System Capacity of F-TDMA Cellular Systems

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Abstract

We study the system capacity of cellular systems with time-division multiple access, slow time-frequency hopping (F-TDMA) and conventional single-user processing at the receivers. System capacity is formally defined as the maximum of the product of the number of users per cell times the user spectral efficiency, for a given maximum outage probability. We adopt an information-theoretic definition of outage as the event that the mutual information of the block-interference channel resulting from a finite number of signal bursts spanned by the transmission of a user code word falls below the actual code rate, because of fading, shadowing and interference. Starting from this definition, we develop a general framework which naturally takes into account many different aspects of F-TDMA cellular systems, like channel reuse, channel utilization, waveform design, time-frequency hopping, voice activity exploitation, handoff and power control strategies. Most importantly, our analysis does not rely on the choice of a particular coding scheme and can be applied to a very large class of systems in order to find guidelines for capacity-maximizing system design. A numerical example based on a typical urban mobile environment shows that there is a considerable capacity gap between actual F-TDMA systems and the limits predicted by our analysis. However, this gap can be filled by carefully designed "practical" systems, which make use of conventional single-user processing and simple coded modulation schemes.

Keywords: Cellular systems, block-interference and block-fading channels, information outage probability.

1 Introduction

Cellular wireless communications systems are the subject of a huge body of work devoted to assessing their ultimate information-theoretic limitations and to compare the efficiency of practical access schemes. We shall distinguish between *information-theoretic* approaches (see [1, 2] for a very comprehensive reference list) and *conventional* approaches (for example,[3, 4, 5, 6, 7]).

Information-theoretic analysis aims at determining the region of rates at which all users can communicate reliably. Wireless channels are characterized by fading whose dynamics depend on the mobile speed and may be much slower than the signaling rate. Without any constraint on the decoding delay and under an ergodic assumption on the fading processes, limiting performances are characterized in terms of average mutual information [8, 9, 10, 11, 12, 1, 2]. On the other hand, under a strict decoding delay constraint fading cannot be treated in an ergodic manner and performance is better characterized in terms of delay-limited capacity [13] or, more generally, in terms of code rate versus *information outage probability* [14, 15]. In all cases, information-theoretic approaches assume complete [11, 12] or partial [1, 2] joint processing (i.e., detection/decoding) of the received signals.

Despite its high performance, joint processing is still not implemented in current cellular standards [16] because of complexity, and may be regarded as an option for future-generation systems.[1] On the contrary, most actual systems (both FDMA/TDMA and CDMA) are based on suboptimal conventional single-user processing, treating either intra-cell and inter-cell interference as additive noise [16].[2]

Here we are interested in assessing the performance of conventional cellular systems. Therefore, we shall assume strictly single-user processing at the receiver, either in the uplink (mobiles to base-station) and in the downlink (base-station to mobiles). This reduces both the inherently different and difficult problems of uplink (multiple-access plus interference) and downlink (broadcast plus interference) to a simple equivalent single-user additive noise channel. We restrict our treatment to the case of intra-cell time-division multiple access with slow frequency-hopping (denoted as F-TDMA), and we briefly discuss an extension to the case of non-orthogonal intra-cell access, as for example CDMA, where also users in the same cell overlap in time and frequency.

Standard approaches to the performance analysis of conventional cellular systems introduce some pragmatic notion of *system capacity*, as a measure of the number of users per cell which can be supported with a given quality of service [3, 4, 5, 6]. Outage is defined as the event that the quality of service for the reference user falls below a given threshold. Outage probability is normally evaluated from the cumulative distribution function (cdf) of C/I (the carrier-to-interference plus noise ratio), as the probability that $C/I < (C/I)_{\rm th}$, where $(C/I)_{\rm th}$ is a threshold determined by the coding and modulation scheme used [5, 3, 7]. This is motivated by the fact that for digital speech transmission with frame-oriented source coding, quality of service can be related to the frame-error rate at the output of the channel decoder [16]. Assuming that each frame is independently encoded and transmitted, a basic quality of service indicator is

[1] A step towards this direction is represented by some recent results [17, 9, 13, 18] showing that all points in the achievable rate region of a Gaussian multiple-access channel (with or without fading) can be obtained by some form of single-user decoding with decision feedback and successive interference cancellation, with a complexity of the order of simple single-user decoding.

[2] Hybrid approaches involve some form of *multiuser detection* (see for example [19, 20] and references therein) followed by single-user independent decoding of each user [21, 22, 23].

the code word error rate (WER). However, constraints on the decoding delay are not captured by this analysis. In fact, the first-order statistics of C/I is not sufficient to characterize the variations of C/I as a function of time.

In this paper, we propose the use of *information outage probability* as the quality of service indicator for a conventional cellular system. In the case of delay-constrained constant-rate transmission and slowly-varying fading (e.g., in the important case of mobile telephony with terminals moving at walking speed, typical of most nowadays cellular situations), the transmission of a code word spans only a small number of fading realizations, so that the "instantaneous" mutual information of the channel is actually a random variable [24, 25]. Following [24], we define an outage as the event that the mutual information of the block-interference channel [26] resulting from a finite number of F-TDMA signal bursts spanned by the transmission of a user code word falls below the actual code rate, because of fading, shadowing and interference. We provide an operational characterization of information outage probability as the achievable WER averaged over the user random coding ensemble and over all the possible realizations of the channel state, as the burst length goes to infinity. Results with practical burst lengths and low-complexity codes show that outage probability is actually closely approached by the average WER of real systems [27, 28]. Hence, our results can be regarded as good estimates of the performance of practical systems with conventional single-user processing at the receiver.

We define the system capacity as the maximum of the product of the number of users/cell times the user spectral efficiency (bit/s/Hz), for a given outage probability. This makes the analysis independent of the coding scheme and allows a simple performance characterization directly in terms of the maximum allowable decoding delay. Expressions of the mutual information necessary for the outage probability evaluation are derived under the assumption that all users signals are Gaussian with flat power spectral density. The Gaussian assumption yields an upperbound to the minimum achievable outage probability [29, 1].

F-TDMA cellular systems are characterized in terms of fundamental system parameters, like the maximum decoding delay ΔT, the user bit-rate R_b, the total bandwidth W and the desired maximum outage probability P_{out}. System design options involve user coding rate, handoff and power control strategies, voice activity exploitation, time-frequency hopping codes and channel utilization, as well as the choice of the user signal bandwidth W_{s}, of the user signaling waveforms and of the reuse cluster size (see definitions in the model of Section 2). The net result of this analysis is a simple and flexible *evaluation tool*, allowing the optimization of many system parameters independently on the particular coding and interleaving scheme used and yielding guidelines for the design of good conventional systems.

The paper is organized as follows. In Section 2 we describe the model of the F-TDMA cellular system under analysis. In Section 3 we define outage probability and system capacity. Section 4 presents expressions for the mutual information needed to compute outage probability. In Section 5 we develop the details of a numerical example. Finally, in Section 6 we present our conclusions.

2 System model

For the time being, we do not distinguish between uplink and downlink, therefore we indicate by "user signals" the signals transmitted over both links, in general. We adopt baseband notation and we represent signals by their complex envelopes. Our system is characterized by a total bandwidth W (Hz), user information bit rate R_b (bit/s) and maximum decoding delay ΔT (s).

Cellular coverage and F-TDMA access. We consider a symmetric infinite (linear or planar) cellular coverage [11, 1, 2], cell 0 being the reference cell. Cells are grouped into *reuse clusters* of size K [30]. Users in cells belonging to the same cluster transmit over different carriers and/or over different time intervals, so that they are mutually orthogonal. Users in different clusters may interfere.

In F-TDMA, transmission is organized in *frames.* The region $(-\infty, \infty) \times [-W/2, W/2]$ of the time-frequency plane is divided into frames $[nT_{\rm f}, (n+1)T_{\rm f}] \times [-W/2, W/2]$ (for $n = \mathbb{Z}$). Each frame is partitioned into time-frequency *slots* of bandwidth $W_{\rm s}$ and duration $T_{\rm s}$. User signals are divided into bursts which occupy one slot. Guard bands and guard intervals are inserted in order to make signal bursts approximately time- and band-limited over the slots. The total number of slots in a frame is $Q = (T_{\rm f}/T_{\rm s})(W/W_{\rm s})$. The slots are equally assigned to the K cells in a cluster, so that each cell has a total of $N_{\rm c} = Q/K$ available slots per frame (we assume that K divides Q). Active users in the system transmit one signal burst per frame, over a given predetermined sequence of slots (hopping sequence). At each time, any user in a given cluster experiences interference from at most one user from each other cluster.

Some access protocol takes care of assigning a hopping sequence belonging to a given Q-ary orthogonal hopping code to each user entering the system. For cluster size K, an orthogonal hopping code can be partitioned into K mutually orthogonal subcodes of size N_c and a different subcode is assigned to each cell in the cluster. If the same hopping code is used in all clusters, two users in different clusters which are assigned the same hopping sequence may interfere over many consecutive bursts. As noted in [31], this highly correlated interference may create very unfavorable worst-case situations where a user suffers from persistently strong interference over a long time. This can be avoided by using different hopping codes in different clusters (a solution based on orthogonal latin squares is proposed in [31]). The ability of the system to randomize the interference over different bursts is referred to as *interferer diversity.*

Channel utilization and voice activity. Because of dynamic channel allocation strategies and/or in order to limit interference between co-channel cells, only a fraction $0 < u \leq 1$ of the $N_{\rm c}$ slots can be used at the same time in each cell, so that the maximum number of connected users (i.e., established calls) per cell is given by $N_{\rm u} = uN_{\rm c}$. The fraction u of slots in use is referred to as *channel utilization.*

In the case of speech transmission, because of voice activity [3], a signal burst may contain "silence". In this case, an option is to leave the slot empty in order to decrease the total interference level (we do not take into account the possibility that the empty slot can be used by another user, as for example in [32]).

Coding, interleaving and decoding. Code words are interleaved, partitioned into blocks and modulated into signal bursts. Because of real-time speech transmission, a maximum interleaving delay ΔT is imposed. The interleaving depth, i.e., the number of bursts spanned by a code word, is $M = \Delta T/T_{\rm f}$. Without loss of generality, we look at the concatenation of a block encoder (e.g., a TCM encoder with trellis termination) with an interleaver and a burst modulator as a single encoder that maps input sequences of length K_b of binary i.i.d. equiprobable random variables onto output sequences of length $N_{\rm s}M$ with elements in the complex signal set $\mathcal{X}$ with normalized average energy per symbol equal to 1 (e.g., $\mathcal{X}$ can be a PSK or a QAM constellation). The output sequence is divided into M blocks of $N_{\rm s}$ symbols, each forming a distinct signal burst. The overall code rate is $\rho = \frac{K_b}{N_{\rm s}M}$ bit/symbol. Following [24], we assume that the interleaving depth M is a small integer while the number of symbols per burst N_s is large, i.e., we are interested in analyzing the system for $N_\mathrm{s} \to \infty$ and finite (small) M.

As anticipated in Section 1, we consider a conventional system with strictly single-user processing. Receivers (either base-station and mobile terminal) are formed by a filter matched to the user of interest, followed by a single-user decoder.

Propagation channel, handoff and power control. The propagation channel is characterized by a frequency-flat gain that takes into account the distance between transmitter and receiver and the effect of *shadowing*, and by a frequency-selective time-varying transfer function that models the *multipath fading* [16].

The frequency-flat gain is a slowly-varying random process and will be considered constant (but random) over a time interval of duration ΔT. Multipath is modeled as a *wide-sense stationary uncorrelated scattering* fading characterized by the impulse response $c(\tau, t)$ [33, Ch. 14]. Since we are interested in slowly-varying fading, typical of most practical cellular situations [24], we make a block-fading [3] approximation [24, 25] and consider $c(\tau, t)$ as constant with respect to t over intervals of duration T_s. Hence, the transmission of M bursts spanned by a code word is characterized by the M impulse responses $\{c_m(\tau) : m = 0, \ldots, M-1\}$ or, equivalently, by the M frequency responses $\{C_m(f) : m = 0, \ldots, M-1\}$, where $C_m(f)$ is the (continuous-time) Fourier transform of $c_m(\tau)$. Although the analysis in the following can be applied to arbitrary fading statistics, for simplicity we assume that $c_m(\tau)$ is complex Gaussian circularly-symmetric [34] with $E[c_m(\tau)] = 0$ and $E[|c_m(\tau)|^2] = \sigma^2(\tau)$. The function $\sigma^2(\tau)$ is referred to as the *multipath intensity profile* of the channel [33, Ch. 14], and it is assumed to be normalized as $\int \sigma^2(\tau) d\tau = 1$. In this way, $|C_m(f)|$ is Rayleigh distributed with $E[|C_m(f)|^2] = 1$ for all f.

Handoff and power control strategies can be incorporated in the model by suitably modifying the statistics of the frequency-flat gain [3, 5]. In any case, no attempt is made to compensate for frequency-selective fading and no *rate* adaptation is considered (as for example in [35]).

Equivalent discrete-time channel model. We assume that the i-th reuse cluster, for $i = 0, 1, \ldots$, is characterized by a unit-energy *user waveform* $s^i(t)$, which may be different for different clusters. Users connected with cells of cluster i employ linear modulation [33] with $s^i(t)$ as elementary waveform. The *waveform spreading factor* is defined as $L = W_\mathrm{s} T$ and it is assumed common to all the $s^i(t)$'s. Hence, the number of symbols (independent Shannon dimensions) in a signal burst is $N_\mathrm{s} = \eta W_\mathrm{s} T_\mathrm{s} / L$, where $\eta \leq 1$ is a factor which takes into account the F-TDMA overhead (guard intervals, guard bands, header and tail symbols and training sequences for channel identification and equalization). From now on we disregard the F-TDMA overhead, which is automatically taken into account by the factor η, and we consider only the information-bearing part of user signals. The signal transmitted by user i over the m-th burst can be written as

$$x^i_m(t) = \sum_{k=0}^{N_\mathrm{s}-1} x^i_m[k] s^i(t - mT_\mathrm{f} - t_m - kT - \tau^i_m) e^{j(2\pi f_m t + \phi^i_m)} \tag{1}$$

[3]The block-fading assumption holds approximately if the product of the channel Doppler bandwidth B_d times the slot duration T_s is ≤ 1 [33], which is verified for low mobile speed [24]. With frequency-hopping, provided that the carrier separation is larger than the channel coherence bandwidth [33], the block-fading assumption with approximately independent blocks holds also if $B_d T_\mathrm{s}$ is much less than 1.

where $x_m^i[k] \in \mathcal{X}$ are complex modulation symbols, τ_m^i, ϕ_m^i are delays and carrier phases and where t_m and f_m are the initial epoch and the carrier frequency of the signal burst m. Without loss of generality we let $\tau_m^0 = \phi_m^0 = 0$ for all m and, since transmission is frame-synchronous, $|\tau_m^i| < T/2$.

Now, we focus on receiver 0 (which can be the mobile or the base-station, depending on which link we are considering). Let A^0 and A_m^i denote the total amplitude gains from transmitter 0 to receiver 0 and from transmitter i to receiver 0, respectively, during the m-th burst. Note that many system features, as interferer diversity, voice activity, channel utilization, handoff and power control strategies, are included in the model by choosing the appropriate joint statistics of the coefficients A^0 and A_m^i (see examples in Section 5). Let $c_m^i(\tau)$ be the multipath channel impulse response from transmitter $i \geq 0$ to receiver 0 during the m-th burst. Finally, we can write the signal at receiver 0 in the m-th burst as

$$y_m(t) = A^0 c_m^0(t) \star x_m^0(t) + \sum_{i=1}^{N-1} A_m^i c_m^i(t) \star x_m^i(t) + n(t) \tag{2}$$

where $\star$ denotes convolution and $n(t)$ is a white circularly-symmetric complex Gaussian noise with (two-sided) power spectral density N_0.

Receiver 0 is a single-user matched filter with perfect knowledge of the channel responses $\{c_m^0(\tau) : m = 0, \ldots, M-1\}$. Let $h_m^i(t) = \left(c_m^i(t)e^{-j2\pi f_m t} \star s^i(t)\right) e^{j\phi_m^i}$. After demodulation (multiplication by $e^{-j2\pi f_m t}$), filtering by $h^0(-t)^*$ and sampling at epochs $kT + mT_{\mathrm{f}} + t_m$ ($k = 0, \ldots, N_{\mathrm{s}} - 1$), we can write the k-th sample of the m-th received burst as

$$y_m[k] = \sum_j A^0 p_m^0[j] x_m^0[k-j] + \sum_{i=1}^{N-1} A_m^i \sum_j p_m^i[j] x_m^i[k-j] + \nu_m[k] \tag{3}$$

where

$$p_m^i[k] = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} c_m^0(\tau)^* c_m^i(\tau') e^{j(2\pi f_m(\tau-\tau')+\phi_m^i)} r_s^i(kT + \tau - \tau' - \tau_m^i) d\tau d\tau' \tag{4}$$

and where we have defined the 0-th and i-th waveforms cross-correlation function

$$r_s^i(\tau) = \int_{-\infty}^{\infty} s^0(t-\tau)^* s^i(t) dt \tag{5}$$

The noise samples $\nu_m[k]$ are obtained as $\nu_m[k] = \int_{-\infty}^{\infty} h_m^0(t-kT)^* n(t) dt$ and have autocorrelation sequence defined by $E[\nu_m[j]\nu_m[j-k]^*] = N_0 p_m^0[k]$. The discrete channel model (3) is the starting point of our analysis.

3 System capacity versus outage probability

Since users transmit signal bursts of $N_{\mathrm{s}} = \eta W_{\mathrm{s}} T_{\mathrm{s}}/L$ symbols every T_{f} seconds with a code rate ρ, the user bit-rate is given by $R_b = \rho\eta W_{\mathrm{s}} T_{\mathrm{s}}/(LT_{\mathrm{f}}) = \rho\eta W/(LKN_{\mathrm{c}})$. Hence, the number of users per cell of a system with channel utilization u is given by

$$N_{\mathrm{u}} = \frac{u\rho}{LK}\frac{\eta W}{R_b} \tag{6}$$

The term $\eta W/R_b$ is just a scale factor that depends on the desired user bit-rate, on the available system bandwidth and on the necessary F-TDMA overhead. Then, it is intuitive to define the system capacity as the ratio $u\rho/(LK)$ users/cell×bit/s/Hz. In the following, we clarify this concept and we formally define system capacity in terms of outage probability.

As noted in [24], under the strict decoding delay constraint ΔT and under the block-fading assumption, the channel cannot be treated in an ergodic stationary fashion. This is because a code word spans a small number M of "channel realizations" (intended here as the realization of the random variables $A_m^i, \tau_m^i, \phi_m^i$, f_m, t_m and of the random channel responses $c_m^i(\tau)$), although the code word length $N_s M$ is very long. Even if all the processes involved during the transmission of a code word are stationary and ergodic, the capacity of this channel in a strict Shannon sense might be zero, since the mutual information between the input (transmitter 0) and the output (receiver 0) during a time span ΔT is a random variable with a certain probability of being below any specified rate $\rho > 0$.

For the sake of notational simplicity, we define the channel state S_m that collects all the random variables which determine the discrete-time channel (3) and we let $\mathbf{S} = \{S_m : m = 0, \dots, M-1\}$ denote the sequence of channel states over the M bursts spanned by a code word. Let $I_M(\mathbf{S})$ denote the "instantaneous" conditional mutual information (in bit/symbol) of the M-block channel as $N_s \to \infty$

$$I_M(\mathbf{S}) = \lim_{N_s \to \infty} \frac{1}{MN_s} I\left(\bigcup_{m=0}^{M-1} \{x_m^0[k]\}_{k=0}^{N_s-1}; \bigcup_{m=0}^{M-1} \{y_m[k]\}_{k=0}^{N_s-1} \middle| \mathbf{S} = \mathbf{S} \right) \tag{7}$$

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

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

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 $p(\mathbf{y}|\mathbf{S})$ conditioned on $\mathbf{S}$ (note that, since $\mathbf{S}$ is random, $I(\mathbf{X};\mathbf{Y}|\mathbf{S} = \mathbf{S})$ is a random variable). The standard conditional average mutual information is obtained by averaging $I_M(\mathbf{S})$ with respect to $\mathbf{S}$.

Assume a code rate ρ bit/symbol and let $\overline{P}_{e|\mathbf{S}}(\rho)$ denote the code word error rate (WER) averaged over the code ensemble of all codes with rate ρ and length $N_s M$, randomly generated according to a given 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 [29, 36]) we can write

$$\lim_{N_s \to \infty} \overline{P}_{e|\mathbf{S}}(\rho) = \mathcal{I}_{\{I_M(\mathbf{S}) < \rho\}} = \begin{cases} 0 & \text{if} \quad I_M(\mathbf{S}) \geq \rho \\ 1 & \text{if} \quad I_M(\mathbf{S}) < \rho \end{cases} \tag{8}$$

($\mathcal{I}_{\mathcal{A}}$ denotes the indicator function of the event $\mathcal{A}$). By averaging $\overline{P}_{e|\mathbf{S}}(\rho)$ with respect to $\mathbf{S}$ and exchanging limit with expectation, we can write [28]

$$\lim_{N_s \to \infty} \overline{P}_e(\rho) = \lim_{N_s \to \infty} E[\overline{P}_{e|\mathbf{S}}(\rho)] = E[\mathcal{I}_{\{I_M(\mathbf{S}) < \rho\}}] = P(I_M(\mathbf{S}) < \rho) = P_{\text{out}}(\rho) \tag{9}$$

where we have defined the outage probability for a given code rate as $P_{\text{out}}(\rho) = P(I_M(\mathbf{S}) < \rho)$ [24]. Equation (9) provides an operational meaning to the information-theoretic outage probability defined above: namely, $P_{\text{out}}(\rho)$ is equal to the WER averaged over the random coding ensemble and over all the possible channel realizations $\mathbf{S}$, in the limit of large N_s.[4]

[4]To be precise, in order to approach $P_{\text{out}}(\rho)$ the burst length N_s should grow to infinity, thus invalidating the assumption of delay constrained transmission. Nevertheless, already for $N_s \simeq 100$ the outage probability predicts surprisingly well the word error probability of good practical codes, as shown in [27, 28]. Burst lengths of this order of magnitude can be regarded as typical of existing cellular standards. For example, in the GSM standard the burst length is $N_s = 114$ symbols [16].

Now we are ready for the following definition:
F-TDMA system capacity. Consider a F-TDMA cellular system as described in Section 2, with reuse K, channel utilization u, user waveforms with spreading L, interleaving depth M. Then, the system capacity under an outage probability constraint $P_{\rm out}$ is given by

$$C_{\rm sys} = \frac{u}{LK} \sup\{\rho \geq 0 \ : \ P(I_M(\mathbf{S}) < \rho) \leq P_{\rm out}\} \qquad \text{user/cell}\times\text{bit/s/Hz} \tag{10}$$

□

Non-orthogonal intra-cell access. From the point of view of a *conventional* system, where strictly single-user processing is employed, the main difference between F-TDMA and a non-orthogonal intra-cell access (e.g., CDMA) is that, with the latter, any user in a given cluster may experience interference from several users in its own cluster plus several users in other clusters [1, 2]. As a consequence, while the expression of the received signal (3) for F-TDMA depends only on the number of interfering cells N, an analogous expression for non-orthogonal access depends explicitly both on N and on the number of users per cell $N_{\rm u}$. On the other hand, users can access the channel at any time and can overlap over the whole system bandwidth W, so that $N_{\rm u}$ is not directly related to the code rate ρ by a relation such as (6).

With the same assumptions made before about single-user processing, total delay and block-fading, for non-orthogonal access we can define an equivalent single-user channel from transmitter 0 to receiver 0 with instantaneous mutual information $I_M(\mathbf{S}, N_{\rm u})$, that explicitly depends on $N_{\rm u}$ as a parameter. The user code word block length is $N_{\rm s}M = \eta\Delta TW/L$ and the user spectral efficiency is simply given by R_b/W bit/s/Hz. The number of users/cell×bit/s/Hz is given by $N_{\rm u}R_b/W = \eta N_{\rm u}\rho/L$. Then, in analogy with what done before for F-TDMA, the system capacity with non-orthogonal intra-cell access under an outage probability constraint is given by

$$C_{\rm sys} = \frac{1}{L} \sup\left\{\rho N_{\rm u} \ : \ P\left(I_M(\mathbf{S}, N_{\rm u}) < \rho\right) \leq P_{\rm out}\right\} \qquad \text{user/cell}\times\text{bit/s/Hz} \tag{11}$$

Conventional F-TDMA and non-orthogonal access systems can be compared in terms of $C_{\rm sys}$ defined in (10) and in (11), respectively. However, this comparison is well beyond the scope of this paper. Results for coded direct-sequence CDMA have been shown in [22] in the case of a single-cell system with frequency-selective Rayleigh fading. We hasten to say that the whole picture changes radically if joint processing at the receiver is allowed (see [1, 2] and references therein).

In the following, we concentrate on F-TDMA and we provide expressions of the mutual information $I_M(\mathbf{S})$ for the computation of $C_{\rm sys}$ given in (10).

4 Mutual information

Usually, transients due to multipath linear distortion are absorbed into guard intervals, in order to avoid interference with bursts transmitted over the same carrier in adjacent time slots. Here we make this assumption, so that the output blocks $\{y_m[k] : k = 0, \ldots, N_{\rm s} - 1\}$ can be treated as conditionally independent given the input and given the sequence of channel realizations $\mathbf{S}$. Then, the mutual information can be written as

$$I_M(\mathbf{S}) = \frac{1}{M}\sum_{m=0}^{M-1} \lim_{N_{\rm s}\to\infty} \frac{1}{N_{\rm s}} I\left(\{x_m^0[k]\}_{k=0}^{N_{\rm s}-1}; \{y_m[k]\}_{k=0}^{N_{\rm s}-1} \Big| \, S_m = S_m\right) \tag{12}$$

In order to proceed further, we assume that the impulse responses $p_m^i[k]$ in (4) have finite energy (that is, $\{p_m^i[k]\} \in l_2(\mathbb{Z})$ for all i, m, with probability 1) and that the user symbols are complex circularly-symmetric Gaussian i.i.d. random variables with $E[x_m^i[k]] = 0$ and $E[|x_m^i[k]|^2] = 1$.

This choice of the input distribution may be motivated as follows:

i) Under the input constraint $E[|x_m^i[k]|^2] \leq 1$ and without delay constraints (i.e., $M \to \infty$), this is the capacity-achieving input distribution in the single-user case ($N = 1$) [24] and in the multiuser case with optimal joint processing of the signals received at all base-stations [11, 12], when the transmitters have no knowledge of the instantaneous channel responses (if the transmitters know all channel responses $A_m^i p_m^i[k]$, the optimal input distribution is still complex circularly-symmetric Gaussian but the best power allocation is obtained by multiuser water-filling [37, 10, 9]).

ii) The Gaussian interference assumption provides a lower bound on the maximum achievable instantaneous mutual information $I_M(\mathbf{S})$ [29, 1], so that the resulting outage probability is an upper bound to the minimum outage probability of a conventional system.

With single-user processing and with the above assumptions, the channel (3), conditioned on the channel state sequence $\mathbf{S}$, reduces to a standard (single-user) additive colored Gaussian noise channel with memory. Hence, by computing explicitly the term inside the limit for $N_{\rm s} \to \infty$ in (12) and by applying the Toeplitz eigenvalue distribution theorem [38] (the finite-energy assumption of the channel impulse responses allows us to do that), we obtain the mutual information in terms of the power spectral densities $Y_m(\theta)$ and $Z_m(\theta)$ [5] of the received sequence $y_m[k]$ and of the equivalent noise sequence $z_m[k] = \sum_{i=1}^{N-1} A_m^i \sum_j p_m^i[j] x_m^i[k-j] + \nu[k]$, respectively (see [24] and references therein). We get

$$I_M(\mathbf{S}) = \frac{1}{M} \sum_{m=0}^{M-1} \int_{-1/2}^{1/2} \log_2 \left(\frac{Y_m(\theta)}{Z_m(\theta)} \right) d\theta \tag{13}$$

where

$$\begin{aligned} Y_m(\theta) &= |A^0|^2 |P_m^0(\theta)|^2 + Z_m(\theta) \\ Z_m(\theta) &= \sum_{i=1}^{N-1} |A_m^i|^2 |P_m^i(\theta)|^2 + N_0 P_m^0(\theta) \end{aligned} \tag{14}$$

and where

$$P_m^i(\theta) = \sum_k p_m^i[k] e^{-j2\pi\theta k} \tag{15}$$

$P_m^0(\theta)$ is real and non-negative. By assuming $P_m^0(\theta) > 0$ for all $\theta \in [-1/2, 1/2]$, we can divide both $Y_m(\theta)$ and $Z_m(\theta)$ by $P_m^0(\theta)$ and obtain

$$I_M(\mathbf{S}) = \frac{1}{M} \sum_{m=0}^{M-1} \int_{-1/2}^{1/2} \log_2 \left(1 + \frac{|A^0|^2 P_m^0(\theta)}{\sum_{i=1}^{N-1} |A_m^i|^2 |P_m^i(\theta)|^2 / P_m^0(\theta) + N_0} \right) d\theta \tag{16}$$

4.1 Strictly band-limited waveforms

In this section we look for simpler expressions for $I_M(\mathbf{S})$ directly in terms of the fading channel frequency responses $C_m^i(f) = \int_0^{T_d} c_m^i(\tau) e^{-j2\pi f \tau} d\tau$ ($i = 0, \ldots, N-1$, $m = 0, \ldots, M-1$). To this end, from (4) and (15) we write

$$P_m^i(\theta) = e^{j\phi_m^i} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} c_m^0(\tau)^* c_m^i(\tau') e^{j2\pi f_m(\tau-\tau')} R_m^i(\theta; \tau, \tau') d\tau d\tau' \tag{17}$$

[5] We use f and θ to denote the continuous-time and the discrete-time Fourier frequencies, respectively.

where we define

$$
\begin{aligned}
R^i_m(\theta;\tau,\tau') &= \sum_k r^i_s(kT+\tau-\tau'-\tau^i_m)e^{-j2\pi\theta k} \\
&= \frac{1}{T}\sum_k S^0\left(\frac{\theta+k}{T}\right)^* S^i\left(\frac{\theta+k}{T}\right) e^{j2\pi(\theta+k)(\tau-\tau'-\tau^i_m)/T}
\end{aligned} \tag{18}
$$

and where $S^i(f)$ is the Fourier transform of $s^i(t)$. The RHS in the last line of (18) is the so called "folded cross-spectrum" that appears in the frequency characterization of linear equalizers [33] and can be obtained by applying Parseval's identity to (5). Then, we specialize our treatment to user waveforms which avoid spectral folding. For simplicity of notation, let $L = 2L'+1$ be an odd integer (it is immediate to check that an analogous final result also holds for L even) and we consider the strictly Nyquist band-limited user waveforms

$$
s^i(t) = \sqrt{\frac{T}{L}}\frac{\sin(\pi t/T)}{\pi t}\sum_{\ell=-L'}^{L'} e^{j(2\pi\ell t/T+\psi^i_\ell)} \tag{19}
$$

where $\{\psi^i_\ell : \ell = -L',\dots,L'\}$ is a sequence of phases in $[-\pi,\pi]$ which can be regarded as the *signature sequence* of the i-th cluster. Without loss of generality we let $\psi^0_\ell = 0$ for all ℓ.

First, we derive the expression of the mutual information and then we comment on the practical relevance of user waveforms (19). The Fourier transform of $s^i(t)$ is $S^i(f) = \sqrt{\frac{T}{L}}\sum_{\ell=-L'}^{L'}\Pi(T(f-\ell/T))e^{j\psi^i_\ell}$ where $\Pi(f)=1$ for $|f|<1/2$ and zero for $|f|>1/2$. By using $S^i(f)$ in (18) we get

$$
R^i_m(\theta;\tau,\tau') = e^{j2\pi\theta(\tau-\tau'-\tau^i_m)/T}\frac{1}{L}\sum_{\ell=-L'}^{L'} e^{j(2\pi\ell(\tau-\tau'-\tau^i_m)/T+\psi^i_\ell)} \qquad \text{for } \theta\in(-1/2,1/2) \tag{20}
$$

Finally, by substituting (20) in (17) we write $P^i_m(\theta)$ as

$$
P^i_m(\theta) = e^{-j(2\pi\theta\tau^i_m/T-\phi^i_m)}\mathbf{c}^0_m(\theta)^\dagger\mathbf{c}^i_m(\theta) \tag{21}
$$

($\dagger$ denotes Hermitian transpose) where we define the column L-vector $\mathbf{c}^i_m(\theta)$ with ℓ-th element

$$
\left[\mathbf{c}^i_m(\theta)\right]_\ell = \frac{1}{\sqrt{L}}C^i_m\left(\frac{\theta+f_mT-\ell}{T}\right)e^{j(2\pi\ell\tau^i_m/T+\psi^i_\ell)} \qquad \text{for} \quad \ell=-L',\dots,L'
$$

Now, we use (21) into (16) and we get the desired expression as

$$
I_M(\mathbf{S}) = \frac{1}{M}\sum_{m=0}^{M-1}\int_{-1/2}^{1/2}\log_2\left(1+\frac{|A^0|^2|\mathbf{c}^0_m(\theta)|^2}{\sum_{i=1}^{N-1}|A^i_m|^2\frac{\left|\mathbf{c}^0_m(\theta)^\dagger\mathbf{c}^i_m(\theta)\right|^2}{|\mathbf{c}^0_m(\theta)|^2}+N_0}\right)d\theta \tag{22}
$$

For practical computation purposes we may discretize the integral in (22). We assume that all the channels $c^i_m(\tau)$ have the same coherence bandwidth B_c [33]. Hence, $C^i_m(f)$ can be well approximated by a piecewise constant function with stepsize smaller than B_c. Let D denote an odd integer $\geq 1/(B_cT)$ (again, an analogous result holds for D even) and define the column L-vector $\mathbf{c}^i_m[j]$ as $\mathbf{c}^i_m(\theta)$ evaluated at $\theta=(j-(D-1)/2)/D$, for $j=0,\dots,D-1$. Conditionally on the signature sequences and channel states, for all m and j the $\mathbf{c}^i_m[j]$'s are complex jointly Gaussian random vectors with circularly-symmetric entries, mean zero and whose covariance matrix can be obtained from the time-frequency fading autocorrelation function [33]. Eventually, $I_M(\mathbf{S})$ can be approximated by

$$I_M(\mathbf{S}) \simeq \frac{1}{MD} \sum_{m=0}^{M-1} \sum_{j=0}^{D-1} \log_2 \left(1 + \frac{|A^0|^2 |\mathbf{c}_m^0[j]|^2}{\sum_{i=1}^{N-1} |A_m^i|^2 \frac{\left|\mathbf{c}_m^0[j]^\dagger \mathbf{c}_m^i[j]\right|^2}{|\mathbf{c}_m^0[j]|^2} + N_0} \right) \tag{23}$$

Remark on user waveforms. Waveforms $s^i(t)$ defined in (19) correspond to orthogonal frequency division multiplexing (OFDM) [39], where the same symbol $x_m^i[k]$ is transmitted over L adjacent subbands spaced by $1/T$, and where a different phase ψ_ℓ^i is given to each ℓ-th subcarrier. In this way, a symbol is spread over a bandwidth $L/T = W_\mathrm{s}$. In a real implementation, the sequences $\{\psi_\ell^i\}$ can be obtained by the same pseudorandom generator, initialized by a different seed for each cluster. Users accessing a cell in a certain cluster are given the corresponding generator seed (in the simulations of Section 5 the sequences $\{\psi_\ell^i\}$ are generated with i.i.d. components, uniformly distributed over $[-\pi, \pi]$, independent for different i's). We believe that waveforms (19) are able to capture the main effects of spreading, still yielding simple expressions for the mutual information. Then, in the following, we restrict our treatment to this case.

Remark on previous results. Expression (22) can be regarded as the generalization of a previous result obtained for the particular case of unspread signals ($L = 1$), single-user ($N = 1$) and ideally power controlled transmission ($A^0 = 1$) given in [24]. In this case, for a given signal-to-noise ratio $\gamma = 1/N_0$ (recall that the alphabet $\mathcal{X}$ has unit average energy per symbol), by letting $W_\mathrm{s} = 1/T$ and $f = \theta W_\mathrm{s}$ we obtain

$$I_M(\mathbf{S}) = \frac{1}{M} \sum_{m=0}^{M-1} \frac{1}{W_\mathrm{s}} \int_{-W_\mathrm{s}/2}^{W_\mathrm{s}/2} \log_2(1 + \gamma |C_m^0(f)|^2) df$$

The average conditional mutual information I is obtained by averaging $I_M(\mathbf{S})$ with respect to the channel realizations. We obtain

$$I = E[I_M(\mathbf{S})] = \log_2(e) \int_0^\infty \log(1 + \gamma v) e^{-v} dv = -\log_2(e) e^{1/\gamma} E_i(-1/\gamma)$$

where we used the fact that, for each f, $|C_m^0(f)|^2$ is exponentially distributed with mean 1 and where $E_i(x)$ is the exponential integral function [40, Sec. 4.337] (the above expression was also obtained in [41, 42]).

With no interference, outage probability can be expressed in a simple closed form in the case of $M = 1$ and a flat fading channel ($C_0^0(f) = C_0^0$, independent of f or, equivalently, when $W_\mathrm{s} \leq B_c$). We obtain [24]

$$P_{\mathrm{out}}(\rho) = 1 - \exp(-(e^{\rho/\log_2(e)} - 1)/\gamma)$$

Generalizations (always with no interference) in the case $M = 2$ with a simple two-path frequency selective fading channel can be found in [24].

4.2 Outage probability for vanishing code rate

In principle, the outage probability $P_{\text{out}}(\rho)$ is immediately obtained from the cdf of $I_M(\mathbf{S})$. Unfortunately, this is difficult, if not impossible, to obtain in closed form even in the simple cases with $L = 1$. Hence, we have to resort to some numerical method (e.g., Monte Carlo simulation).

An exception is represented by the case of very low coding rate ($\rho \to 0$). As it is clear from definition (10), for vanishing ρ the system capacity is also vanishing. Nevertheless, for some applications it might be interesting to study the limit of $P_{\text{out}}(\rho)$ as $\rho \to 0$, for a given ratio $\gamma_b = \mathcal{E}_b/N_0$, where $\mathcal{E}_b$ is the average received energy per bit in the case of ideal power control (i.e., for $|A^0| = 1$). Since $\mathcal{X}$ has unit average energy per symbol, for a code rate ρ we get $\gamma_b = 1/(\rho N_0)$. The limit of the outage probability $\underline{P}_{\text{out}}$ can be written as

$$\underline{P}_{\text{out}} = \lim_{\rho\to 0} P(I_M(\mathbf{S}) < \rho) = \lim_{\rho\to 0} E[\mathcal{I}_{\{I_M(\mathbf{S})<\rho\}}] = E[\lim_{\rho\to 0} \mathcal{I}_{\{I_M(\mathbf{S})<\rho\}}] = P(\lim_{\rho\to 0} I_M(\mathbf{S})/\rho < 1) \quad (24)$$

By letting $N_0 = 1/(\gamma_b\rho)$ in (23) we get the limit

$$\lim_{\rho\to 0} I_M(\mathbf{S})/\rho = \frac{\gamma_b \log_2(e)|A^0|^2}{MD} \sum_{m=0}^{M-1} \sum_{j=0}^{D-1} |\mathbf{c}_m^0[j]|^2 \quad (25)$$

Finally, in the case of ideal power control ($|A^0|^2 = 1$), we use (25) in the RHS of (24) and we obtain

$$\underline{P}_{\text{out}} = P\left(\sum_{m=0}^{M-1} \sum_{j=0}^{D-1} |\mathbf{c}_m^0[j]|^2 < \frac{MD}{\gamma_b \log_2(e)} \right) = P(\mu < \varepsilon) \quad (26)$$

The limiting outage probability is given by the cdf of μ evaluated in ε. This can be easily computed by recognizing that μ is a quadratic form of correlated complex Gaussian random variables [43].

In general, it is interesting to observe that for $\rho \to 0$ the effect of interference is eliminated (both μ and ε are independent of interference) and C_{sys} is vanishing.

5 Numerical results

In the framework developed in the previous sections we can compare different system parameter choices in terms of C_{sys}. Our approach yields a fair comparison provided that the assumptions of the underlying model hold. These must be verified case-by-case.

We consider a planar regular hexagonal coverage with $k_t = 3$ tiers of interfering cells around reference cell 0. For simplicity, we assume omni-directional antennas, although simple fixed spatial filtering (e.g., cell sectorization [16, 3, 5]) could be easily included in the model. From the geometry of the hexagonal coverage [30], the number of interfering users is $N = 1+3k_t(k_t+1)$ and the possible reuse cluster sizes are given by $K = i^2 + ij + j^2$ for all non-zero integer pairs (i, j).

In order to make our results independent of the specific hopping codes, we consider the following interferer diversity limiting cases: i) Pairs of users interfere over all consecutive bursts (*no interferer diversity*). ii) A user experiences interference from different (independent) users over different bursts (*maximum interferer diversity*).

We consider two idealized handoff strategies: i) Each mobile is instantaneously connected to the closest base-station (*distance-driven handoff* (DDH)). ii) Each mobile is instantaneously connected with the base-station with the most favorable total path gain (distance plus shadowing) among a set of close base-stations (*path-driven handoff* (PDH)). As for power control, we consider the following cases: i) Constant transmitted power (*no power control*). ii) Each transmitter knows ideally the frequency-flat channel gain and compensates for it perfectly (*ideal slow power control*).

We model voice activity of the i-th interferer in burst m by an i.i.d. Bernoulli random variable β_m^i with $P(\beta_m^i = 1) = p$ (for user 0, $\beta_m^0 = 1$ with probability 1, since the outage probability must be evaluated only when user 0 actually transmits a code word). $p = 1$ models either the case where voice activity is not exploited and the case of continuous data transmission. Voice activity of each individual user is typically modeled by a Markov chain with two states (the "on" and the "off" state) [44]. Here, the i.i.d. Bernoulli model is motivated by the fact that, with random hopping codes, user 0 interferes with randomly selected sets of $N-1$ statistically independent users in each burst. Hence, by assuming stationarity of all voice activity processes, p can be interpreted as the *stationary probability* of the "on" state of the voice activity Markov chain, assumed to be the same for all users. [6] Channel utilization u can be incorporated in the model by replacing p with the product up.

5.1 Monte Carlo simulation

Monte Carlo simulation is used in order to evaluate the outage probability $P(I_M(\mathbf{S}) < \rho)$. In the following example we focus on the uplink. The main system parameters are summarized in Table 1 [24, 3, 5].

Path gain statistics. The frequency-flat channel gain is given by $d^{-\alpha}\xi$, where d is the distance between transmitter and receiver and $\xi = 10^{\sigma_{\rm sh}\zeta/10}$, with $\zeta \sim \mathcal{N}(0,1)$, is a log-normal random variable modeling the shadowing [16]. We represent the coverage area in the two-dimensional plane and define the set of interfering cell centers by $Z = \{z^i : i = 0, \ldots, N-1\}$, where $z^0 = 0$ is the reference cell center (the cell radius is normalized to 1). With DDH, the Bernoulli random variables β_m^i are generated i.i.d. according to the voice activity distribution $P(\beta_m^i = 1) = p$ (or up, in the case $u < 1$). Then, for all $i > 0$ and $m = 0, \ldots, M-1$ for which $\beta_m^i = 1$, the position x_m^i of mobile i during burst m is generated uniformly over a disk of radius 1 centered in z^i, together with two log-normal random variables ξ_m^i and ξ_m^{i0}. The first represents the shadowing from mobile i to cell center z^i, while the second represents the shadowing from mobile i to cell center z^0.

In the case of no interferer diversity, β_m^i, x_m^i, ξ_m^i and ξ_m^{i0} are constant with m, while in the case of maximum interferer diversity they are independent for different m's. For user 0, the position x^0 and the log-normal gain ξ^0 common to all M bursts are generated. Then, $|A^0|^2$ and $|A_m^i|^2$ (for $i > 0$) are given by

$$|A^0|^2 = \begin{cases} \xi^0 |x^0 - z^0|^{-\alpha} & \text{no power control} \\ 1 & \text{ideal power control} \end{cases}$$

[6] With maximum interferer diversity, interference comes from M mutually disjoint sets of $N-1$ users on the M bursts spanned by a user 0 code word. Then, for $N_{\rm u} > M$, the Bernoulli model is exact. With no interference diversity, interference comes from *the same* set of $N-1$ users for all M bursts. Then, the Bernoulli model is exact only if code words from different users are time-aligned (in other words, if an interferer is "on" on the first burst, it will stay "on" for all the M bursts). Since talkspurts are relatively long with respect to the code word duration, this model is expected to provide accurate results also for the more realistic case where code words are not time-aligned.

$$|A^i_m|^2 = \begin{cases} \beta^i_m \xi^{i0}_m |x^i_m - z^0|^{-\alpha} & \text{no power control} \\ \beta^i_m \frac{\xi^{i0}_m}{\xi^i_m} \left(\frac{|x^i_m - z^0|}{|x^i_m - z^i|}\right)^{-\alpha} & \text{ideal power control} \end{cases} \quad (27)$$

PDH is implemented as follows [5]. Let $\mathcal{C}_i$ denote the set formed by cell i and by its 6 surrounding adjacent cells, which can be either in the same cluster or in a different cluster. Mobile i in burst m, with position x^i_m, is actually connected with cell i if the event $\Omega^i_m = \{i = \arg\max_{j \in \mathcal{C}_i} \{\xi^j_m |x^i_m - z^j|^{-\alpha}\}\}$ occurs. (ξ^j_m are mutually independent log-normal random variables and z^j are the centers of the cells in $\mathcal{C}_i$). In [5] it is shown that $P(\Omega^i_m)$ is negligible for $|x^i_m - z^i| > 1.5$. Hence, positions x^i_m are generated independently and uniformly distributed over a disk of radius 1.5 centered in z^i. The gains $|A^0|^2$ and $|A^i_m|^2$ ($i > 0$) are still given by (27), but their statistics are different since they are conditioned on the occurrence of the events Ω^i_m.

With ideal power control and DDH, since x^0 and ξ^0 are statistically independent and x^0 is uniformly distributed over a disk of radius 1 centered in z^0, we can calculate the average transmitted energy per symbol as

$$\mathcal{E} = E\left[\frac{|x^0 - z^0|^\alpha}{\xi^0}\right] = \frac{2}{\alpha+2} \exp\left(\frac{1}{2}\left(\frac{\log 10}{10}\sigma_{\rm sh}\right)^2\right)\Bigg|_{\substack{\alpha=4 \\ \sigma_{\rm sh}=8}} \simeq 1.82 \quad (28)$$

With ideal power control and PDH, from Monte Carlo simulation we get $\mathcal{E} = 0.23$. We compared the different handoff and power control strategies for the same *transmitted* average SNR, given by $\mathcal{E}/(\rho N_0) = \mathcal{E}\gamma_b$ (in our examples, we considered $\mathcal{E}\gamma_b = 10$ dB). In passing, we note that power control with DDH increases the average transmitted power, while power control with PDH achieves a significant power saving.

Fading channel statistics. We make the simplifying assumption that all fading channels $c^i_m(\tau)$ have the same statistics, for all i and m. We consider the Rayleigh fading channel model given in [45] for a typical urban environment. The multipath intensity profile is given by

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

with $t_0 = 1$ μs and $T_d = 7$ μs. A sensible value for D in (23) for this channel is $D = 3 \times \lceil W_{\rm s}/(2 \cdot 10^5) \rceil$, i.e., 3 discretization subbands per 200 kHz bandwidth.

In our examples we consider interleaving depths $M = 1, 2, 4, 8$.[7] The most restrictive and most important assumption made in the following is that fading channel realizations in the $M \le 8$ bursts spanned by a code word are independent. This is motivated by the fact that, with $B_d = 50$ Hz and $\Delta T = 100$ ms, we have about $B_d \Delta T = 5$ degrees of freedom in time. Provided that the spacing between hopping carriers is sufficient, we can easily make up to 8 independent fading blocks. This assumption makes the mutual information $I_M(\mathbf{S})$ independent of the hopping sequences.

5.2 Outage probability results

Limiting outage probability. Fig. 1 shows the limit $\underline{P}_{\rm out}$ of $P_{\rm out}(\rho)$ as $\rho \to 0$ vs. $\mathcal{E}\gamma_b$, for $K = L = 1$, $W_{\rm s} = 200$ kHz and ideal power control, in the cases of DDH and PDH. These curves have been computed analytically via the residue method and Laplace inversion [46, 43].

[7] In the IS-54 standard $M = 2$, in the GSM half-rate standard $M = 4$ and in the GSM full-rate standard $M = 8$ [16].

Joint effect of handoff, power control, interferer diversity and voice activity. Figs. 2 and 3 show $P_{\rm out}(\rho)$ vs. ρ for a system with $K = L = u = 1$ and $W_{\rm s} = 200$ kHz, in the case of DDH with no power control and in the case of PDH with ideal power control, respectively. Each figure includes four curves, corresponding to the following combinations of interferer diversity and voice activity: Case a) no interferer diversity and $p = 1$; Case b) no interferer diversity and $p = 3/8$; Case c) maximum interferer diversity and $p = 1$; Case d) maximum interferer diversity and $p = 3/8$. For the sake of chart readability, we included only the results for $M = 8$ ($W_{\rm s} = 200$ kHz and $M = 8$ correspond to the GSM full-rate standard [16]). We notice that PDH and power control have a large impact on $P_{\rm out}(\rho)$ if used jointly, as expected.

The outage probability performance improves as the interleaving depth M increases. However, this improvement is not the same for all cases a), b), c) and d). For example, Fig. 4 shows $P_{\rm out}(\rho)$ vs. ρ for the same system parameters as before, in the cases a) and d) for $M = 1, 2, 4, 8$. We observe that interleaving depth has a large impact on the outage probability for $p = 3/8$ and maximum interferer diversity (case d), while this effect is very reduced for $p = 1$ and no interferer diversity (case a). This fact has been noticed in [31] for particular coded modulation schemes.

Effect of channel reuse, channel utilization and signal bandwidth expansion. Fig. 5 shows $P_{\rm out}(\rho)$ vs. ρ for $L = u = 1$, $W_{\rm s} = 200$ kHz and $K = 1, 3, 4, 7$. Fig. 6 shows $P_{\rm out}(\rho)$ vs. ρ for $K = L = 1$, $W_{\rm s} = 200$ kHz and u ranging from 0.1 to 1.0. Fig. 7 shows $P_{\rm out}(\rho)$ vs. ρ for $K = u = 1$, spreading factors $L = 1, 2, \ldots, 8$ and signal bandwidths $W_{\rm s} = 0.2 \times L$ MHz. Finally, Fig. 8 shows $P_{\rm out}(\rho)$ vs. ρ for $K = L = u = 1$ and $W_{\rm s}$ ranging from 200 kHz to 1.6 MHz. Again, for the sake of chart readability only the curves for $M = 8$ are shown. All sets of curves are obtained with maximum interferer diversity, voice activity $p = 3/8$, PDH and ideal power control.

By comparing these charts we note that the most effective countermeasure to reduce outage probability is increasing K (Fig. 5), followed by decreasing u (Fig. 6), by signal bandwidth expansion with spreading (Fig. 7) and by signal bandwidth expansion without spreading (Fig. 8). Similar results are obtained for shorter interleaving depths ($M = 1, 2$ and 4). On the other hand, $C_{\rm sys}$ is decreased by a factor $u/(KL)$, so that, for a given outage probability, simple signal bandwidth expansion with no waveform spreading (i.e., pure coding) may yield the largest $C_{\rm sys}$, as we will see next.

5.3 System capacity results

In this section we present results in terms of system capacity. All the results of this section are obtained with PDH and ideal power control. We fix the desired outage probability threshold as $P_{\rm out} = 10^{-2}$. Since $P_{\rm out}(\rho)$ is non-decreasing with ρ, we obtain the corresponding code rate by numerically solving for ρ the equation $P_{\rm out}(\rho) = P_{\rm out}$. This yields the maximum code rate that attains the required $P_{\rm out}$. Hence, the corresponding $C_{\rm sys}$ is obtained by multiplying this value by $u/(KL)$.

Figs. 9 and 10 show $C_{\rm sys}$ vs. $W_{\rm s}$ in the range from 200 kHz to 1.6 MHz, for $M = 1, 2, 4, 8$, $u = 1$, $K = 1$ and $K = 4$, respectively, in the case of maximum interferer diversity and $p = 3/8$. The solid lines correspond to the case $L = 1$, while the dashed lines correspond to the case of L proportional to $W_{\rm s}$. We note that the case $L = 1$ always outperforms the corresponding case with $L > 1$ in terms of $C_{\rm sys}$. Moreover, for L proportional to $W_{\rm s}$, $C_{\rm sys}$ decreases with $W_{\rm s}$ (except in some cases for $M = 1$) while, for $L = 1$, $C_{\rm sys}$ increases with $W_{\rm s}$. Note also that $C_{\rm sys}$

increases slowly for large $W_{\rm s}$. Similar results (not shown because of space limitations) were obtained for $K = 3$ and $K = 7$. The largest value of $C_{\rm sys}$ for $M = 1$ is achieved by $K = 4$, for $M = 2$ by $K = 3$ and for $M = 4, 8$ by $K = 1$. This shows that channel reuse $K = 1$ provides the largest system capacity provided that a sufficient interleaving is possible ($M \geq 4$, in our case). For very correlated fading situations and/or very strict decoding delay constraints (i.e., very short interleaving depth), reuse $K > 1$ might be a better choice.

Fig. 11 shows $C_{\rm sys}$ vs. u, in the case $K = L = 1$, $W_{\rm s} = 1.6$ MHz, $M = 1, 2, 4, 8$, maximum interferer diversity, with and without voice activity exploitation. In both cases, $u = 1$ maximizes $C_{\rm sys}$ for $M \geq 4$, while for shorter interleaving depth some $u < 1$ is optimal. Without voice activity exploitation ($p = 1$), the largest value of $C_{\rm sys}$ for $M = 1$ is achieved by $u \simeq 0.2$ and for $M = 2$ by $u \simeq 0.45$. With voice activity exploitation ($p = 3/8$), the largest value of $C_{\rm sys}$ for $M = 1$ is still achieved by $u \simeq 0.2$ and for $M = 2$ by $u \simeq 0.7$.

In order to give an idea about how many users per cell can be served by a well-designed *conventional* system approaching $C_{\rm sys} \simeq 1.5$, with $M = 8$, $L = u = 1$ and $W_{\rm s} = 200$ kHz, assuming a system bandwidth $W = 12.8$ MHz, a user bit-rate $R_b = 9.6$ kbit/s and a F-TDMA overhead of 30% (i.e., a factor $\eta = 0.7$ in (6)) [8], we get $N_{\rm u} \simeq 1400$ users/cell. This system has 64 hopping carriers. With delay constraint $\Delta T = 100$ ms we get frame duration $T_{\rm f} = 12.5$ ms, 22 time slots per frame of duration $T_{\rm s} = 568.2\mu$s, $N_{\rm s} = \eta W_{\rm s} T_{\rm s} \simeq 80$ coded symbols and 34 overhead (training plus guard) symbols, yielding a total of 114 symbols per burst and a code word length $N_{\rm s}M = 640$. Actually, voice activity with $p = 3/8$ is rather optimistic [6, 3]. In the same conditions as above, with $p = 1$ we get $C_{\rm sys} = 0.5$ which yields $N_{\rm u} = 460$. Then, with more realistic voice activity statistics we may expect $400 \leq N_{\rm u} \leq 1400$ users/cell.

6 Conclusions

In this paper we considered a F-TDMA cellular system with conventional single-user processing. This assumption reduces both the inherently different problems of uplink (multiple-access plus interference) and downlink (broadcast plus interference) to an equivalent single-user additive noise channel. For this system we provided a definition of system capacity in terms of users/cell×bit/s/Hz, for a desired maximum outage probability. Outage probability has been defined as the probability that the mutual information of the M-burst channel resulting from F-TDMA transmission in a mobile environment with frequency selective slow fading falls below the actual user code rate. Expressions for the mutual information under the assumption that all users transmit with Gaussian input distribution and flat power spectral density have been derived. Since Gaussian noise minimizes mutual information over all types of additive noise with the same second-order statistics, Gaussian inputs yield an upperbound to the minimum achievable outage probability. Also, we provided an operational characterization of outage probability as the achievable WER averaged over the user random coding ensemble and over all the possible realizations of the channel state, as the burst length goes to infinity. Results with practical burst lengths and low-complexity codes show that outage probability is actually a good performance indicator for the average WER of real systems.

A numerical example covering the interesting case of a typical urban mobile environment and regular hexagonal coverage was developed in detail. Driven by this example, we can make

[8]In the GSM standard, a training sequence of 26 symbols, a guard time of 8.25 symbols and 3 header and tail symbols are inserted in each burst carrying 114 encoded symbols [16], corresponding to an overhead of about 26 %.

some remarks on capacity-maximizing F-TDMA system design:

i) Coding over several bursts, power control, path-driven handoff, interferer diversity and voice activity exploitation must be jointly implemented in order to increase $C_{\rm sys}$ and achieve channel reuse and utilization 1. If the interleaving depth is small and/or if interferer diversity and voice activity are not exploited, $C_{\rm sys}$ is maximized by some optimal K and u. Waveform spreading $L > 1$ seems always to decrease $C_{\rm sys}$. This is certainly due to the single-user processing scheme considered in this paper. In fact, as shown in [22], a simple multiuser detection scheme consisting of a MMSE interference canceler followed by a bank of single-user decoders may yield better $C_{\rm sys}$ for $L > 1$. It is likely to expect even a larger impact of user waveform design if more sophisticated joint processing schemes are employed in the base-stations (as for example joint processing for users in the same cluster of neighboring cells as in [1, 2]).

ii) The behavior of $C_{\rm sys}$ is rather flat around the optimal values of u. This insensitivity of system capacity with respect to channel utilization implies that the fraction u of channels in use can be fixed by some practical protocol driven by reasons different from pure system capacity maximization (e.g., reservation of some free channels for accommodating handoff requests with high probability) with negligible system capacity decrease, provided that u is not too far from its optimal value.

iii) Because of frequency selectivity, $C_{\rm sys}$ is increased by expanding the signal bandwidth with $L = 1$. This fact suggests that good strategies for maximizing $C_{\rm sys}$ are either coded TDMA with some powerful channel estimation and equalization technique, or coded OFDM with a sufficient number D of subcarriers (in our case, for $W_{\rm s} = 1.6$ MHz, we need $D \geq 24$). From the channel estimation and equalization viewpoint, the latter might lead to an easier implementation, especially if the number of carriers is not very large.

iv) The largest system capacity value found in our simulations is $C_{\rm sys} \simeq 1.7$, achieved by $K = L = u = 1$, $W_{\rm s} = 1.6$ MHz and $\rho \simeq 1.7$ bit/symbol (for the fading statistics considered here, larger signal bandwidths yield only negligible capacity increases). Hence, even if we assumed the channel input alphabet $\mathcal{X}$ to be infinite and Gaussian distributed, the best code rate can be approached by standard QAM coded modulation (e.g., 16QAM with a powerful concatenated coding scheme with rate slightly less than 1/2). There is no need for high code rates and complicated signal alphabets. The use of coding/signaling schemes matched to the block-fading frequency-selective channel emerges as a key point for maximizing system capacity (results on code construction for block-fading channels can be found in [27, 28, 47]).

v) We showed that a carefully designed conventional cellular system can achieve a fairly large system capacity. In order to achieve further improvements, future-generation cellular systems should exploit new non-conventional techniques. Among them, we may indicate transmission/reception diversity [48, 49, 50] and multiuser joint processing.

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ΔT	100 ms
B_d	50 Hz
α	4
σ_{sh}	8 dB
k_t	3
N	37
K	1, 3, 4, 7
M	1, 2, 4, 8
p	3/8, 1
u	from 0 to 1
W_{s}	from 0.2 to 1.6 MHz
L	from 1 to 8

Table 1: Parameter values used in the numerical example of Section 5.

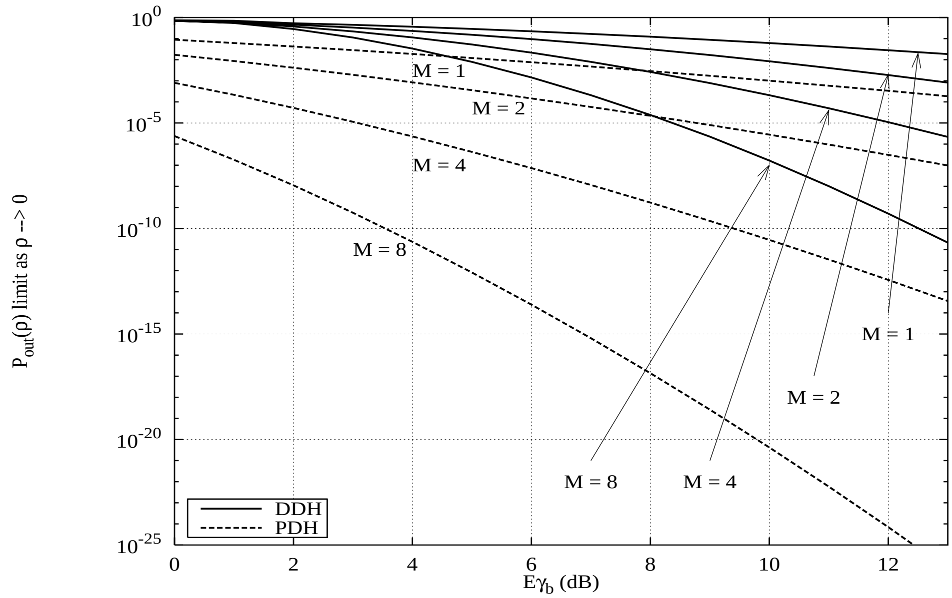

Figure 1: $\underline{P}_{\rm out}$ vs. $\mathcal{E}\gamma_b$ for $K = L = 1$, $W_{\rm s}$ =200 kHz, ideal power control with DDH and PDH.

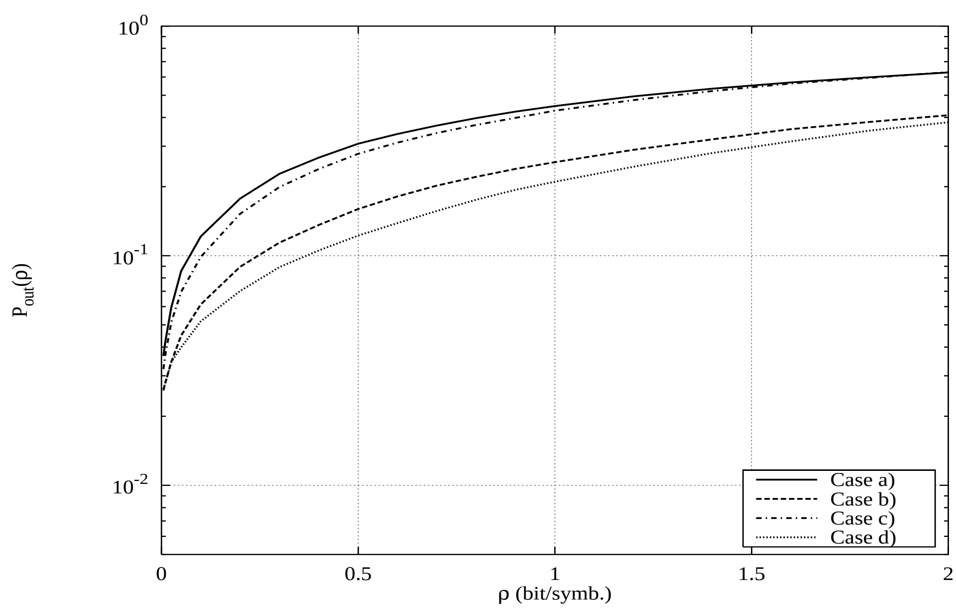

Figure 2: $P_{\rm out}(\rho)$ vs. ρ for $K = L = u = 1$, $M = 8$, $W_{\rm s}$ =200 kHz, DDH and no power control. Case a) no interferer diversity, voice activity $p = 1$; Case b) no interferer diversity, voice activity $p = 3/8$; Case c) max. interferer diversity, voice activity $p = 1$; Case d) max. interferer diversity, voice activity $p = 3/8$.

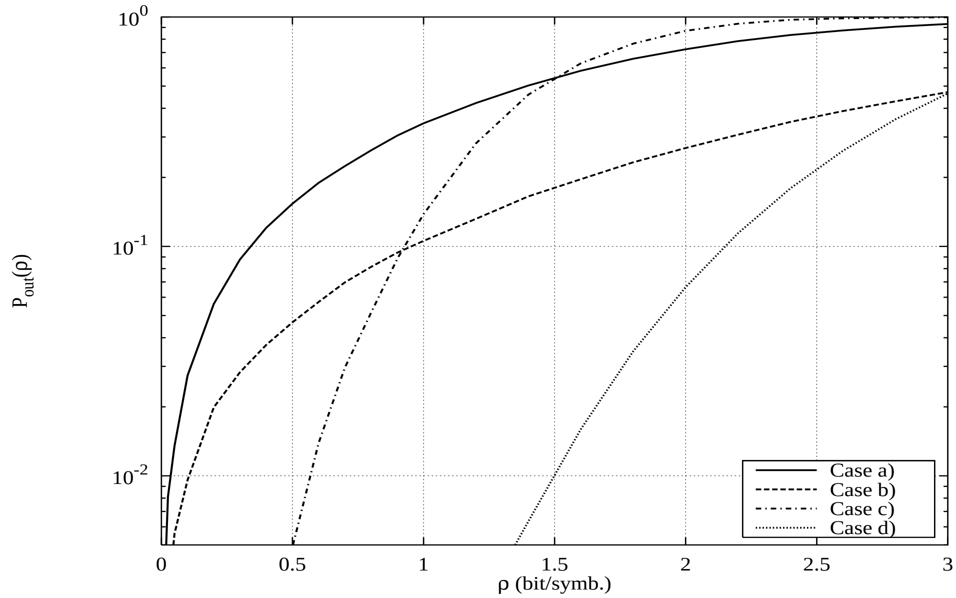

Figure 3: $P_{\rm out}(\rho)$ vs. ρ for $K = L = u = 1$, $M = 8$, $W_{\rm s}$ =200 kHz, PDH and ideal power control. Case a) no interferer diversity, voice activity $p = 1$; Case b) no interferer diversity, voice activity $p = 3/8$; Case c) max. interferer diversity, voice activity $p = 1$; Case d) max. interferer diversity, voice activity $p = 3/8$.

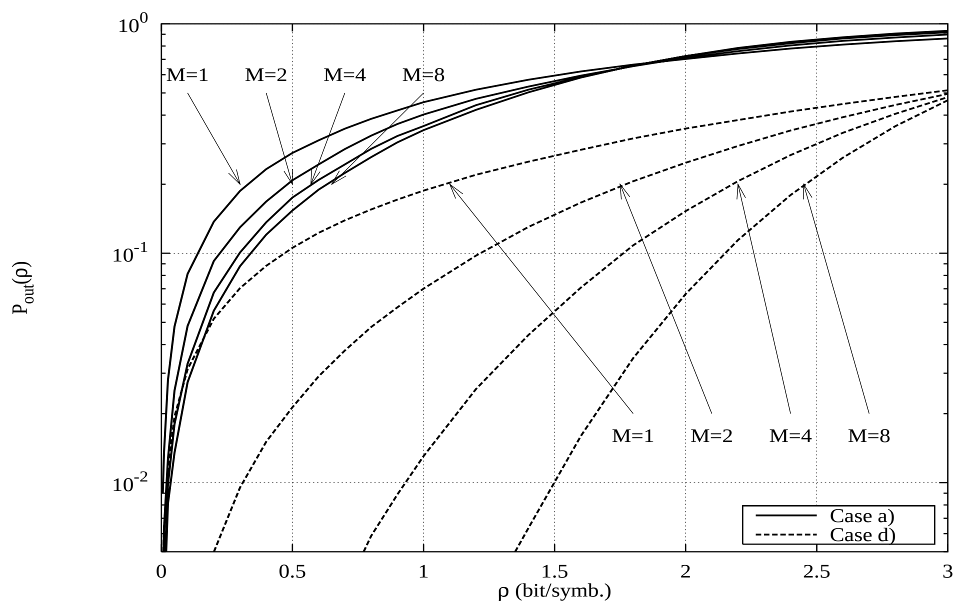

Figure 4: $P_{\rm out}(\rho)$ vs. ρ for $K = L = u = 1$, $M = 1, 2, 4, 8$, $W_{\rm s}$ =200 kHz, PDH and ideal power control. Case a) no interferer diversity, voice activity $p = 1$; Case d) max. interferer diversity, voice activity $p = 3/8$.

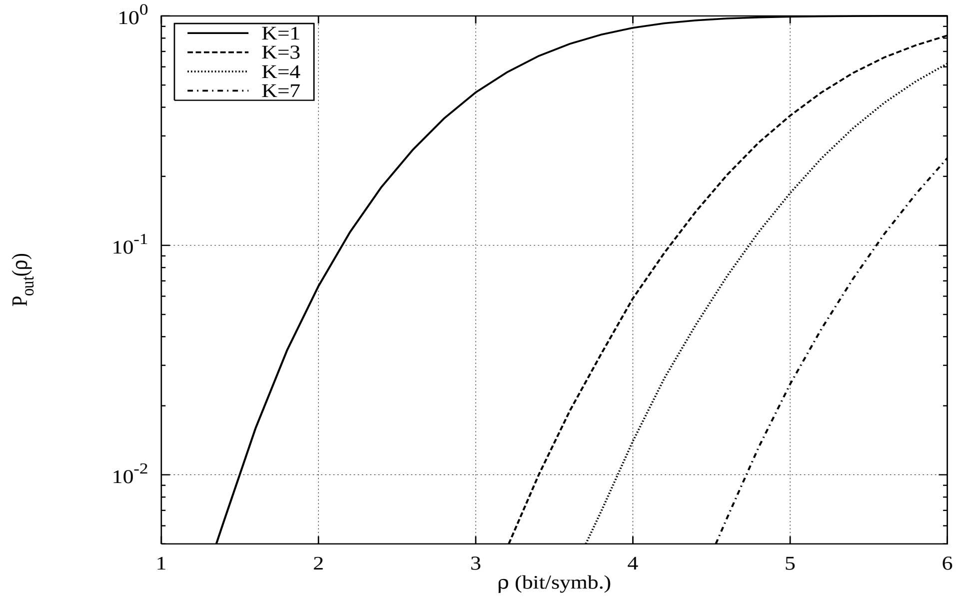

Figure 5: $P_{\text{out}}(\rho)$ vs. ρ for $L = u = 1$, $M = 8$, $W_s = 200$ kHz, PDH, ideal power control, maximum interferer diversity, voice activity $p = 3/8$ and for reuse cluster sizes $K = 1, 3, 4, 7$.

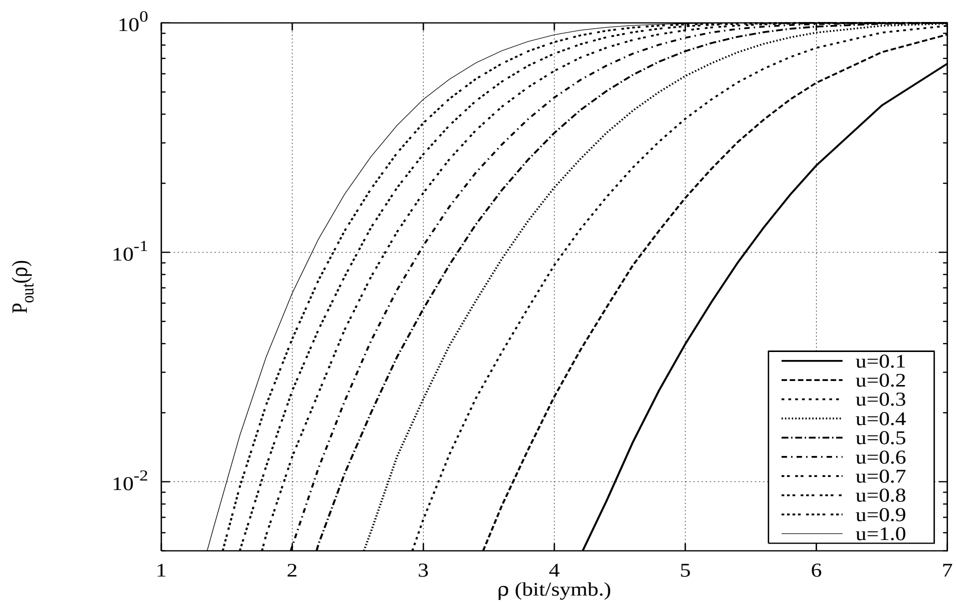

Figure 6: $P_{\text{out}}(\rho)$ vs. ρ for $K = L = 1$, $M = 8$, $W_s = 200$ kHz, PDH, ideal power control, maximum interferer diversity, voice activity $p = 3/8$ and channel utilization u ranging from 0.1 to 1.

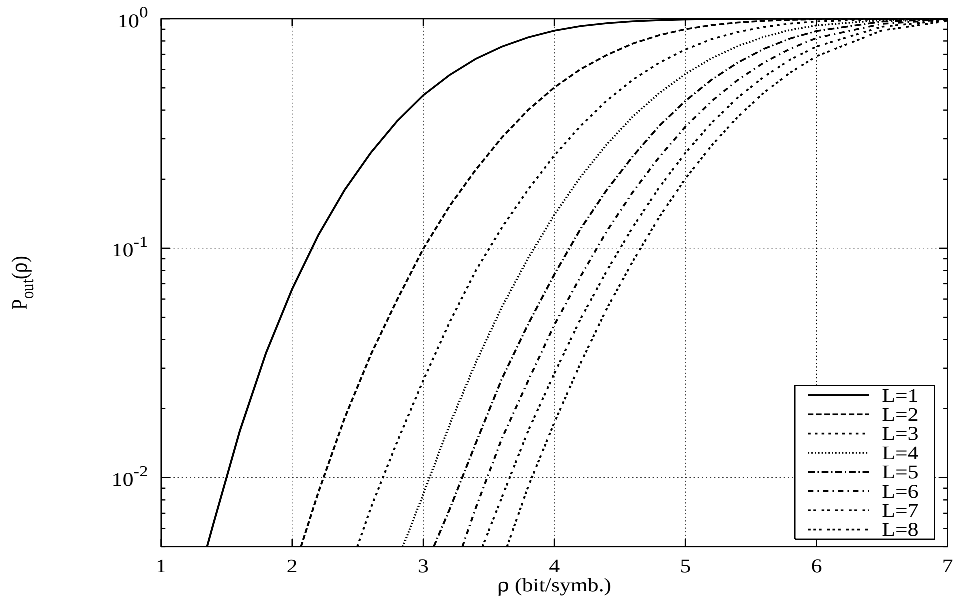

Figure 7: $P_{\text{out}}(\rho)$ vs. ρ for $K = u = 1$, $M = 8$, PDH, ideal power control, maximum interferer diversity, voice activity $p = 3/8$ and for spreading factors $L = 1, 2, 3, 4, 5, 6, 7, 8$ with $W_{\text{s}} = 0.2 \times L$ MHz.

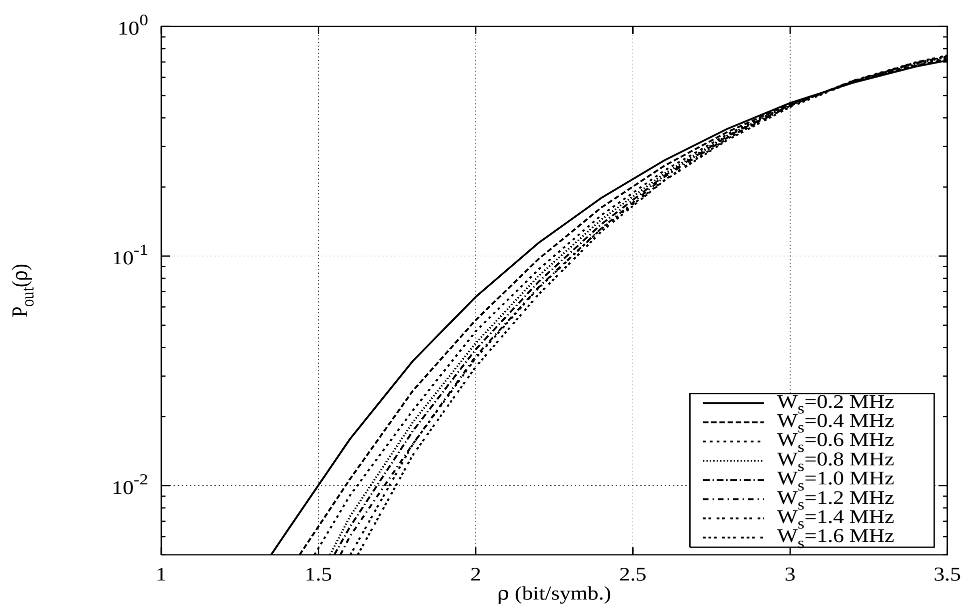

Figure 8: $P_{\text{out}}(\rho)$ vs. ρ for $K = L = u = 1$, $M = 8$, PDH, ideal power control, maximum interferer diversity, voice activity $p = 3/8$ and for signal bandwidths from 200 kHz to 1.6 MHz with steps of 200 kHz.

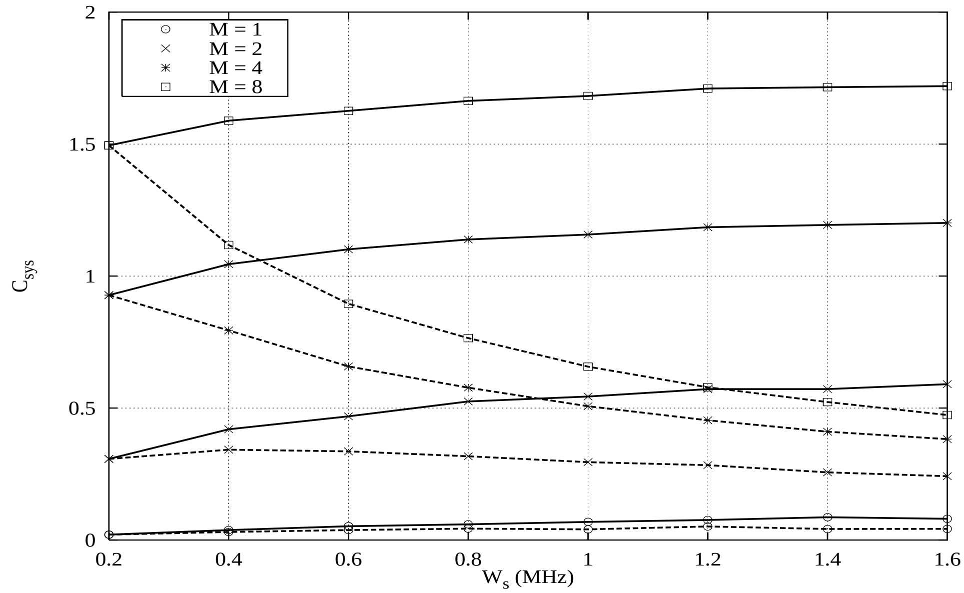

Figure 9: $C_{\rm sys}$ vs. $W_{\rm s}$ for $K = 1$, $u = 1$ PDH, ideal power control, maximum interferer diversity and voice activity $p = 3/8$. Solid lines are for the case of no spreading ($L = 1$) while dashed lines are for signals with spreading (L proportional to $W_{\rm s}$).

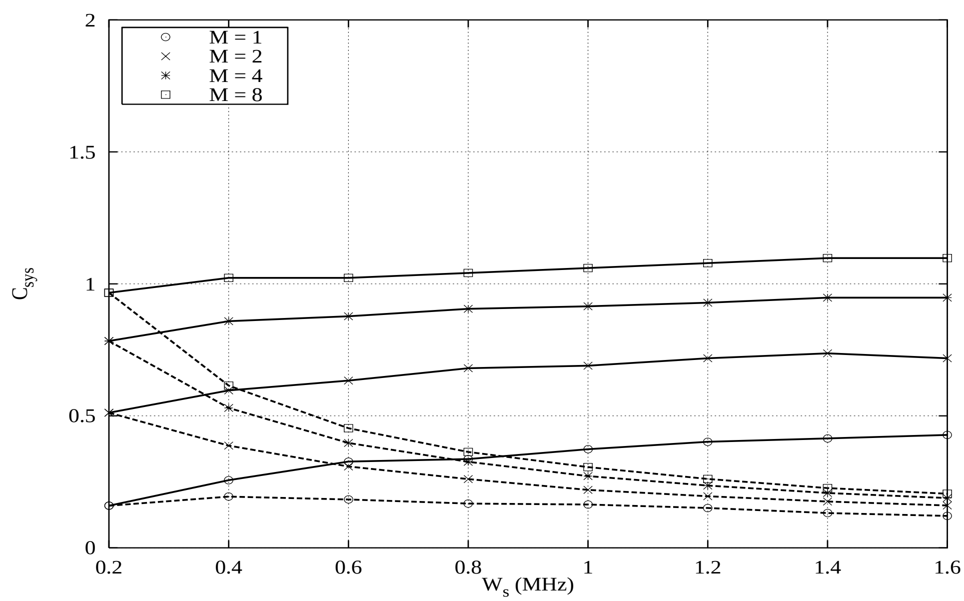

Figure 10: $C_{\rm sys}$ vs. $W_{\rm s}$ for $K = 4$, $u = 1$ PDH, ideal power control, maximum interferer diversity and voice activity $p = 3/8$. Solid lines are for the case of no spreading ($L = 1$) while dashed lines are for signals with spreading (L proportional to $W_{\rm s}$).

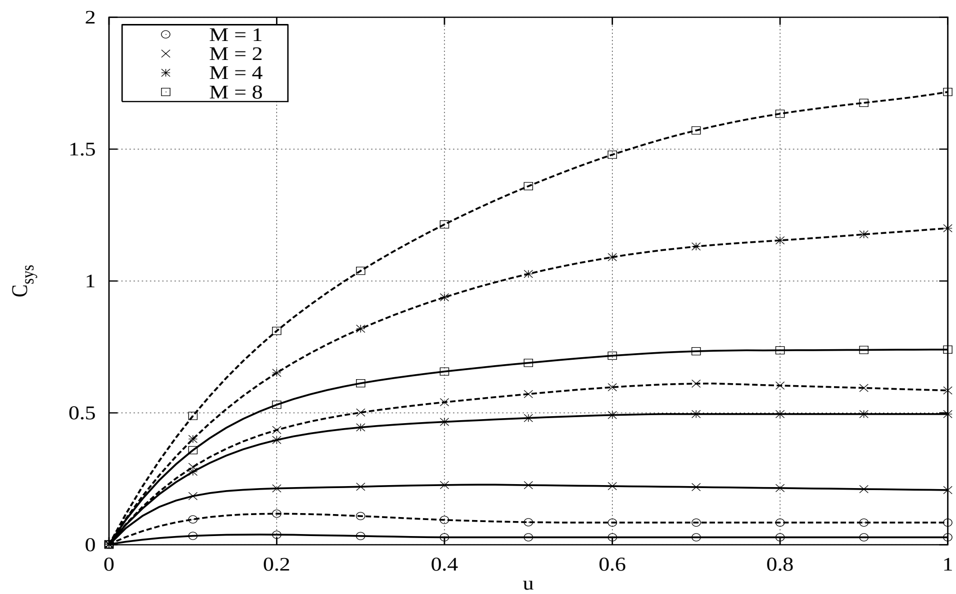

Figure 11: C_{sys} vs. u for $K = L = 1$, $W_{\text{s}} = 1.6$ MHz, PDH, ideal power control, maximum interferer diversity. Solid lines: no voice activity ($p = 1$). Dashed lines: voice activity ($p = 3/8$).

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