
Corso Duca degli Abruzzi 24
I-10129 Torino, Italy*
<name>@polito.it

Abstract: In this chapter we study the optimum tradeoff of coding vs. spreading in a single-cell CDMA mobile communication system with block-fading, block-synchronous (but symbol-asynchronous) transmission, and slow power control. The optimization criterion we choose is based on system capacity, as measured in users/cell×bit/s/Hz,

We adopt an information-theoretic definition of *outage*: this is the event that the mutual information experienced by a user code word falls below the actual user code rate. The system capacity is then defined as above under decoding-delay and outage-probability constraints. We examine the conventional single-user receiver and a linear MMSE multiuser receiver. Our results show that, with ideal power control and optimum coding/spreading tradeoff, capacities close to 1 user/cell×bit/s/Hz are achievable by the conventional receiver, while the capacity gain offered by MMSE multiuser detection is moderate. With non-ideal power control MMSE multiuser detection is more attractive: in fact, it proves to be very robust to residual power-control errors, while conventional detection suffers from a large capacity degradation.

*This research was supported by the Italian Space Agency (ASI).

INTRODUCTION

In this contribution we consider the following problem: Given a system whose bandwidth W and bit rate R_b are assigned, what fraction of the overall bandwidth-expansion factor W/R_b should be allocated to coding, and what to spreading in order to optimize performance? As we shall see, the solution to this problem depends heavily on system assumptions. Hence, we describe our approach to the problem through a case study which, although involving a special system model, outlines the basic solution philosophy. Specifically, we consider here a single-cell CDMA mobile communication system with block-fading, block-synchronous (but symbol-asynchronous) transmission, and slow power control. The optimization criterion is based on system capacity, measured in users/cell×bit/s/Hz and defined under decoding-delay and outage-probability constraints.

Channel model and its constraints

The Gaussian multiple-access channel (MAC), as well as its generalization to a channel in which users may have different signaling waveforms while transmission may be symbol-asynchronous, have been extensively studied [16]. Typical mobile communication channels differ considerably from the Gaussian MAC, since they suffer from time-varying propagation vagaries. These can be either frequency-flat, like shadowing and path attenuation due to changing distance, or frequency-selective, like multipath fading.

In mobile telephony, a strict constraint on the decoding delay is imposed and channel variations are normally rather slow with respect to the maximum allowed delay ΔT (usually a few tens of milliseconds). This situation makes it appropriate to use a channel model introduced in [7, 10] (see also [8, 9]) and known as *block fading* model. It is motivated by the fact that, in many mobile radio situations, the channel coherence time is much longer than one symbol interval, and hence several transmitted symbols are affected by the same fading value. The transmission of a user code word is characterized by a small number of "channel states," intended as the realization of the random variables which determine the channel. For example, the fading time-variation is characterized by its Doppler bandwidth B_d [12, Ch. 14]. The number of "almost independent" realizations of the fading channel during a time span ΔT is, roughly speaking, $\lfloor B_d \Delta T \rfloor + 1$. For a portable handset held by a pedestrian, which is by far the most common case in today's personal communications, $B_d \simeq 0$ so that the channel is constant (but random) during the transmission of a code word [1].

Use of this channel model allows one to introduce a delay constraint for transmission, which is realistic whenever infinite-depth interleaving is not a reasonable assumption. In fact, the block-fading model assumes that a code word of length $n = MN_s$ spans M blocks of N_s symbols each (a group of M blocks will be referred to as a *frame*.) The value of the fading in each block is constant, and each block is sent through an independent channel. An interleaver spreads the code symbols over the M blocks. M is now a measure of the interleaving *delay* of the system: in fact, $M = 1$ (or $N_s = n$) corresponds to no interleaving, while $M = n$ (or $N_s = 1$) corresponds to perfect interleaving. Thus, results obtained for different values of M illustrate the effects of finite decoding delay.

As discussed in [10], with the block fading model if $M < \infty$ the notions of *average mutual information* and *average capacity* have no practical operational meaning. On the contrary, the "instantaneous" mutual information of the channel is a random variable and the Shannon capacity is zero, since the probability that the mutual information is below any specified code rate $\mathcal{R} > 0$ is strictly positive. In this framework, it makes sense to study the capacity for an assigned outage probability, the latter being defined as the probability that the mutual information of the random channel experienced by the transmission of a user code word falls below the actual code rate [10].

In real mobile communication channels synchronization among users has a major impact on the receiver design. For example, consider a transmission scheme where users transmit signal blocks possibly separated in time and assume that a block of user 0 overlaps partially with a block of user 1 and partially with a block of user 2. These are characterized by different average signal levels, carrier phases, transmission delays and linear distortion due to the multipath fading channel. The channel "seen by user 0" exhibits fast (almost instantaneous) variations of the interference parameters which are difficult, if not impossible, to track. Hence, the use of a multiuser detection scheme (e.g., an adaptive interference canceler [6]) may not be possible with block-asynchronous transmission. On the other hand, synchronization at the block level is not difficult to achieve and implies neither frame nor symbol synchronization: it is just a coarse quantization of the time axis, obtained by distributing a common clock to all users in the same cell through the downlink channel (from the base station to the mobile).

Outline of this chapter

In this chapter we consider a simplified model for the link from mobile to base station (the "uplink") in a single-cell CDMA mobile communication system with block fading, block-synchronous (possibly frame and symbol-asynchronous) transmission, and slow power control. We study the capacity of this system, expressed as the maximum of the product of the number of users per cell times the user spectral efficiency, under a constraint on delay and on outage probability. We examine the conventional single-user receiver as well as the linear minimum mean-square error (MMSE) multiuser receiver [14]. With ideal power control, by choosing a target outage probability 10^{-2}, we find that capacities close to 1 user/cell$\times$bit/s/Hz can be achieved by the single-user receiver even with moderate interleaving depth. In this case, the MMSE multiuser receiver offers only a moderate capacity increase with respect to the conventional single-user receiver. On the other hand, with non-ideal power control the latter receiver suffers from a very large capacity degradation, while the MMSE multiuser receiver proves to be considerably robust.

SYSTEM MODEL

Let the uplink channel of a single-cell mobile communication system be characterized by a total (two-sided) system bandwidth W, user information bit rate R_b (bit/s) and maximum allowable decoding delay ΔT. The main assumptions underlying our model are:

- Access is time-slotted, with slots of duration T_s. Each user signal block occupies one time slot and spans the whole system bandwidth.
- A control channel from the receiver to the transmitters implements slow power control. No form of channelized FDMA/TDMA with dynamic centralized allocation is allowed, nor do the transmitters have knowledge of the channel state.
- User codes are selected independently, and decoding is strictly single-user and feedforward. No joint decoding and/or feedback decoding (viz., any form of "onion peeling") is considered.
- The propagation channel is modeled as a frequency-selective Rayleigh block-fading channel. Frequency-flat attenuation due to propagation distance is assumed to be perfectly compensated by power control. However, because of non-ideal power control, a residual log-normal shadowing may be present. This is modeled as the power gain $10^{\sigma_{\rm sh} s/10}$, where s is normally distributed with mean zero and unit variance and $\sigma_{\rm sh}$ (expressed in dB) is the residual *shadowing factor* [4].

Coding and interleaving. In order to introduce time diversity, code words are interleaved and transmitted over M separate signal blocks. We assume that guard times longer than the maximum channel memory are inserted, so that the channel can be considered blockwise memoryless. Encoders produce sequences of length $N_s M$ with elements in the complex signal set $\mathcal{X}$ with unit energy per symbol (as defined before, N denotes the number of symbols per block.) The overall code rate is $\mathcal{R}$ bit/symbol. Following [10], we assume that M (the interleaving depth) is a small integer while the number of symbols per block N_s is large, i.e., we are interested in analyzing the system in the limit for $N_s \to \infty$ and finite (small) M.

Fading channel. Because of the block-fading assumption, during the transmission of a code word the channel is described by the sequence of M impulse responses $\{c_m(\tau) : m = 0, \dots, M-1\}$ or equivalently by the sequence of M frequency responses $\{C_m(f) : m = 0, \dots, M-1\}$, where

$$C_m(f) \triangleq \int_{-\infty}^{\infty} c_m(\tau) e^{-j2\pi f \tau}\, d\tau$$

is the (continuous-time) Fourier transform of $c_m(\tau)$. The support of $c_m(\tau)$ is $[0, T_d]$, where T_d is referred to as the *channel delay spread* [12, Ch. 14]-[2, Ch.

13]. For each positive integer n and for all epochs $\{\tau_i : i = 0, \dots, n-1\}$, the complex random variables $\{c_m(\tau_i) : i = 0, \dots, n-1\}$ are zero-mean jointly Gaussian with i.i.d. real and imaginary parts. Finally, we assume that $c_m(\tau)$ and $c_m(\tau')$ are independent for $\tau \neq \tau'$ (independent scattering assumption). The channel is characterized by the *multipath intensity profile* [12, Ch. 7] $\sigma^2(\tau) = E[|c_m(\tau)|^2]$ with normalized total power $\int_0^{T_d} \sigma^2(\tau) d\tau = 1$. The fading time-frequency second-order statistics is characterized by the time-frequency autocorrelation function

$$\Phi_C(m, \Delta\theta) = E[C_n(\theta) C_{n-m}(\theta - \Delta\theta)^*] \tag{1.1}$$

where we assume implicitly that the fading channel is wide-sense stationary both in time and in frequency.

Equivalent discrete-time channel. Let N denote the number of users transmitting at the same time. We assume that the signal transmitted by user i over block m is linearly modulated and characterized by the unit-energy *signature waveform* $s_m^i(t)$, for $i = 0, \dots, N-1$ and $m = 0, \dots, M-1$. Since $s_m^i(t)$ is allowed to change from block to block, we are considering the possibility of time-varying signature waveforms. The number of symbols (i.e., independent complex Shannon dimensions) for a block of duration $T_{\rm s}$ is given by $N_{\rm s} = WT_{\rm s}/L$, where $1/T$ is the symbol rate and $L = WT$ is the *spreading factor* of the waveform $s_m^i(t)$ (assumed to be the same for all i and m).

The signal received at the base station during the m-th block can be written as

$$y_m(t) = \sum_{i=0}^{N-1} A_m^i c_m^i(t) \star x_m^i(t) + n(t) \tag{1.2}$$

where A_m^i and $c_m^i(t)$ are the amplitude gain and the fading channel from transmitter i to the base station, $\star$ denotes convolution, $n(t)$ is a complex Gaussian white noise with i.i.d. real and imaginary parts and autocorrelation function $E[n(t)n(t-\tau)^*] = N_0\delta(\tau)$, and $x_m^i(t)$ is the signal of user i transmitted during the m-th block, given by

$$x_m^i(t) = \sum_{k=0}^{N_{\rm s}-1} x_m^i[k] s_m^i(t - kT - \tau_m^i - mT_{\rm f}) e^{j\phi_m^i} \tag{1.3}$$

Here, $x_m^i[k] \in \mathcal{X}$, τ_m^i and ϕ_m^i are the delay and carrier phase of user i during block m, and $T_{\rm f}$ is the block time separation.

Let user 0 be the reference user. Receiver 0 is assumed to be a linear time-varying piecewise-constant filter characterized by the sequence of impulse responses $\{w_m(-\tau)^* : m = 0, \dots, M-1\}$, followed by a sampler at the symbol rate $1/T$ and by a decoder matched to the reference-user code (decoder 0). The sequence of output samples in block m is given by

$$y_m[k] = \sum_{i=0}^{N-1} \sum_j A_m^i p_m^i[j] x_m^i[k-j] + \nu_m[k] \tag{1.4}$$

where we define

$$\begin{aligned} h_m^i(t) &\triangleq c_m^i(t) \star s_m^i(t) e^{j\phi_m^i} \\ p_m^i[k] &\triangleq w_m^*(-t) \star h_m^i(t-\tau_m^i)|_{t=kT+\tau_m^0+mT_\mathrm{f}} \\ \nu_m[k] &\triangleq w_m^*(-t) \star n(t)|_{t=kT+\tau_m^0+mT_\mathrm{f}} \end{aligned} \tag{1.5}$$

The noise sequence $\nu_m[k]$ has autocorrelation function

$$E[\nu_m[j]\nu_m^*[j-k]] = N_0 r_{w,m}[k] \tag{1.6}$$

where

$$r_{w,m}[k] \triangleq \int_{-\infty}^{\infty} w_m(\tau) w_m^*(\tau - kT)\, d\tau$$

Decoder 0 makes a decision on the user 0 message based on the observation

$$\{y_m[k] : k = 0, \ldots, N_\mathrm{s} - 1\} \qquad \text{for} \quad m = 0, \ldots, M-1$$

SYSTEM CAPACITY VERSUS OUTAGE PROBABILITY

Outage probability. Let the *channel state* S_m denote the set of random variables A_m^i, τ_m^i, ϕ_m^i, of random impulse responses $c_m^i(\tau)$ and, in the case of random signature waveform selection, of the waveforms $s_m^i(t)$ which determine the discrete-time channel (1.4). For the sake of notational simplicity, let $\mathbf{S} = \{S_m : m = 0, \ldots, M-1\}$ denote the sequence of channel states over the M blocks spanned by a user code word and denote by $I_M(\mathbf{S})$ the instantaneous conditional mutual information (in bit/symbol) as $N_\mathrm{s} \to \infty$:

$$I_M(\mathbf{S}) \triangleq \lim_{N_\mathrm{s}\to\infty} \frac{1}{MN_\mathrm{s}} I\left(\bigcup_{m=0}^{M-1} \{x_m^0[k]\}_{k=0}^{N_\mathrm{s}-1} ; \bigcup_{m=0}^{M-1} \{y_m[k]\}_{k=0}^{N_\mathrm{s}-1} \middle| \mathbf{S} = \mathbf{S} \right) \tag{1.7}$$

where (with a slight abuse of notation [10]) we indicate by $I(\mathbf{X};\mathbf{Y}|\mathbf{S}=\mathbf{S})$ the functional

$$I(\mathbf{X};\mathbf{Y}|\mathbf{S}=\mathbf{S}) \triangleq \sum_{\mathbf{x}\in\mathcal{X}} \sum_{\mathbf{y}\in\mathcal{Y}} p(\mathbf{x},\mathbf{y}|\mathbf{S}) \log_2 \left(\frac{p(\mathbf{x},\mathbf{y}|\mathbf{S})}{p(\mathbf{x}|\mathbf{S})p(\mathbf{y}|\mathbf{S})} \right)$$

where $(\mathbf{X},\mathbf{Y}) \in \mathcal{X} \times \mathcal{Y}$ are random vectors jointly distributed according to $p(\mathbf{x},\mathbf{y}|\mathbf{S})$ with marginals $p(\mathbf{x}|\mathbf{S})$ and $p(\mathbf{y}|\mathbf{S})$ conditionally on $\mathbf{S}$ (note that if $\mathbf{S}$ is a random vector, $I(\mathbf{X};\mathbf{Y}|\mathbf{S}=\mathbf{S})$ is a random variable. The standard conditional average mutual information is obtained by averaging $I_M(\mathbf{S})$ with respect to $\mathbf{S}$.

Outage probability is defined as

$$P_\mathrm{out}(\mathcal{R}) \triangleq P(I_M(\mathbf{S}) < \mathcal{R}) \tag{1.8}$$

An operational motivation of the above definition of outage probability is the following. Assume a code rate $\mathcal{R}$ bit/symbol, and let $\overline{P}_{e|\mathbf{S}}(\mathcal{R})$ denote the message error probability averaged over the code ensemble of all codes with rate $\mathcal{R}$

and length MN_s, randomly generated according to a certain input probability distribution and conditioned with respect to the sequence of channel realizations $\mathbf{S}$. From the channel coding theorem and its strong converse (see [3] and references therein) we can write

$$\lim_{N_s \to \infty} \overline{P}_{e|\mathbf{S}}(\mathcal{R}) = \mathcal{I}_{\{I_M(\mathbf{S}) < \mathcal{R}\}} = \begin{cases} 0 & \text{if} \quad I_M(\mathbf{S}) \geq \mathcal{R} \\ 1 & \text{if} \quad I_M(\mathbf{S}) < \mathcal{R} \end{cases} \tag{1.9}$$

($\mathcal{I}_{\mathcal{A}}$ denotes the indicator function of the event $\mathcal{A}$). By averaging $\overline{P}_{e|\mathbf{S}}(\mathcal{R})$ with respect to $\mathbf{S}$ and exchanging limit with expectation, we can write

$$\lim_{N_s \to \infty} \overline{P}_e(\mathcal{R}) = \lim_{N_s \to \infty} E[\overline{P}_{e|\mathbf{S}}(\mathcal{R})] = E[\mathcal{I}_{\{I_M(\mathbf{S}) < \mathcal{R}\}}] = P(I_M(\mathbf{S}) < \mathcal{R}) = P_{\text{out}}(\mathcal{R}) \tag{1.10}$$

Hence, the information-theoretic outage probability defined by (1.8) is equal to the message error probability averaged over the random code ensemble and over all the possible channel realizations $\mathbf{S}$, in the limit for large N_s.

System capacity. Let R_b denote the user information bit-rate, i.e., the number of bit/s transmitted by a user during a block of duration T_s [2]. The user spectral efficiency (measured in bit/s/Hz) is given by $R_b/W = \mathcal{R}N_s/(WT_s) = \mathcal{R}/L$. Hence, for fixed R_b and W and for a desired outage probability P_{out}, the system capacity is defined by

$$C_{\text{sys}} \triangleq \frac{R_b}{W} \max\{N : P(I_M(\mathbf{S})/L < R_b/W) < P_{\text{out}}\} \qquad \text{users/cell} \times \text{bit/s/Hz} \tag{1.11}$$

In the following we assume that user codes are randomly generated with i.i.d. components according to a complex Gaussian distribution with i.i.d. real and imaginary parts (i.e., with flat power spectral density). An argument supporting this choice is that transmitters have no knowledge of the fading channels and fading amplitudes are identically distributed for all frequencies. However, we do not claim that this choice leads to any kind of optimality in the present setting.

RECEIVER DESIGN AND MUTUAL INFORMATION

The receiver filters $w_m^*(-t)$ can be designed according to several criteria. Here we examine the conventional single-user receiver and the linear MMSE multiuser receiver. In both cases we assume that the receiver has perfect channel-state information, i.e., that it knows A_m^i, τ_m^i, ϕ_m^i, $c_m^i(\tau)$ and $s_m^i(t)$ for all i and m.

Conventional single-user receiver. In this case the receiving filter is the single-user matched filter $w_m^*(-t) = [h_m^0(-t)]^*$.

Linear MMSE multiuser receiver. Linear MMSE multiuser receiver has been independently rediscovered by many after [14]. The goal here is to design $w_m^*(-t)$ such that the mean-square error $\varepsilon^2(w_m) = E[|y_m[k] - x_m^0[k]|^2]$

is minimized. It can be shown that the linear MMSE multiuser receiver is formed by a bank of N filters with the i-th filter matched to the i-th user signal, followed by a (possibly infinite) vector tapped delay line with coefficients $\mathbf{w}_m^\dagger[j]$. By assuming very long blocks ($N_{\rm s} \to \infty$) we obtain the D-transform $\mathbf{w}_m(D) = \sum_j \mathbf{w}_m[j]D^j$ of the sequence of coefficients $\mathbf{w}_m[j]$ as [3]

$$\mathbf{w}_m(D) = \frac{1}{A_m^0}\mathrm{col}_0\left\{\left[\mathbf{R}_m(D^{-1}) + \frac{1}{\gamma_m}\mathbf{E}_m^{-1}\right]^{-1}\right\} \tag{1.12}$$

where we have defined

- The signal-to-noise ratio of user 0 over block m, $\gamma_m \triangleq (A_m^0)^2/N_0$.
- The diagonal matrix of the received energies per symbol, normalized to the energy of signal 0

  $$\mathbf{E}_m \triangleq \mathrm{diag}\left(1, \mathcal{E}_m^1, \ldots, \mathcal{E}_m^{N-1}\right)$$

  where $\mathcal{E}_m^i = (A_m^i/A_m^0)^2$.
- The cross-correlation matrix sequence $\mathbf{R}_m[k]$ whose entry (i,j) is

  $$\{\mathbf{R}_m[k]\}_{ij} \triangleq e^{-j\phi_m^{ij}} \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} c_m^i(\tau)^* c_m^j(\tau') r_{s,m}^{ij}(kT + \tau - \tau' + \tau_m^{ij})d\tau d\tau' \tag{1.13}$$

  where $\phi_m^{ij} \triangleq \phi_m^i - \phi_m^j$ and $\tau_m^{ij} \triangleq \tau_m^i - \tau_m^j$, and where

  $$r_{s,m}^{ij}(\tau) \triangleq \int_{-\infty}^{\infty} s_m^i(t-\tau)^* s_m^j(t)dt \tag{1.14}$$

- The D-transform $\mathbf{R}_m(D) \triangleq \sum_k \mathbf{R}_m[k]D^k$.

By using (discrete-time) inverse Fourier transform, we obtain the resulting minimum MSE as

$$\varepsilon_{\min,m}^2 = \int_{-1/2}^{1/2}\left\{\left[\gamma_m \mathbf{R}_m(\lambda) + \mathbf{E}_m^{-1}\right]^{-1}\right\}_{00} d\lambda \tag{1.15}$$

where, with a slight abuse of notation, we let $\mathbf{R}_m(\lambda) \triangleq \mathbf{R}_m(D)|_{D=e^{j2\pi\lambda}}$.

Mutual information

Because of the channel blockwise memoryless assumption we can write the mutual information $I_M(\mathbf{S})$ as the sum of M contributions, one for each block. By assuming i.i.d. complex Gaussian symbols for all users, the channel (1.4) with input $x_m^0[k]$ and output $y_m[k]$ conditioned on $\mathbf{S}$ is a single-user discrete-time channel with ISI and correlated Gaussian noise. Moreover, we assume that the impulse responses $p^i_m[k]$ are square-summable with probability 1 (this is always met in practice since the channel impulse responses have finite energy). Hence, the mutual information can be obtained as an application of the Toeplitz distribution theorem [5]. Let $Y_m(f)$ denote the power spectral density of $y_m[k]$ conditioned on S_m and let $Z_m(f)$ denote the power spectral density of $y_m[k]$ conditioned on both the input sequence $x^0_m[k]$ and S_m. Hence we can write

$$I_M(\mathbf{S}) = \frac{1}{M}\sum_{m=0}^{M-1}\int_{-1/2}^{1/2}\log_2\frac{Y_m(f)}{Z_m(f)}\,df \tag{1.16}$$

With the single-user receiver, assuming $\{\mathbf{R}_m(\lambda)\}_{00} > 0$ for all $\lambda \in [-1/2, 1/2]$ we obtain

$$I_M(\mathbf{S}) = \frac{1}{M}\sum_{m=0}^{M-1}\int_{-1/2}^{1/2}\log_2\left(1 + \frac{\{\mathbf{R}_m(\lambda)\}_{00}}{\sum_{i=1}^{N-1}\mathcal{E}^i_m\frac{|\{\mathbf{R}_m(\lambda)\}_{0i}|^2}{\{\mathbf{R}_m(\lambda)\}_{00}} + 1/\gamma_m}\right)d\lambda \tag{1.17}$$

With the linear MMSE multiuser receiver, after some algebra we obtain

$$I_M(\mathbf{S}) = \frac{1}{M}\sum_{m=0}^{M-1}\int_{-1/2}^{1/2} -\log_2\left(\left\{\left[\gamma_m\mathbf{R}_m(\lambda) + \mathbf{E}_m^{-1}\right]^{-1}\right\}_{00}\right)d\lambda \tag{1.18}$$

By comparing the MMSE given in (1.15) with the above expression, application of Jensen's inequality to the convex function $-\log_2(x)$ yields the inequality

$$I_M(\mathbf{S}) \geq -\log_2\left[\prod_{m=0}^{M-1}\varepsilon^2_{\min,m}\right]^{1/M} \tag{1.19}$$

where the argument of the logarithm in the RHS is the geometric mean of the MMSE of the M blocks. Equality is obtained when the argument of $\log_2(\cdot)$ in (1.18) does not depend on frequency, i.e., for a memoryless channel. This is the case, for example, of symbol-synchronous transmission and frequency-flat fading.

Strictly Nyquist band-limited waveforms. In the following we limit ourselves to strictly Nyquist band-limited waveforms of the type

$$s^i_m(t) = \frac{\mathrm{sinc}(t/T)}{\sqrt{TL}}\sum_{\ell=0}^{L-1}e^{j(2\pi\ell t/T + \psi^i_m[\ell])} \tag{1.20}$$

where $\mathrm{sinc}(t) \triangleq \sin(\pi t)/(\pi t)$ and where $\{\psi^i_m[\ell] : \ell = 0, \dots, L-1\}$ is the spreading sequence of user i over block m. Using waveforms (1.20) is tantamount to orthogonal frequency division multiplexing (OFDM) [1] where the same symbol $x^i_m[k]$ is transmitted over L adjacent subbands spaced by $1/T$ and where a different phase $\psi^i_m[\ell]$ is assigned to each ℓ-th subcarrier. This choice of the user waveforms allows us to write $\mathbf{R}_m(\lambda)$ in a very simple way in terms of the fading channel frequency responses $C_m^i(f) = \int_0^{T_d} c_m^i(\tau)e^{-j2\pi f\tau}d\tau$. Although choice (1.20) is admittedly restrictive, we believe it is not overly so, inasmuch as it allows us to reach conclusions that are expected to hold for other waveforms as well.

In this case, after long but straightforward calculations we can write

$$\mathbf{R}_m(\lambda) = \frac{1}{L}\mathbf{C}_m(\lambda)^\dagger \mathbf{C}_m(\lambda)$$

where $\mathbf{C}(\lambda)$ is the $L \times N$ matrix whose entry (ℓ, i) is

$$\{\mathbf{C}_m(\lambda)\}_{\ell i} \triangleq e^{j(\phi_m^i - 2\pi\lambda\tau_m^i/T)} C_m^i\left(\frac{\lambda+\ell}{T}\right) e^{j(\psi_m^i[\ell] - 2\pi\ell\tau_m^i/T)} \tag{1.21}$$

Note that $\text{rank}(\mathbf{R}_m(\lambda)) \leq \min\{L, N\}$ and, for $L < N$, $\mathbf{R}_m(\lambda)$ is not invertible.

Tradeoff between coding and spreading. If $L = 1$, then $\mathbf{R}_m(\lambda)$ has rank 1. Thus, we write $\mathbf{R}_m(\lambda) = \mathbf{a}(\lambda)\mathbf{a}^\dagger(\lambda)$ and use the matrix identity

$$\left[\mathbf{I} + \mathbf{a}\mathbf{a}^\dagger\right]^{-1} = \mathbf{I} - \frac{1}{1 + |\mathbf{a}|^2}\mathbf{a}\mathbf{a}^\dagger$$

(where $\mathbf{a}$ is a column N-vector) to see that for $L = 1$ the two mutual informations (1.17) and (1.18) coincide. Hence, for signaling waveforms of the type (1.20) with spreading $L = 1$ the conventional single-user receiver and the linear MMSE multiuser receiver give exactly the same result in terms of outage probability and system capacity. Moreover, in this case the mutual information does not depend on the user relative delays (so that symbol-synchronous and symbol-asynchronous transmission are equivalent in terms of outage and system capacity). In order to exploit the ability of the MMSE receiver to cancel interference we need some spreading ($L > 1$). Hence, for a given spectral efficiency $R_b/W = \mathcal{R}/L$ coding and spreading must be traded off: if we want to increase the waveform spreading factor L we must increase the code rate $\mathcal{R}$ accordingly. The best solution is the one that maximizes system capacity for given channel statistics.

Asymptotic analysis. We can gain more intuition on the coding-spreading trade-off from the asymptotic characterization of the system capacity of a very idealized CDMA single-cell system presented by D. Tse and S. Hanly in [15]. Consider the uplink of a symbol-synchronous Direct-Sequence (DS) CDMA system with a single cell, N users and spreading sequences of length L chips. No fading and no shadowing or power control errors are taken into account, so that the channel is purely AWGN. The spreading sequences are real or complex and randomly generated i.i.d. sequences with mean zero and variance $1/L$ (classical examples are BPSK and QPSK spreading sequences). Once the sequences are generated, they are assigned to the users permanently. Although for every finite

L the system capacity as defined by (1.11) is a random variable depending on the random spreading sequence assignement, a very strong characterization of system capacity is possible for very large systems, i.e., those in which both L and N grow to infinity while keeping constant the ratio $\alpha = N/L$.

Specifically, assume independent random coding [3] for each user, where users' symbols $x_m^i[k]$ are generated i.i.d. according to a complex circularly symmetric Gaussian distribution with mean zero and per-component variance 1/2. Moreover, assume that all users are received with the same amplitude $A_m^i = A$ (perfect power control and no fading). From the results of [15] we obtain the asymptotic system capacity as $L \to \infty$ and $N/L \to \alpha$ with a conventional single-user receiver as

$$C_{\rm sys} = \log_2(1+\beta) \max\left\{ \left(\frac{1}{\beta} - \frac{N_0}{A^2} \right), 0 \right\} \tag{1.22}$$

where β is the desired signal-to-interference plus noise ratio (SINR) at the receiver output. Under the same assumptions, the asymptotic system capacity with a linear MMSE multiuser receiver is given by

$$C_{\rm sys} = \log_2(1+\beta)(1+\beta) \max\left\{ \left(\frac{1}{\beta} - \frac{N_0}{A^2} \right), 0 \right\} \tag{1.23}$$

Two comments are in order to illustrate the above formulas:

- Since the channel is AWGN and the received power from each user is kept constant, the mutual information as derived previously is constant if conditioned on the choice of the random spreading sequences. For $L \to \infty$, it is possible to prove that the mutual information converges in probability to the constant $\log_2(1+\beta)$ (where β depends on α and on the SNR A^2/N_0). Then, by letting the code rate $\mathcal{R}$ equal to $\log_2(1+\beta)$, the outage probability is zero.

- It is convenient to parameterize $C_{\rm sys}$ in terms of β, which plays the role of a design parameter. We can write $C_{\rm sys} = NR_b/W = N\mathcal{R}/L = \alpha\mathcal{R}$. Since in order to have zero outage probability, we let $\mathcal{R} = \log_2(1+\beta)$, then the number α of users$\times$cell per dimension is given by

  $$\alpha = \max\left\{ \left(\frac{1}{\beta} - \frac{N_0}{A^2} \right), 0 \right\} \tag{1.24}$$

  for the conventional single-user receiver, and by

  $$\alpha = (1+\beta) \max\left\{ \left(\frac{1}{\beta} - \frac{N_0}{A^2} \right), 0 \right\} \tag{1.25}$$

  for the linear MMSE receiver.

Fig. 1.1 shows $C_{\rm sys}$ vs. β, for SNR= 10 dB. We notice that the maximum system capacity with the conventional receiver is obtained as $\beta \to 0$, i.e., for

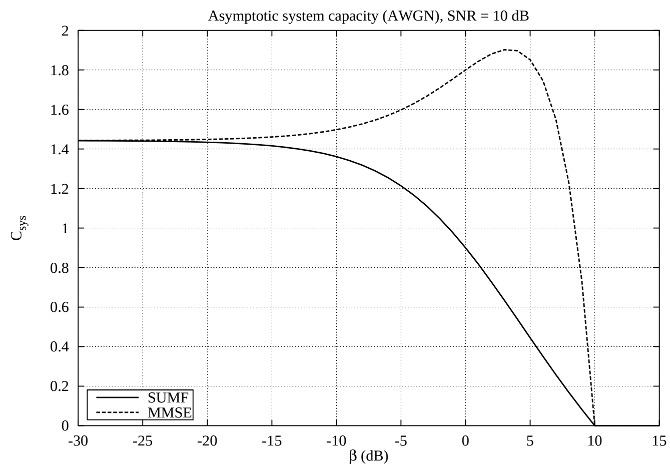


Figure 1.1 System capacity vs. β, the signal-plus-interference to noise ratio at the receiver output for SNR= 10 dB, a single-user matched-filter receiver (continuous line) and a linear MMSE receiver (dotted line). The channel is ideal AWGN.

a very large number of users×cell per dimension ($\alpha \to \infty$) and very low-rate coding ($\mathcal{R} \to 0$). This is in agreement with Viterbi's findings and with the IS-95 return link philosophy [17, 13]. The system capacity attained by the linear MMSE receiver is larger than that of the conventional receiver for all values of β. We notice that in this case there exists an optimal β maximizing the system capacity. This represents (at least asymptotically and in the case of AWGN and perfect power control) the best trade-off between coding and spreading. Let $\beta_{\rm opt}$ be the capacity maximizing β, and assume a system with $N \gg 1$ users. Then, the optimal coding rate is approximately given by $\mathcal{R}_{\rm opt} = \log_2(1+\beta_{\rm opt})$ and the optimal spreading length is approximately given by $L_{\rm opt} = N/\alpha$, where α is given by (1.25) and evaluated for $\beta = \beta_{\rm opt}$.

NUMERICAL RESULTS

To substantiate the above results with numerical examples, here we borrow some system parameters from the cellular CDMA standard IS-95 [13]. We consider system bandwidth $W = 1.25$ MHz, user bit-rate $R_b = 9.6$ kb/s and interleaving depths $M = 1, 2$, and 4. The resulting user spectral efficiency is $\mathcal{R}/L = R_b/W \simeq 7.7 \cdot 10^{-3}$. We make the simplifying assumption that all fading channels $c_m^i(\tau)$ are i.i.d. for all i and m. We consider the Rayleigh fading channel model given in [18] for a typical urban environment, with a multipath intensity profile

$$\sigma^2(\tau) = \begin{cases} \dfrac{e^{-\tau/t_0}}{t_0(1-e^{-T_d/t_0})} & 0 \le \tau \le T_d \\ 0 & \text{elsewhere} \end{cases}$$

with $t_0 = 1$ μs and $T_d = 7$ μs.

Since outage probability does not seem to be amenable to a closed-form expression, we resorted to Monte Carlo simulation. In order to compute mutual information, we discretize the integration domain $[-1/2, 1/2]$ into frequency intervals, in each of which the matrix $\mathbf{R}_m(\lambda)$ does not vary appreciably with λ. The number of discretization intervals D is chosen as $D = \lfloor W/(LB_c) \rfloor + 1$ where B_c is the channel coherence bandwidth [12, Ch. 14]-[2, Ch. 13]. We found that accurate results for the fading model considered here can be obtained by using $B_c = 66.6$ kHz.

Spreading sequences $\{\psi_m^i[\ell] : \ell = 0, \ldots, L-1\}$ are assumed to be i.i.d. randomly generated for all users according to a uniform distribution over $[-\pi, \pi]$. Independently generated sequences are used in different blocks.

The received energy per symbol $(A_m^i)^2$ is modeled as a log-normal random variable with residual shadowing factor σ_{sh}. We considered the values $\sigma_{\text{sh}} = 0, 2$ and 8 dB. For $\sigma_{\text{sh}} = 0$ we have ideal power control. In our simulations we consider a signal-to-noise ratio in the absence of residual shadowing (or, equivalently, with ideal power control) equal to $1/N_0 = 10$ dB (recall that the symbols in $\mathcal{X}$ have unit average energy). We assume shadowing to be a process so slow that it can be considered as constant over all the M blocks spanned by a user code word. Hence, A_m^0 does not depend on m. However, we assume that a slot hopping scheme is applied so that two users can interfere in at most one block out of M. Then, the A_m^i, for $i > 0$, are i.i.d. for different i's and m's, since user 0 "sees" different interfering users in each block. A method based on orthogonal latin squares for designing hopping schemes with the above property is advocated in [11]. Use of such hopping schemes is highly desirable since in this way the *interferer diversity* of the system is equal to the interleaving depth M, and the probability of worst case situations where a user experiences persistently strong interference over all the blocks spanned by a code word is reduced.

Finally, we note that in order to reduce the amount of computations in the Monte Carlo simulation of the MMSE multiuser outage probability we can compute the argument of the logarithm in (1.18) as

$$\left\{ \left[\gamma_m \mathbf{R}_m(\lambda) + \mathbf{E}_m^{-1} \right]^{-1} \right\}_{00} = \left(\{\mathbf{U}(\lambda)\}_{00} \right)^{-2}$$

where $\mathbf{U}(\lambda)\mathbf{U}^\dagger(\lambda)$ is the Cholesky factorization of $[\gamma_m \mathbf{R}(\lambda) + \mathbf{E}_m^{-1}]$ and $\mathbf{U}(\lambda)$ is upper triangular.

Outage probability results

The outage probability $P_{\text{out}}(\mathcal{R}) = P(I_M(\mathbf{S}) < \mathcal{R})$ was computed for $M = 1, 2, 4$, $L = 1, 2, 4, 8, 16$, $N = 10, 20, 30, 40, 50, 60, 70, 80, 90, 100$ and $\sigma_{\text{sh}} = 0, 2$ and 8 dB, for both the single-user receiver and the linear MMSE multiuser receiver. Because of space limitations we include here only a subset of our results.

Fig. 1.2 shows $P_{\text{out}}(\mathcal{R})$ vs. the user spectral efficiency $\mathcal{R}/L$ for $M = 1, L = 1$ (above) and for $M = 1, L = 16$ (below), in the case of ideal power control ($\sigma_{\text{sh}} = 0$ dB). As expected, the performance of the two receivers for $L = 1$ is the same. On the contrary, for $L > 1$ the MMSE receiver performs uniformly better than the single-user receiver.

Fig. 1.3 shows $P_{\text{out}}(\mathcal{R})$ vs. $\mathcal{R}/L$ for $\sigma_{\text{sh}} = 2$ dB (above) and $\sigma_{\text{sh}} = 8$ dB (below), for $M = 1, L = 16$. Fig. 1.4 shows analogous results for $M = 4, L = 16$. We note that with non-ideal power control the single-user receiver suffers from a large performance degradation in terms of outage probability. For example, in the case $\sigma_{\text{sh}} = 8$ dB, the single-user yields $P_{\text{out}}(\mathcal{R}) > 10^{-1}$ already with $N = 10$ users. On the contrary, the linear MMSE multiuser receiver is much more robust to power control errors and achieves low outage probabilities even for $\sigma_{\text{sh}} = 8$ dB. This fact may be viewed as a redefinition in terms of outage probability of the *near-far resistance* of MMSE multiuser detectors (see [6] and references therein): the MMSE multiuser receiver is able to cope with unbalanced signal power situations.

As expected, interleaving depth $M > 1$ provides a benefit. This effect is particularly worthy of notice when hopping schemes with maximum interferer diversity are employed, as in these simulations. In this way, a diversity order equal to the interleaving depth M is achieved against the residual log-normal interference.

System capacity results

Fig. 1.5 shows P_{out} as a function of N, for $\mathcal{R}/L = 7.7 \cdot 10^{-3}$, $M = 1, 2, 4$, $L = 1, 2, 4, 8, 16$ and $\sigma_{\text{sh}} = 0$ dB, for the single-user (above) and for the linear MMSE multiuser receivers (below), respectively. A usual value for the desired outage probability is $P_{\text{out}} = 10^{-2}$. In this case we see that with $M = 4$ and $L = 16$ the single-user and the MMSE multiuser receiver can accommodate $N \simeq 105$ and $N \simeq 125$ user/cell, respectively. This corresponds to $C_{\text{sys}} \simeq 0.8$ (single-user) and $C_{\text{sys}} \simeq 0.96$ (multiuser) user/cell$\times$bit/s/Hz.

Fig. 1.6 shows P_{out} as a function of N, for $\mathcal{R}/L = 7.7 \cdot 10^{-3}$, $M = 1, 2, 4$, $L = 2, 16$ and $\sigma_{\text{sh}} = 2$ dB, for the single-user (above) and for the linear MMSE multiuser receivers (below), respectively. We observe that an increase of waveform spreading from 2 to 16 yields a large improvement only with the MMSE multiuser receiver. For a desired outage probability $P_{\text{out}} = 10^{-2}$, with $M = 4$ and $L = 16$ the single-user and the MMSE multiuser receiver can accommodate $N \simeq 66$ and $N \simeq 95$ user/cell, respectively, corresponding to $C_{\text{sys}} \simeq 0.5$ (single-user) and $C_{\text{sys}} \simeq 0.73$ (multiuser) user/cell$\times$bit/s/Hz.

Finally, Fig. 1.7 shows P_{out} as a function of N, for $\mathcal{R}/L = 7.7 \cdot 10^{-3}$, $M = 1, 2, 4$, $L = 2, 16$ and $\sigma_{\text{sh}} = 8$ dB, for the single-user (above) and for the linear MMSE multiuser receivers (below), respectively. Here we see that the single-user receiver cannot achieve outage probabilities smaller than 10^{-1}. On the contrary, with the MMSE multiuser receiver it is possible to accommodate up to

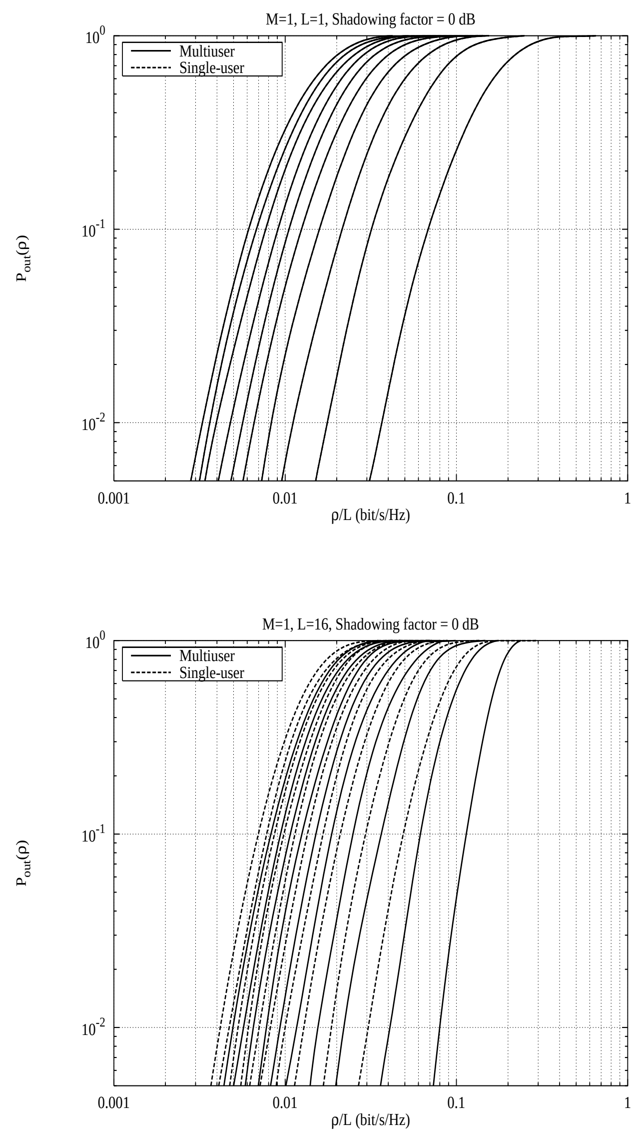
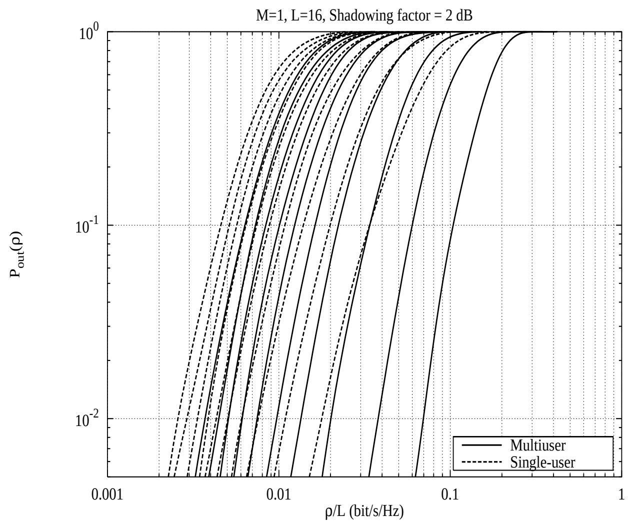


Figure 1.2 $P_{\rm out}(\rho)$ vs. ρ/L for $M = 1$, $L = 1$ (above) and $L = 16$ (below). Curves for $N = 10, 20, \ldots, 100$ users are shown. For each family of curves, the rightmost corresponds to $N = 10$ and the leftmost to $N = 100$.



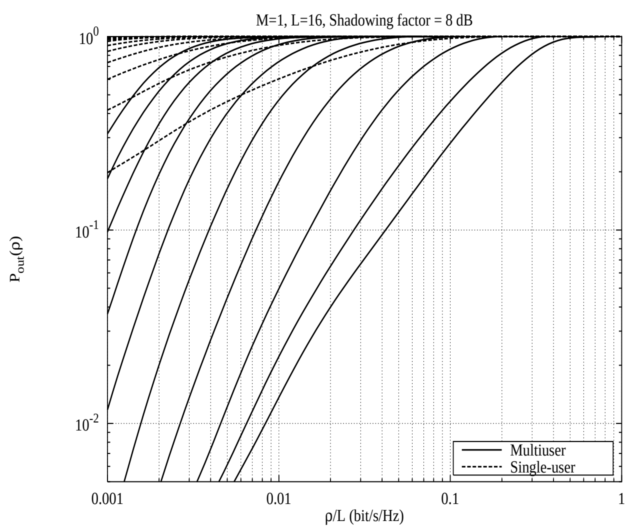


Figure 1.3 $P_{\rm out}(\rho)$ vs. ρ/L for $M = 1$, $L = 16$, with residual shadowing factor $\sigma_{\rm sh} = 2$ dB (above) and $\sigma_{\rm sh} = 8$ dB (below). Curves for $N = 10, 20, \ldots, 100$ users are shown. For each family of curves, the rightmost corresponds to $N = 10$ and the leftmost to $N = 100$.

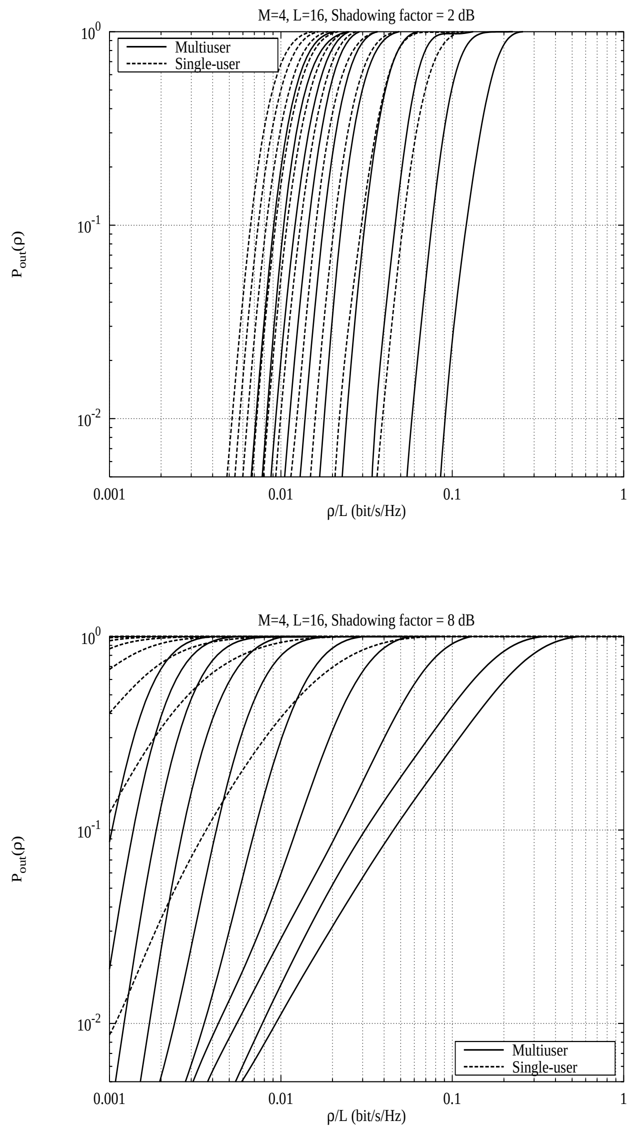


Figure 1.4 $P_{\rm out}(\rho)$ vs. ρ/L for $M = 4$, $L = 16$, with residual shadowing factor $\sigma_{\rm sh} = 2$ dB (above) and $\sigma_{\rm sh} = 8$ dB (below). Curves for $N = 10, 20, \ldots, 100$ users are shown. For each family of curves, the rightmost corresponds to $N = 10$ and the leftmost to $N = 100$.

20 users (with $M = 4$ and $L = 16$), even in this very adverse residual shadowing conditions. The resulting system capacity is $C_{\rm sys} \simeq 0.15$ user/cell$\times$bit/s/Hz.

CONCLUSIONS

In this paper we examined the tradeoff between coding and spreading in a single-cell CDMA mobile communication system with block-fading, block-synchronous transmission and slow power control. The cost function selected for optimization was *system capacity*, expressed as the maximum of the product of the number of users per cell times the user spectral efficiency, under an outage probability constraint. Despite some necessary simplifications, the model chosen here takes into account many features of real-world systems, such as time-varying random spreading waveforms, non-ideal power control, symbol-asynchronous transmission, and multipath fading.

We derived expressions for the mutual information characterizing the M-block random channel spanned by the transmission of a user code word in the cases of conventional single-user receiver of the linear MMSE multiuser receiver. From these expressions it was possible to compute outage probability and system capacity by Monte Carlo simulation. The tradeoff between coding and spreading and interleaving depth was examined for different receiver structures. Notice finally that our analysis is independent of the particular coding and modulation scheme adopted.

Our results show that with ideal power control and moderate interleaving depth ($M = 4$) and spreading ($L = 16$), capacities close to 1 user/cell$\times$bit/s/Hz can be obtained by both the single-user and the MMSE multiuser receivers, for outage probability $P_{\rm out} = 10^{-2}$. On the contrary, the MMSE multiuser receiver proves to be very robust to power control inaccuracies while the single-user receiver breaks down. Hence, a precise power control algorithm is a key issue in conventional CDMA systems, while MMSE multiuser detection allows for (moderately) unbalanced signal powers. This fact may justify the implementation of linear MMSE multiuser detection. We note in passing that, with block-synchronous transmission, adaptive MMSE multiuser detection may be implemented also with random time-varying spreading waveforms.

Two major aspects that have not been taken into account here are voice activity and inter-cell interference [4]. On one hand, voice activity detection in the transmitter increase the system capacity, since the number of users simultaneously transmitting is just a fraction of the total number of users per cell N. On the other hand, inter-cell interference decreases system capacity as it increase the interference total power. Moreover, also cell sectorization should be taken into account in order to asses the system capacity of an actual system.

Notes

1. Normal Doppler bandwidths with moving vehicles and carrier frequencies around 1 GHz are about 100 Hz, so that with $\Delta T = 100$ ms we get not more than 10 fading "blocks".

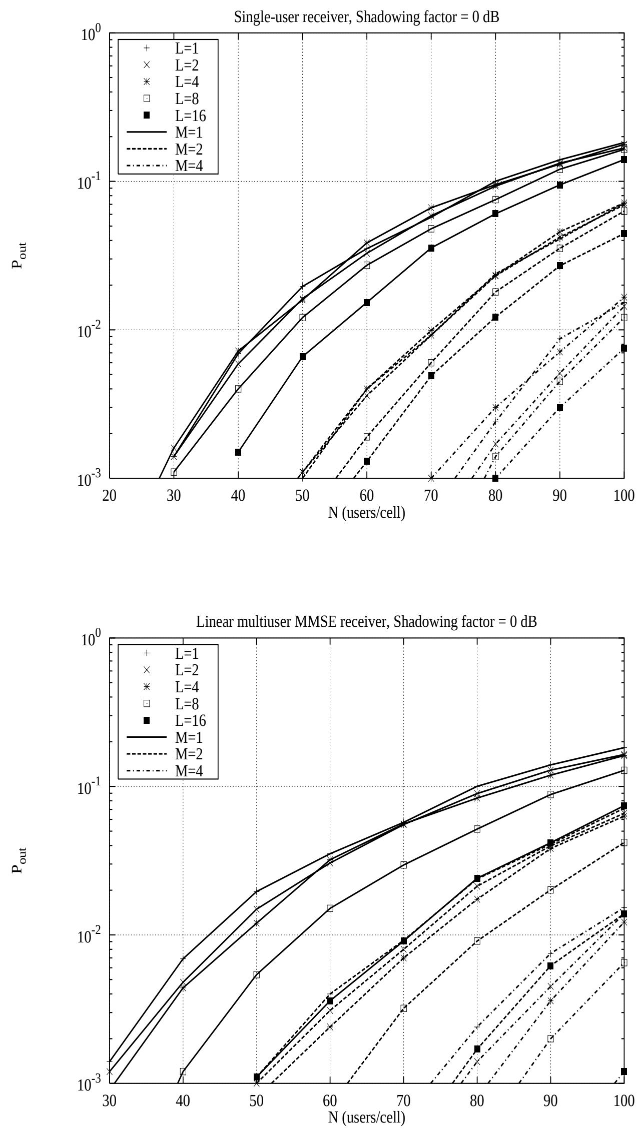


Figure 1.5 $P_{\rm out}$ vs. N for fixed $\rho/L = 7.7 \cdot 10^{-3}$, for the single-user receiver (above) and for the linear MMSE multiuser receiver (below) with interleaving depths $M = 1, 2, 4$, spreading factors $L = 1, 2, 4, 8, 16$ and ideal power control ($\sigma_{\rm sh} = 0$ dB).

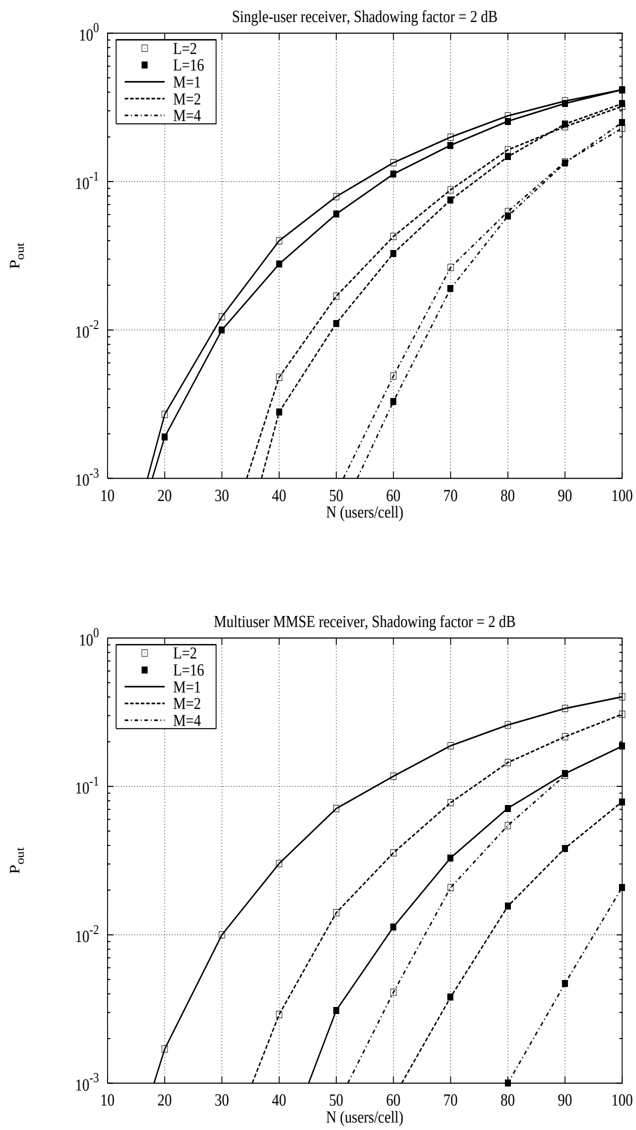
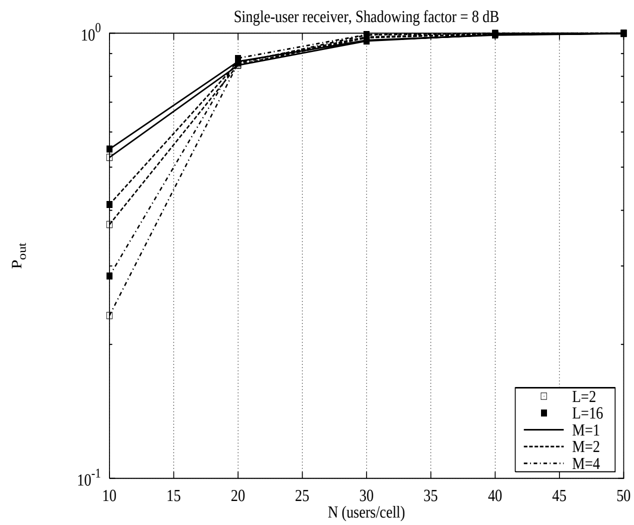


Figure 1.6 P_{out} vs. N for fixed $\rho/L = 7.7 \cdot 10^{-3}$, for the single-user receiver (above) and for the linear MMSE multiuser receiver (below) with interleaving depths $M = 1, 2, 4$, spreading factors $L = 2, 16$ and residual shadowing factor $\sigma_{\mathrm{sh}} = 2$ dB.



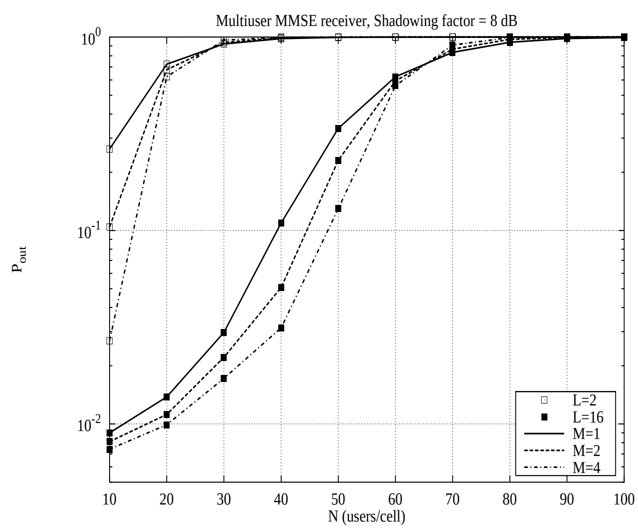


Figure 1.7 $P_{\rm out}$ vs. N for fixed $\rho/L = 7.7 \cdot 10^{-3}$, for the single-user receiver (above) and for the linear MMSE multiuser receiver (below) with interleaving depths $M = 1, 2, 4$, spreading factors $L = 2, 16$ and residual shadowing factor $\sigma_{\rm sh} = 8$ dB.

2. Note that in a slotted transmission the average user bit-rate depends also on the number of transmitted blocks per second. Here, we assume that users transmit continuously, so that the burst bit-rate coincides with the information bit-rate R_b. In the case of a variable rate system, blocks can be transmitted with a duty cycle $< 100\%$ in order to accommodate lower rates.

3. We number row and columns of N-vectors and $N \times N$ matrices from 0 to $N-1$ and $\mathrm{row}_i\{\mathbf{M}\}$, $\mathrm{col}_i\{\mathbf{M}\}$ and $\{\mathbf{M}\}_{ij}$ denote the i-th row, the i-th column and the (i,j) element of $\mathbf{M}$, respectively.

References

[1] M. Alard and R. Lassalle, "Principles of modulation and channel coding for digital broadcasting for mobile receivers," *European Broadcasting Union Review,* No. 224, Aug. 1987.

[2] S. Benedetto and E. Biglieri, *Principles of Digital Transmission with Wireless Applications.* New York: Plenum, 1998.

[3] T. Cover and J. Thomas, *Elements of Information Theory.* New York: J. Wiley & Sons, 1991.

[4] K. Gilhousen, I. Jacobs, R. Padovani, A. Viterbi, L. Weaver and C. Wheatly III, "On the capacity of a cellular CDMA system," *IEEE Trans. on Vehic. Tech.,* Vol. 40, No. 2, pp. 303 – 312, May 1991.

[5] U. Grenander and G. Szegö, *Toeplitz Forms and Their Applications.* New York: Chelsea, 1983.

[6] M. Honig, U. Madhow and S. Verdu, "Blind adaptive multiuser detection," *IEEE Trans. on Inform. Theory,* Vol. 41, No. 4, pp. 944 – 960, July 1995.

[7] G. Kaplan and S. Shamai (Shitz), "Error probabilities for the block-fading Gaussian channel," *A.E.Ü.,* Vol. 49, No. 4, pp. 192 – 205, 1995.

[8] R. Knopp, *Coding and Multiple Access over Fading Channels*, Ph.D. Thesis, Ecole Polytechnique Fédérale de Lausanne, Lausanne, Switzerland, 1997.

[9] R. Knopp, P. A. Humblet, "Maximizing Diversity on Block Fading Channels," *Proceedings of ICC '97,* Montréal, Canada, 8 – 12 June, 1997.

[10] L. Ozarow, S. Shamai, and A. D. Wyner, "Information theoretic considerations for cellular mobile radio," *IEEE Trans. on Vehic. Tech.,* Vol. 43, No. 2, May 1994.

[11] G. Pottie and R. Calderbank, "Channel coding strategies for cellular mobile radio," *IEEE Trans. on Vehic. Tech.,* Vol. 44, No. 4, pp. 763 –769, Nov. 1995.

[12] J. Proakis, *Digital Communications.* New York: McGraw-Hill, 1983.

[13] T. Rappaport, *Wireless Communications.* Englewood Cliffs, NJ: Prentice-Hall, 1996.

[14] J. Salz, "Digital transmission over cross-coupled linear channels," AT&T *Tech. J.*, Vol. 64, no. 6, July-Aug. 1985.

[15] D. Tse and S. Hanly, "Linear Multiuser Receivers: Effective Interference, Effective Bandwidth and Capacity," Memorandum No. UCB/ERL M98/1,

UC Berkeley, Jan. 1998. Also: submitted for publication to *IEEE Trans. on Inform. Theory.*

[16] S. Verdú, “The capacity region of the symbol-asynchronous Gaussian multiple-access channel,” *IEEE Trans. on Inform. Theory,* Vol. 35, No. 4, pp. 733 – 751, July 1989.

[17] A. J. Viterbi, *CDMA – Principles of Spread-Spectrum Communications,* Reading, MA: Addison-Wesley, 1995.

[18] COST 207 Management Committee, *COST 207: Digital land mobile radio communications (Final Report).* Commission of the European Communities, 1989.

Ezio Biglieri was born in Aosta (Italy). He studied Electrical Engineering at Politecnico di Torino (Italy), where he received his Dr. Engr. degree in 1967.

From 1968 to 1975 he was with the Istituto di Elettronica e Telecomunicazioni, Politecnico di Torino, first as a Research Engineer, then as an Associate Professor (jointly with Istituto Matematico). In 1975 he was made a Professor of Electrical Engineering at the University of Napoli (Italy). In 1977 he returned to Politecnico di Torino as a Professor in the Department of Electrical Engineering. From 1987 to 1989 he was a Professor of Electrical Engineering at the University of California, Los Angeles. Since 1990 he has been again a Professor with Politecnico di Torino.

He has held visiting positions with the Department of System Science, UCLA, the Mathematical Research Center, Bell Laboratories, Murray Hill, NJ, the Bell Laboratories, Holmdel, NJ, the Department of Electrical Engineering, UCLA, the Telecommunication Department of the Ecole Nationale Supérieure de Télécommunications, Paris, France, and the University of Sydney, Australia.

He was elected three times to the Board of Governors of the IEEE Information Theory Society, in which he is now serving as its 1st Vice-President.

From 1988 to 1991 he was an Editor of the IEEE Transactions on Communications, from 1991 to 1994 an Associate Editor of the IEEE Transactions on Information Theory, and since 1997 an Editor of the IEEE Communications Letters. Since 1997 he has been the Editor in Chief of the European Transactions on Telecommunications. He has edited three books and co-authored five, among which the recent *Principles of Digital Transmission with Wireless Applications* (New York: Plenum, 1998).

Professor Biglieri is a Fellow of the IEEE.

Giuseppe Caire was born in Torino (Italy) on May 21, 1965. He received the B.Sc. in Electrical Engineering from Politecnico di Torino (Italy), in 1990, the M.Sc. in Electrical Engineering from Princeton University in 1992, and the Ph.D. from Politecnico di Torino in 1994. As a recipient of the AEI G. Someda Scholarship in 1991, he was with the European Space Agency (ESTEC, Noordwijk, The Netherlands) from May 1994 till February 1995. He was also a recipient of the COTRAO Scholarship in 1996 and of a CNR Scholarship in 1997. He visited the Institut Eurecom, Sophia Antipolis, France, in 1996, and Princeton University in Summer 1997. Presently, he is assistant professor at the Dipartimento di Elettronica of the Politecnico di Torino. He is co-author of more than 20 papers in international journals and more than 40 in international conferences, and he is author of three international patents filed with the European Space Agency.

His current interests are in the field of digital communications theory, information theory, coding theory and digital signal processing, with particular focus on wireless applications.

Giorgio Taricco received the "Laurea in Ingegneria Elettronica" (summa cum laude) from Politecnico di Torino in 1985. In 1986 he received the Andreoni Graduation Award. From 1985 to 1987 he was with CSELT (Torino, Italy) involved in the development of the coding subsystem of GSM. He holds two international patents on applications of Reed-Solomon co-decoding algorithms with error-and-erasure correction. Since 1988 he has been with Politecnico di Torino in the Department of Electrical Engineering. His current research interests are in the area of coded modulations and information theory applied to satellite and mobile radio systems.