

# Spectral Efficiency of CDMA Downlink Cellular Networks with Matched Filter

Nicolas Bonneau,<sup>1</sup> Mérouane Debbah,<sup>2</sup> and Eitan Altman<sup>1</sup>

<sup>1</sup>MAESTRO, INRIA Sophia Antipolis, 2004 Route des Lucioles, B.P. 93, 06902 Sophia Antipolis, France

<sup>2</sup>Mobile Communications Group, Institut Eurécom, 2229 Route des Crêtes, B.P. 193, 06904 Sophia Antipolis, France

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In this contribution, the performance of a downlink code division multiple access (CDMA) system with orthogonal spreading and multicell interference is analyzed. A useful framework is provided in order to determine the optimal base station coverage for wireless frequency selective channels with dense networks where each user is equipped with a matched filter. Using asymptotic arguments, explicit expressions of the spectral efficiency are obtained and provide a simple expression of the network spectral efficiency based only on a few meaningful parameters. Contrarily to a common misconception which asserts that to increase spectral efficiency in a CDMA network, one has to increase the number of cells, we show that, depending on the path loss and the fading channel statistics, the code orthogonal gain (due to the synchronization of all the users at the base station) can compensate and even compete in some cases with the drawbacks due to intercell interference. The results are especially realistic and useful for the design of dense networks.

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## 1. INTRODUCTION

An important problem that arises in the design of CDMA systems concerns the deployment of an efficient architecture to cover the users. Increasing the number of cells in a given area yields indeed a better coverage but increases at the same time intercell interference. The gain provided by a cellular network is not at all straightforward and depends on many parameters: path loss, type of codes used, receiving filter, and channel characteristics. Previous studies have already studied the spectral efficiency of an uplink CDMA multicell network with Wyner's model [1] or with simple interference models [2–9]. However, none has taken explicitly into account the impact of the code structure (orthogonality) and the multipath channel characteristics.

This contribution is a first step into analyzing the complex problem of downlink CDMA multicell networks, using a new approach based on unitary random matrix theory. The purpose of this contribution is to determine, for a dense and infinite multicell network, the optimal distance between base stations. A downlink frequency selective fading CDMA scheme where each user is equipped with a linear matched filter is considered. The users are assumed to be uniformly

distributed along the area. Only orthogonal access codes are considered, as the users are synchronized within each cell. The problem is analyzed in the asymptotic regime: very dense networks are considered where the spreading length  $N$  tends to infinity, the number of users per meter  $d$  tends to infinity, but the load per meter  $d/N = \alpha$  is constant. The analysis is mainly based on asymptotic results of unitary random matrices [10]. One of the great features of this tool is that performance measures such as signal-to-interference-plus-noise ratio (SINR) [11] or spectral efficiency [12] have very simple forms in the large system limit, independent of the particular CDMA code structure. Moreover, the theoretical results were shown to be very accurate predictions of the system's behavior in the finite size case (spreading length  $N$  of 256 for the SINR [13] or number of antennas 8 for the mutual information [14]).

This paper is structured as follows: in Section 2, the cellular CDMA model is introduced. In Section 3, the SINR expression is derived and an asymptotic analysis of the spectral efficiency with matched filter in case of downlink orthogonal CDMA is provided. Finally in Section 4, discussions as well as numerical simulations are provided in order to validate our analysis.

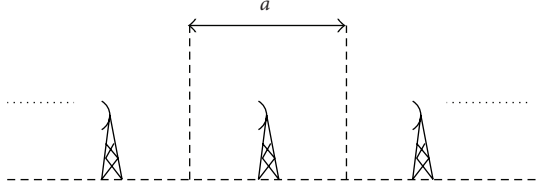


FIGURE 1: Representation of a CDMA cellular network.

## 2. CDMA CELLULAR MODEL

### 2.1. Cellular model

Without loss of generality and in order to ease the understanding, we focus our analysis on a one-dimensional (1D) network. This scenario represents, for example, the case of the deployment of base stations along a motorway (users, i.e., cars are supposed to move along the motorway). Discussions on the two-dimensional (2D) case are given in Section 3.3. An infinite length base station deployment is considered (see Figure 1). The base stations are supposed equidistant with interbase station distance  $a$ . The spreading length  $N$  is fixed and is independent of the number of users. The number of users per cell is  $K = da$  ( $d$  is the density of the network). Note that as the size of the cell increases, each cell accommodates more users (with the constraint  $da \leq N$ ). However, the total power emitted by the network does not increase since the same number of users is served by the network.

### 2.2. Downlink CDMA model

In the following, upper case and lower case boldface symbols will be used for matrices and column vectors, respectively.  $(\cdot)^T$  will denote the transpose operator,  $(\cdot)^*$  conjugation, and  $(\cdot)^H = ((\cdot)^T)^*$  hermitian transpose.  $\mathbb{E}$  denotes the expectation operator. The general case of downlink wide-band CDMA is considered where the signal transmitted by the base station in cell  $p$  to user  $j$  has complex envelope

$$x_{pj}(t) = \sum_n s_{pj}(n) v_{pj}(t - nT). \quad (1)$$

In (1),  $v_{pj}(t)$  is a weighted sum of elementary modulation pulses  $\psi(t)$  which satisfy the Nyquist criterion with respect to the chip interval  $T_c$  ( $T = NT_c$ ):

$$v_{pj}(t) = \sum_{\ell=1}^N v_{p\ell j} \psi(t - (\ell - 1)T_c). \quad (2)$$

The signal is transmitted over a frequency selective channel with impulse response  $c_{pj}(\tau)$ . Under the assumption of slowly varying fading, the continuous time signal  $y_j^p(t)$  received by user  $j$  in cell  $p$  has the form

$$y_j^p(t) = \sum_q \sum_n \sum_{k=1}^K s_{qk}(n) \int \sqrt{P_q(x_j)} c_{qk}(\tau) v_{qk} \times (t - nT - \tau) d\tau + n(t), \quad (3)$$

where  $n(t)$  is the complex white Gaussian noise. In (3), the index  $q$  stands for the cells, the index  $n$  for the transmitted symbol, and the index  $k$  for the users (in each cell there are  $K$  users). User  $j$  is determined by his position  $x_j$ . The signal (after pulse-matched filtering by  $\psi^*(-t)$ ) is sampled at the chip rate to get a discrete-time received signal of user  $j$  in cell  $p$  of a downlink CDMA system that has the form

$$\mathbf{y}^p(x_j) = \sum_q \sqrt{P_q(x_j)} \mathbf{C}_{qj} \mathbf{W}_q \mathbf{s}_q + \mathbf{n}. \quad (4)$$

$x_j$  are the coordinates of user  $j$  in cell  $p$ .  $\mathbf{y}^p(x_j)$  is the  $N \times 1$  received vector,  $\mathbf{s}_q = [s_q(1), \dots, s_q(K)]^T$  is the  $K \times 1$  transmit vector of cell  $q$ , and  $\mathbf{n} = [n(1), \dots, n(N)]^T$  is an  $N \times 1$  noise vector with zero mean and variance  $\sigma^2$  Gaussian independent entries.  $P_q(x_j)$  represents the path loss between base station  $q$  and user  $j$  whereas matrix  $\mathbf{C}_{qj}$  represents the  $N \times N$  Toeplitz structured frequency selective channel between base station  $q$  and user  $j$ . Each base station has an  $N \times K$  code matrix  $\mathbf{W}_q = [\mathbf{w}_q^1, \dots, \mathbf{w}_q^K]$ . User  $j$  is subject to intra-cell interference from other users of cell  $p$  as well as intercell interference from all the other cells.

### 2.3. Assumptions

The following assumptions are rather technical in order to simplify the analysis.

#### 2.3.1. Code structure model

In the downlink scenario, Walsh-Hadamard codes are usually used. However, in order to get interpretable expressions of the SINR, isometric matrices obtained by extracting  $K < N$  columns from a Haar unitary matrix will be considered. An  $N \times N$  random unitary matrix is said to be Haar distributed if its probability distribution is invariant by right (or equivalently left) multiplication by deterministic unitary matrices. In spite of the limited practical use, these random matrices represent a very useful analytical tool as simulations [13] and show that their use provides similar performances as Walsh-Hadamard codes. *Note that each cell uses a different isometric code matrix.*

#### 2.3.2. Multipath channel

We consider the case of a multipath channel. Under the assumption that the number of paths from base station  $q$  to any user  $j$  is given by  $L_{qj}$ , the model of the channel is given by

$$c_{qj}(\tau) = \sum_{\ell=0}^{L_{qj}-1} \eta_{qj}(\ell) \psi(\tau - \tau_{qj}(\ell)), \quad (5)$$

where we assume that the channel is invariant during the time considered. In order to compare channels at the same signal-to-noise ratio, we constrain the distribution of the i.i.d. fading coefficients  $\eta_{qj}(\ell)$  such as

$$\mathbb{E}[\eta_{qj}(\ell)] = 0, \quad \mathbb{E}[|\eta_{qj}(\ell)|^2] = \frac{\rho}{L_{qj}}. \quad (6)$$

Usually, fading coefficients  $\eta_{qj}(\ell)$  are supposed to be independent with decreasing variance as the delay increases. In all cases,  $\varrho$  is the average power of the channel, such as  $\mathbb{E}[|c(\tau)|^2] = \sum_{\ell=0}^{L_{qj}-1} \mathbb{E}[|\eta_{qj}(\ell)|^2] = \varrho$ . For each base station  $q$ , let  $h_{qj}(i)$  be the discrete Fourier transform of the fading process  $c_{qj}(\tau)$ . The frequency response of the channel at the receiver is given by

$$h_{qj}(f) = \sum_{\ell=0}^{L_{qj}-1} \eta_{qj}(\ell) e^{-j2\pi f \tau_{qj}(\ell)} |\Psi(f)|^2, \quad (7)$$

where we assume that the transmit filter  $\Psi(f)$  and the receive filter  $\Psi^*(-f)$  are such that, given the bandwidth  $W$ ,

$$\Psi(f) = \begin{cases} 1 & \text{if } -\frac{W}{2} \leq f \leq \frac{W}{2}, \\ 0 & \text{otherwise.} \end{cases} \quad (8)$$

Sampling at the various frequencies  $f_1 = -W/2$ ,  $f_2 = -W/2 + 1/NW$ ,  $\dots$ ,  $f_N = -W/2 + ((N-1)/N)W$ , we obtain the coefficients  $h_{qj}(i)$ ,  $1 \leq i \leq N$ , as

$$h_{qj}(i) = h_{qj}(f_i) = \sum_{\ell=0}^{L_{qj}-1} \eta_{qj}(\ell) e^{-j2\pi(i/N)W\tau_{qj}(\ell)} e^{j\pi W\tau_{qj}(\ell)}. \quad (9)$$

Note that  $\mathbb{E}[|h_{qj}(i)|^2] = \varrho$ .

Notice that the assumption on the code structure model enables us to simplify the model (4) in the following way. Let  $\mathbf{C}_{qj} = \mathbf{U}_{qj} \mathbf{H}_{qj} \mathbf{V}_{qj}$  denote the singular value decomposition of  $\mathbf{C}_{qj}$ .  $\mathbf{U}_{qj}$  and  $\mathbf{V}_{qj}$  are unitary, while  $\mathbf{H}_{qj}$  is a diagonal matrix with elements  $\{\tilde{h}_{qj}(1), \dots, \tilde{h}_{qj}(N)\}$ .

Since  $\mathbf{U}_{qj}$  and  $\mathbf{V}_{qj}$  are unitary,  $\mathbf{U}_{qj}^H \mathbf{n}$  and  $\mathbf{V}_{qj} \mathbf{W}_q$  have respectively the same distribution as  $\mathbf{n}$  and  $\mathbf{W}_q$  (the distribution of a Haar distributed matrix is left unchanged by left multiplication by a constant unitary matrix). As a consequence, model (4) is equivalent to

$$\mathbf{y}^p(x_j) = \sum_q \sqrt{P_q(x_j)} \mathbf{H}_{qj} \mathbf{W}_q \mathbf{s}_q + \mathbf{n}, \quad (10)$$

where  $\mathbf{H}_{qj}$  is a diagonal matrix with diagonal elements  $\{\tilde{h}_{qj}(i)\}_{i=1 \dots N}$ . Note that for any user  $j$  in cell  $p$ , when  $N \rightarrow \infty$ , the coefficients  $\tilde{h}_{qj}(i)$  tend to the discrete Fourier transform of the channel impulse response  $h_{qj}(i)$  given by (9) (see [15]).

### 2.3.3. Path loss

The general path loss  $P_q(x_j)$  depends on a path loss factor which characterizes the type of attenuation. The greater the factor is, the more severe the attenuation is. In Section 3, we will derive expressions for an exponential path loss  $P_q(x_j) = P e^{-\gamma|x_j - m_q|}$  [16], where  $m_q$  are the coordinates of base station  $q$ . Note that in the usual model, the attenuation is of the polynomial form:  $P_q(x_j) = P/(|x_j - m_q|)^\beta$ . The polynomial path loss will be considered through simulations in Section 4 to validate our results. We use the exponential form for the sake

of calculation simplicity and therefore put the framework in the most severe path loss scenario in favor of the multicell approach.

## 3. PERFORMANCE ANALYSIS

### 3.1. General SINR formula

In all the following, without loss of generality, we will focus on user  $j$  of cell  $p$ . We assume that the user does not know the codes of the other cells as well as the codes of the other users within the same cell. As a consequence, the user is equipped with the matched filter receiver  $\mathbf{g}_{pj} = \mathbf{H}_{pj} \mathbf{w}_p^j$ .

The output of the matched filter is given by

$$\begin{aligned} \mathbf{g}_{pj}^H \mathbf{y}^p(x_j) &= \sqrt{P_p(x_j)} \mathbf{g}_{pj}^H \mathbf{H}_{pj} \mathbf{w}_p^j s_p(j) \\ &+ \sqrt{P_p(x_j)} \mathbf{g}_{pj}^H \mathbf{H}_{pj} \mathbf{W}_p^{(-j)} \begin{bmatrix} s_p(1) \\ \vdots \\ s_p(K) \end{bmatrix}_{(K-1) \times 1} \\ &+ \sum_{q \neq p} \sqrt{P_q(x_j)} \mathbf{g}_{pj}^H \mathbf{H}_{qj} \mathbf{W}_q \mathbf{s}_q + \mathbf{g}_{pj}^H \mathbf{n}, \end{aligned} \quad (11)$$

where  $\mathbf{W}_p^{(-j)} = [\mathbf{w}_p^1, \dots, \mathbf{w}_p^{j-1}, \mathbf{w}_p^{j+1}, \dots, \mathbf{w}_p^K]$ . From (11), we obtain the expression for the output SINR of user  $j$  in cell  $p$  with coordinates  $x_j$  and code  $\mathbf{w}_p^j$ :

$$\text{SINR}(x_j, \mathbf{w}_p^j) = \frac{S^*(x_j)}{I_1(x_j) + I_2(x_j) + \sigma^2 \mathbf{w}_p^j{}^H \mathbf{H}_{pj}^H \mathbf{H}_{pj} \mathbf{w}_p^j}, \quad (12)$$

where

$$S^*(x_j) = P_p(x_j) |\mathbf{w}_p^j{}^H \mathbf{H}_{pj}^H \mathbf{H}_{pj} \mathbf{w}_p^j|^2, \quad (13)$$

$$I_1(x_j) = \sum_{q \neq p} P_q(x_j) \mathbf{w}_p^j{}^H \mathbf{H}_{pj}^H \mathbf{H}_{qj} \mathbf{W}_q \mathbf{W}_q^H \mathbf{H}_{qj}^H \mathbf{H}_{pj} \mathbf{w}_p^j, \quad (14)$$

$$I_2(x_j) = P_p(x_j) \mathbf{w}_p^j{}^H \mathbf{H}_{pj}^H \mathbf{H}_{pj} \mathbf{W}_p^{(-j)} \mathbf{W}_p^{(-j)H} \mathbf{H}_{pj}^H \mathbf{H}_{pj} \mathbf{w}_p^j. \quad (15)$$

Note that the SINR is a random variable with respect to the channel model. For a fixed  $d$  (or  $K = da$ ) and  $N$ , it is extremely difficult to get some insight on expression (12). In order to provide a tractable expression, we will analyze (12) in the asymptotic regime ( $N \rightarrow \infty$ ,  $d \rightarrow \infty$ , but  $d/N \rightarrow \alpha$ ) and show in particular that  $\text{SINR}(x_j, \mathbf{w}_p^j)$  converges almost surely to a random value  $\text{SINR}_{\text{lim}}(x_j, p)$  independent of the code  $\mathbf{w}_p^j$ . Usual analysis based on random matrices use the ratio  $K/N$  [17], also known as the load of the system. In our case, the ratio  $K/N$  is equal to  $\alpha a$ .

**Proposition 1.** *When  $N$  grows towards infinity and  $d/N \rightarrow \alpha$ , the SINR of user  $j$  in cell  $p$  in downlink CDMA with orthogonal*

spreading codes and matched filter is given by

$$\begin{aligned} \text{SINR}_{\text{lim}}(x_j, p) &= \frac{P_p(x_j) \left( (1/W) \int_{-W/2}^{W/2} |h_{pj}(f)|^2 df \right)^2}{I_1(x_j) + I_2(x_j) + (\sigma^2/W) \int_{-W/2}^{W/2} |h_{pj}(f)|^2 df}, \\ I_1(x_j) &= \frac{\alpha a}{W} \sum_{q \neq p} P_q(x_j) \int_{-W/2}^{W/2} |h_{pj}(f)|^2 |h_{qj}(f)|^2 df, \\ I_2(x_j) &= \frac{\alpha a}{W} P_p(x_j) \left( \int_{-W/2}^{W/2} |h_{pj}(f)|^4 df \right. \\ &\quad \left. - \frac{1}{W} \left( \int_{-W/2}^{W/2} |h_{pj}(f)|^2 df \right)^2 \right). \end{aligned} \quad (16)$$

*Proof.* See Appendix B.  $\square$

### 3.2. Spectral efficiency

We would like to quantify the number of bps/Hz the system is able to provide to all the users. It has been shown [18] that the interference plus noise at the output of the matched filter for randomly spread systems can be considered as Gaussian when  $K$  and  $N$  are large enough. In this case, the mean (with respect to the position of the users and the fading) spectral efficiency of cell  $p$  is given by

$$C^p = \frac{1}{N} \mathbb{E}_{x,h} \left[ \sum_{j=1}^K \log_2 (1 + \text{SINR}(x_j, \mathbf{w}_p^j)) \right]. \quad (17)$$

In the asymptotic case and due to invariance by translation, the spectral efficiency per cell is the same for all cells. As a consequence, the network spectral efficiency is infinite. Without loss of generality, we will consider a user in cell 0 ( $x_j \in [-a/2, a/2]$ ) and the corresponding asymptotic SINR is denoted by  $\text{SINR}_{\text{lim}}(x_j)$ . Assuming the same distribution for all the users in cell 0, we drop the index  $j$ . The measure of performance in this case is the number of bits per second per

hertz per meter (bps/Hz/meter) the system is able to deliver

$$\begin{aligned} C &= \frac{1}{a} \frac{K}{N} \mathbb{E}_{x,h} [\log_2 (1 + \text{SINR}_{\text{lim}}(x))] \\ &= \alpha \mathbb{E}_{x,h} [\log_2 (1 + \text{SINR}_{\text{lim}}(x))]. \end{aligned} \quad (18)$$

According to the size of the network, the total spectral efficiency scales linearly with the factor  $C$ . If we suppose that in each cell, the statistics of the channels are the same, then denoting  $P_0(x) = P(x)$ ,  $h_0(f) = h(f)$ , and  $L_0 = L$ , we obtain the following proposition from (16).

**Proposition 2.** *When the spreading length  $N$  grows towards infinity and  $d/N \rightarrow \alpha$ , the asymptotic spectral efficiency per meter of downlink CDMA with random orthogonal spreading codes, general path loss, and matched filter is given by*

$$\begin{aligned} C(a) &= \frac{\alpha}{a} \mathbb{E}_h \left[ \int_{-a/2}^{a/2} \log_2 \left( 1 + \frac{P(x) \left( (1/W) \int_{-W/2}^{W/2} |h(f)|^2 df \right)^2}{I(x) + (\sigma^2/W) \int_{-W/2}^{W/2} |h(f)|^2 df} \right) dx \right], \\ I(x) &= \frac{\alpha a}{W} \sum_{q \neq 0} P_q(x) \int_{-W/2}^{W/2} |h(f)|^2 |h_q(f)|^2 df \\ &\quad + \frac{\alpha a}{W} P(x) \left( \int_{-W/2}^{W/2} |h(f)|^4 df - \frac{1}{W} \left( \int_{-W/2}^{W/2} |h(f)|^2 df \right)^2 \right) \end{aligned} \quad (19)$$

with  $a \in [0, 1/\alpha]$ .

### 3.3. 2D network

In the case of a 2D network, the expression for the general SINR (16) from Proposition 1 is still valid if we admit that  $x_j = (x_j^1, x_j^2)$  represents the coordinates of the user considered,  $d$  is the density of users *per square meter*, and  $a = |\mathcal{C}|$  is the *surface* of the cell  $\mathcal{C}$ . The expression (19) for the spectral efficiency from Proposition 2 can be immediately rewritten with a double integration over the surface of the cell:

$$\begin{aligned} C(a) &= \frac{\alpha}{a} \mathbb{E}_h \left[ \iint_{\mathcal{C}} \log_2 \left( 1 + \frac{P(x^1, x^2) \left( (1/W) \int_{-W/2}^{W/2} |h(f)|^2 df \right)^2}{I(x^1, x^2) + (\sigma^2/W) \int_{-W/2}^{W/2} |h(f)|^2 df} \right) dx^1 dx^2 \right], \\ I(x^1, x^2) &= \frac{\alpha a}{W} \sum_{q \neq 0} P_q(x^1, x^2) \int_{-W/2}^{W/2} |h(f)|^2 |h_q(f)|^2 df \\ &\quad + \frac{\alpha a}{W} P(x^1, x^2) \left( \int_{-W/2}^{W/2} |h(f)|^4 df - \frac{1}{W} \left( \int_{-W/2}^{W/2} |h(f)|^2 df \right)^2 \right) \quad \text{with } a \in [0, 1/\alpha]. \end{aligned} \quad (20)$$

### 3.4. Simplifying assumptions

#### 3.4.1. Equi-spaced delays

For ease of understanding of the impact of the number of paths on the orthogonality gain, we suppose that in each cell  $q$ , all the users have the same number of paths  $L_q$  and the delays are uniformly distributed according to the bandwidth:

$$\tau_q(\ell) = \frac{\ell}{W}. \quad (21)$$

Hence, replacing  $h(f)$  and  $h_q(f)$  with their expression with respect to the temporal coefficients (7) and using (21) and (19) from Proposition 2 reduces to

$$\begin{aligned} C(a) &= \frac{\alpha}{ar} \mathbb{E}_\eta \left[ \int_{-a/2}^{a/2} \log_2 \left( 1 + \frac{P(x) \left( \sum_{\ell=0}^{L-1} |\eta(\ell)|^2 \right)^2}{I(x) + \sigma^2 \sum_{\ell=0}^{L-1} |\eta(\ell)|^2} \right) dx \right], \\ I(x) &= \alpha a \sum_{q \neq 0} P_q(x) \left( \left( \sum_{\ell=0}^{L-1} |\eta(\ell)|^2 \right) \left( \sum_{\ell=0}^{L_q-1} |\eta_q(\ell)|^2 \right) \right. \\ &\quad \left. + \sum_{\ell=0}^{\min(L, L_q)-1} \sum_{\ell' \neq \ell} \eta(\ell) \eta(\ell')^* \eta_q(\ell)^* \eta_q(\ell') \right) \\ &\quad + \alpha a P(x) \left( \sum_{\ell=0}^{L-1} \sum_{\ell' \neq \ell} |\eta(\ell)|^2 |\eta(\ell')|^2 \right). \end{aligned} \quad (22)$$

In the case of a single path (i.e.,  $L_q = 1$  for all  $q$ ), the signal is only affected by flat fading, therefore orthogonality is preserved and intracell interference vanishes:

$$C(a) = \frac{\alpha}{ar} \mathbb{E}_\eta \left[ \int_{-a/2}^{a/2} \log_2 \left( 1 + \frac{P(x) |\eta|^2}{\alpha a \sum_{q \neq 0} P_q(x) |\eta_q|^2 + \sigma^2} \right) dx \right]. \quad (23)$$

#### 3.4.2. Exponential path loss and ergodic case

In the case of an exponential path loss, explicit expressions of the spectral efficiency can be derived when  $L_q \rightarrow \infty$  for all  $q$ , referred in the following as the ergodic case. Although  $L_q$  grows large, we suppose  $L_q$  to be negligible with respect to  $N$  (see the appendix). The impact of frequency reuse is also considered. In other words,  $r$  adjacent cells may use different frequencies to reduce the amount of interference. This point is a critical issue to determine the impact of frequency reuse on the spectral efficiency of downlink CDMA networks.

**Proposition 3.** *When the spreading length  $N$  and the number of paths  $L_q$  (for all  $q$ ) grow towards infinity with  $d/N \rightarrow \alpha$  and  $L_q/N \rightarrow 0$ , the asymptotic spectral efficiency per meter of downlink CDMA with random orthogonal spreading codes, exponential path loss, frequency reuse  $r$ , and matched filter is*

given by

$$\begin{aligned} C(a) &= \frac{\alpha}{ar} \int_{-a/2}^{a/2} \log_2 \left( 1 + \frac{P e^{-\gamma|x|} (\mathbb{E}[|h|^2])^2}{I(x) + \sigma^2 \mathbb{E}[|h|^2]} \right) dx, \\ I(x) &= \alpha a P (\mathbb{E}[|h|^2])^2 \frac{2e^{-\gamma ar}}{1 - e^{-\gamma ar}} \cosh(\gamma x) \\ &\quad + \alpha a P e^{-\gamma|x|} (\mathbb{E}[|h|^4] - (\mathbb{E}[|h|^2])^2) \end{aligned} \quad (24)$$

with  $a \in [0, 1/\alpha]$ .

*Proof.* See Appendix C.  $\square$

In the case where  $a \rightarrow 0$ , the number of bps/Hz/meter depends only on the fading statistics, the path loss, and the factor  $\alpha = d/N$  through

$$C(0) = \alpha \log_2 \left( 1 + \frac{P \mathbb{E}[|h|^2]}{\sigma^2 + 2\alpha P \mathbb{E}[|h|^2]/\gamma} \right). \quad (25)$$

For the proof, let  $a \rightarrow 0$  in (24).

## 4. DISCUSSION

In all the following discussion,  $P = 1$ ,  $\sigma^2 = 10^{-7}$ ,  $\alpha = 10^{-2}$ , and  $r = 1$  (unless specified otherwise).

### 4.1. Path loss versus orthogonality

We would like to quantify the impact of path loss on the overall performance of the system when considering downlink unfaded CDMA. In this case,

$$\begin{aligned} C(a) &= \frac{\alpha}{a} \int_{-a/2}^{a/2} \log_2 \left( 1 + \frac{P(x)}{I(x) + \sigma^2} \right) dx, \\ I(x) &= \alpha a \sum_{q \neq 0} P_q(x) \end{aligned} \quad (26)$$

with  $a \in [0, 1/\alpha]$ .

In Figures 2 and 3, we have plotted the spectral efficiency per meter with respect to the intercell distance for an exponential ( $\gamma = 1, 2, 3$ ) and polynomial ( $\beta = 4$ ) path loss, without frequency selective fading. Remarkably, for each path loss factor, there is an optimum intercell distance which maximizes the users' spectral efficiency. This surprising result shows that there is no need into packing base stations without bound if one can remove completely the effect of frequency selective fading. It can be shown that optimal spacing depends mainly on the path loss factor  $\gamma$  and increases with a decreasing path loss factor.

### 4.2. Ergodic fading versus orthogonality

We would like to quantify the impact of the channel statistics on the intercell distance. In other words, in the case of limited path loss, should one increase or reduce the cell size? A neat framework can be formulated in the case of exponential path loss with vanishing values of the path loss factor  $\gamma$  and ergodic fading. Although the spectral efficiency tends to zero,

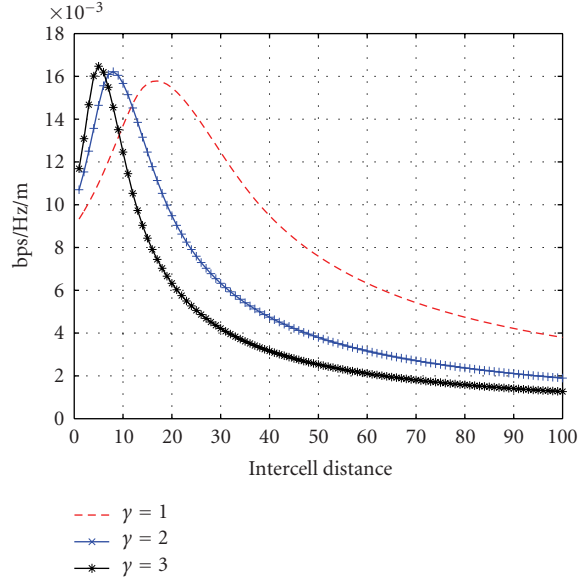


FIGURE 2: Spectral efficiency versus intercell distance (in meters) in the case of exponential path loss and no fading:  $\sigma^2 = 10^{-7}$ ,  $P = 1$ , and  $\gamma = 1, 2, 3$ .

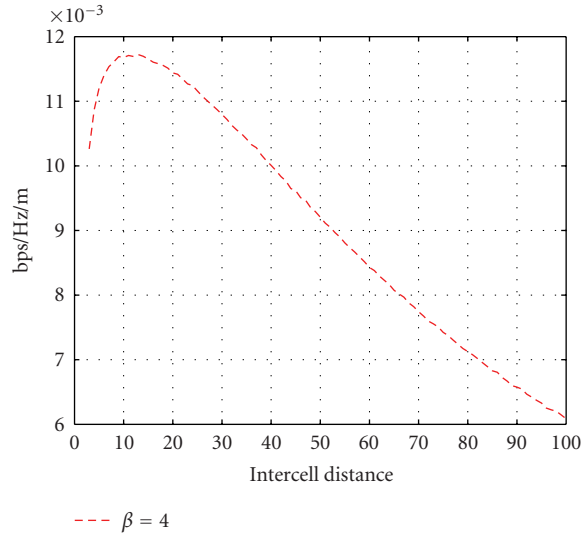


FIGURE 3: Spectral efficiency versus intercell distance (in meters) in the case of polynomial path loss ( $\beta = 4$ ) and no fading:  $\sigma^2 = 10^{-7}$ ,  $P = 1$ .

one can infer on the behavior of the derivative of the spectral efficiency which is given by

$$\frac{\partial C}{\partial a} \propto \left( \frac{3}{2} - \frac{\mathbb{E}[|h|^4]}{(\mathbb{E}[|h|^2])^2} \right) \quad (27)$$

with  $a \in [0, 1/\alpha]$ .

For the proof, see Appendix D.

This simplified case (exponential path loss with ergodic fading) is quite instructive on the impact of frequency selective fading on orthogonal downlink CDMA. In the ergodic

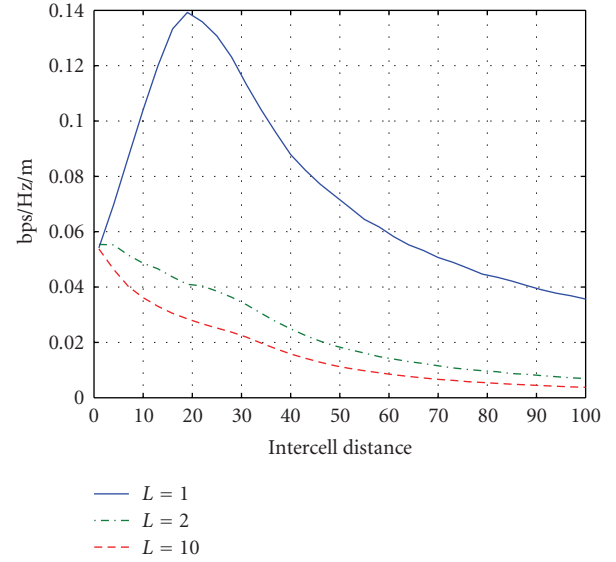


FIGURE 4: Spectral efficiency versus intercell distance (in meters) in the case of exponential path loss and multipath fading:  $\sigma^2 = 10^{-7}$ ,  $P = 1$ ,  $\gamma = 1$ , and  $L = 1, 2, 10$ .

case and with limited path loss, the optimum number of cells depends only on how “peaky” the channel is through the kurtosis  $T = \mathbb{E}[|h|^4]/(\mathbb{E}[|h|^2])^2$ . If  $T > 3/2$ , orthogonality is severely destroyed by the channel and one has to decrease the cell size whereas if  $T \leq 3/2$ , one can increase the cell size.<sup>1</sup>

### 4.3. Number of paths versus orthogonality

We would like to quantify the impact of the number of multipaths on the overall performance of the system. In Figures 4 and 5, we have plotted the spectral efficiency per meter with respect to the intercell distance for an exponential ( $\gamma = 1$ ) and polynomial ( $\beta = 4$ ) path loss, in each case for numbers of multipaths  $L = 1$ ,  $L = 2$ , and  $L = 10$  (supposing an equal number of paths is generated by each cell) and Rayleigh fading. For  $L = 1$ , fading does not destroy orthogonality and as a consequence, an optimum intercell distance is obtained as in the nonfading case. However, for any value of  $L > 1$ , the optimum intercell distance is equal to 0.

### 4.4. Impact of reuse factor

In Figure 6, we consider a realistic case with ergodic Rayleigh frequency selective fading and  $\gamma = 2$  and reuse factor  $r = 1, 2, 3$ . The spectral efficiency has been plotted for various values of the intercell distance. The curve shows that the users’ rate decreases with increasing intercell distance, which is mainly due to frequency selective fading. Note that the best

<sup>1</sup> The value 3/2 is mainly dependent on the type of path loss (exponential, polynomial, ...).

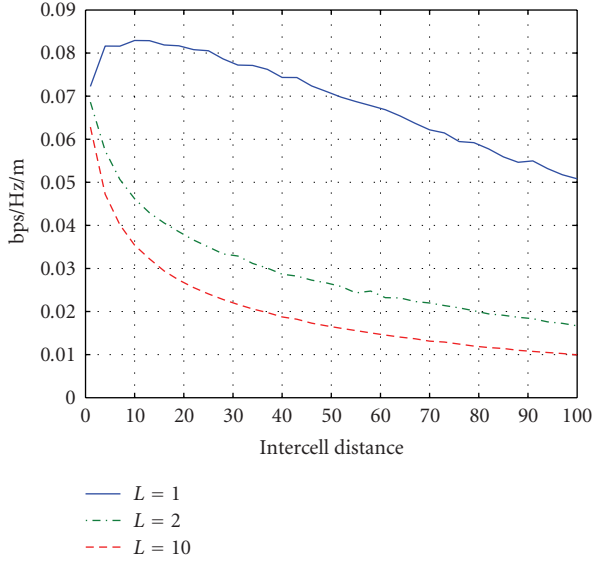


FIGURE 5: Spectral efficiency versus intercell distance (in meters) in the case of polynomial path loss ( $\beta = 4$ ) and multipath fading:  $\sigma^2 = 10^{-7}$ ,  $P = 1$ , and  $L = 1, 2, 10$ .

spectral efficiency is achieved for a reuse factor of 1, meaning that all base stations should use all the available bandwidth.

**4.5. General discussion**

We would like to show that, in a cellular system, multipath fading is in fact more dramatic than path loss and restoring orthogonality through diversity (multiple antennas at the base station) and equalization techniques (MMSE, ...) pays off. To visually confirm this fact, Figure 7 plots for a path loss factor  $\gamma = 2$  the spectral efficiency per meter with respect to the intercell distance in the ergodic Rayleigh frequency selective fading, nonfading, and intercell interference-free case (i.e.,  $(1/a) \int_{-a/2}^{a/2} \log_2(1 + P(x)/\sigma^2) dx$ ). The figure shows that one can more than triple the spectral efficiency per meter by restoring orthogonal multiple access for any intercell distance. Moreover, for small values of the intercell distance, greater gains can be achieved if one removes intercell interference (by exploiting the statistics of the intercell interference, e.g.). Note also that even with fading and intercell interference, the capacity gain with respect to the number of base stations is not linear and therefore, based on economic constraints, the optimal interbase station distance can be determined. Hence, based on the quality of service targets for each user, the optimum intercell distance can be straightforwardly derived.

**5. CONCLUSION**

Using asymptotic arguments, an explicit expression of the spectral efficiency was derived and was shown to depend only on a few meaningful parameters. This contribution is also very instructive in terms of future research directions. In the "traditional point of view" of cellular systems, the general

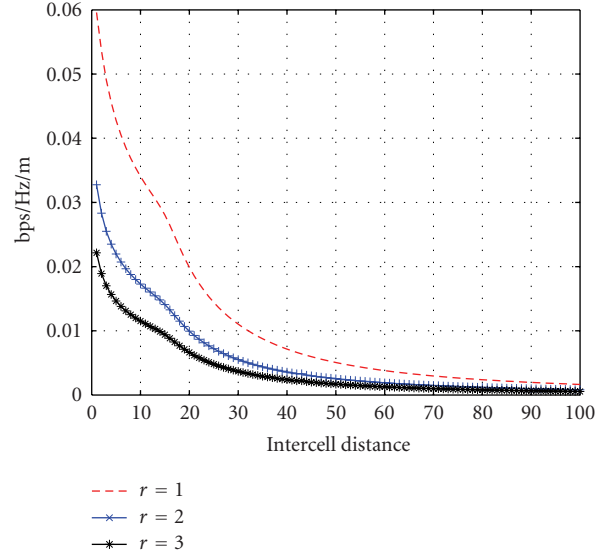


FIGURE 6: Effect of the reuse factor: spectral efficiency versus intercell distance (in meters) in the case of exponential path loss and fading:  $\sigma^2 = 10^{-7}$ ,  $\gamma = 2$ ,  $P = 1$ , and  $r = 1, 2, 3$ .

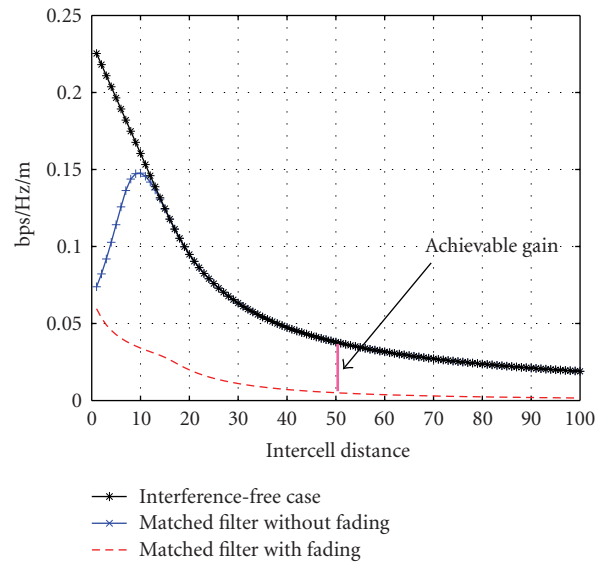


FIGURE 7: Spectral efficiency versus intercell distance (in meters) in the case of exponential path loss with fading, without fading, and interference-free case (one cell):  $\sigma^2 = 10^{-7}$ ,  $P = 1$ , and  $\gamma = 2$ .

guidance to increase the cell size has always been related to an increase in the transmitted power to reduce path loss. However, these results show that path loss is only the second part of the story and the first obstacle is on the contrary frequency selective fading since path loss does not destroy orthogonal multiple access whereas frequency selective fading does. These considerations show therefore that all the effort must be focused on combating frequency selective fading through diversity and equalization techniques in order to

restore orthogonality. Finally, note that the results presented in this paper deal only with the downlink and any deployment strategy should take also into account the uplink traffic as in [19].

## APPENDIX

### A. PRELIMINARY RESULTS

Let  $\mathbf{X}_p$  be an  $N \times N$  random matrix with i.i.d. zero-mean unit variance Gaussian entries. The matrix  $\mathbf{X}_p(\mathbf{X}_p^H \mathbf{X}_p)^{-1/2}$  is unitary Haar distributed (see [13] for more details). The code matrix  $\mathbf{W}_p = [\mathbf{w}_p^1, \dots, \mathbf{w}_p^K]$  is obtained by extracting  $K$  orthogonal columns from  $\mathbf{X}_p(\mathbf{X}_p^H \mathbf{X}_p)^{-1/2}$ . In this case, the entries of matrix  $\mathbf{W}_p$  verify [10] that

$$\mathbb{E}[|w_p^1(i)|^2] = \frac{1}{N}, \quad 1 \leq i \leq N, \quad (\text{A.1})$$

$$\mathbb{E}[|w_p^1(i)|^2 |w_p^k(i)|^2] = \frac{1}{N(N+1)}, \quad k > 1, \quad (\text{A.2})$$

$$\mathbb{E}[w_p^1(i)^* w_p^k(i) w_p^1(l) w_p^k(l)^*] = -\frac{1}{N(N^2-1)}, \quad k > 1, i \neq l. \quad (\text{A.3})$$

### B. PROOF OF PROPOSITION 1

*Term  $S^*$*

Let us focus on the term  $S^*$  of (13). As  $N \rightarrow \infty$ , (13) becomes

$$S^* = \mathbf{w}_p^{jH} \mathbf{H}_{pj}^H \mathbf{H}_{pj} \mathbf{w}_p^j = \sum_{i=1}^N |h_{pj}(i)|^2 |w_p^j(i)|^2. \quad (\text{B.1})$$

Using (A.1), it is rather straightforward to show that

$$S^* \rightarrow \lim_{N \rightarrow \infty} \frac{1}{N} \sum_{i=1}^N |h_{pj}(i)|^2 \quad (\text{B.2})$$

in the mean-square sense. Therefore,

$$S^* \xrightarrow{N \rightarrow \infty} \frac{1}{W} \int_{-W/2}^{W/2} |h_{pj}(f)|^2 df. \quad (\text{B.3})$$

Formula (B.3) stems from the fact that as  $N \rightarrow \infty$ , the eigenvalues  $|h_{pj}(i)|^2$  of  $\mathbf{H}_{pj}^H \mathbf{H}_{pj}$  correspond to the squared frequency response of the channel in the case of a Toeplitz structure of  $\mathbf{H}_{pj}$  (see [15]).

*Term  $I_1$*

Let us now derive the term  $I_1$  of (14). It can be shown that (since  $\mathbf{w}_p^j$  is independent of  $\mathbf{W}_q$ , see proof in [13])

$$\begin{aligned} & \mathbf{w}_p^{jH} \mathbf{H}_{pj}^H \mathbf{H}_{qj} \mathbf{W}_q \mathbf{W}_q^H \mathbf{H}_{qj}^H \mathbf{H}_{pj} \mathbf{w}_p^j \\ & - \frac{1}{N} \text{trace}(\mathbf{W}_q \mathbf{W}_q^H \mathbf{H}_{qj}^H \mathbf{H}_{pj} \mathbf{H}_{pj}^H \mathbf{H}_{qj}) \rightarrow 0. \end{aligned} \quad (\text{B.4})$$

Therefore, each term  $I_q^*$  in the sum in (14) can be calculated as

$$\begin{aligned} I_q^* &= \mathbf{w}_p^{jH} \mathbf{H}_{pj}^H \mathbf{H}_{qj} \mathbf{W}_q \mathbf{W}_q^H \mathbf{H}_{qj}^H \mathbf{H}_{pj} \mathbf{w}_p^j \\ &\rightarrow \frac{1}{N} \text{trace}(\mathbf{W}_q \mathbf{W}_q^H \mathbf{H}_{qj}^H \mathbf{H}_{pj} \mathbf{H}_{pj}^H \mathbf{H}_{qj}) \\ &\rightarrow \frac{1}{N} \sum_{k=1}^K \sum_{i=1}^N |h_{pj}(i)|^2 |h_{qj}(i)|^2 |w_q^k(i)|^2. \end{aligned} \quad (\text{B.5})$$

Using (A.1), it is rather straightforward to show that

$$I_q^* \rightarrow \lim_{N \rightarrow \infty} \frac{1}{N^2} \sum_{k=1}^K \sum_{i=1}^N |h_{pj}(i)|^2 |h_{qj}(i)|^2 \quad (\text{B.6})$$

in the mean-square sense. Therefore,

$$I_q^* \xrightarrow[N \rightarrow \infty]{d/N \rightarrow \alpha} \frac{\alpha a}{W} \int_{-W/2}^{W/2} |h_{pj}(f)|^2 |h_{qj}(f)|^2 df. \quad (\text{B.7})$$

*Term  $I_2$*

Finally, let us derive the asymptotic expression of  $I_2$  in (15). The proof follows here a different procedure as  $\mathbf{w}_p^j$  is not independent of  $\mathbf{W}_p^{(-j)}$ . However, one can show that

$$\begin{aligned} I_2 &= P_p(x_j) \mathbf{w}_p^{jH} \mathbf{H}_{pj}^H \mathbf{H}_{pj} \mathbf{W}_p^{(-j)} \mathbf{W}_p^{(-j)H} \mathbf{H}_{pj}^H \mathbf{H}_{pj} \mathbf{w}_p^j \\ &= P_p(x_j) \sum_{\substack{k=1 \\ k \neq j}}^K (\mathbf{w}_p^{jH} \mathbf{H}_{pj}^H \mathbf{H}_{pj} \mathbf{w}_p^k)^2 \\ &= P_p(x_j) \sum_{\substack{k=1 \\ k \neq j}}^K \left( \sum_{i=1}^N |h_{pj}(i)|^2 w_p^j(i)^* w_p^k(i) \right)^2 \\ &= P_p(x_j) \sum_{\substack{k=1 \\ k \neq j}}^K \sum_{i=1}^N \sum_{l=1}^N |h_{pj}(i)|^2 |h_{pj}(l)|^2 w_p^j(i)^* \\ &\quad \times w_p^j(l) w_p^k(i) w_p^k(l)^*, \end{aligned} \quad (\text{B.8})$$

and using (A.2) and (A.3), it can be shown that  $I_2$  converges in the mean-square sense to

$$\begin{aligned} & P_p(x_j) \lim_{N \rightarrow \infty} \frac{1}{N(N+1)} \sum_{\substack{k=1 \\ k \neq j}}^K \sum_{i=1}^N |h_{pj}(i)|^4 \\ & - \frac{1}{N(N^2-1)} \sum_{\substack{k=1 \\ k \neq j}}^K \sum_{i=1}^N \sum_{\substack{l=1 \\ l \neq i}}^N |h_{pj}(i)|^2 |h_{pj}(l)|^2. \end{aligned} \quad (\text{B.9})$$

Therefore,

$$I_2 \xrightarrow[\frac{d/N-\alpha}{N-\alpha}]{} P_p(x_j) \frac{\alpha\alpha}{W} \left( \int_{-W/2}^{W/2} |h_{pj}(f)|^4 - \frac{1}{W} \left( \int_{-W/2}^{W/2} |h_{pj}(f)|^2 df \right)^2 \right). \quad (\text{B.10})$$

### C. PROOF OF PROPOSITION 3

Note that as  $L_q \rightarrow \infty$ ,

$$\sum_{\ell=0}^{L_q-1} |\eta_q(\ell)|^2 \rightarrow \mathbb{E} \left[ \sum_{\ell=0}^{L_q-1} |\eta_q(\ell)|^2 \right] = \mathbb{E}[|h|^2], \quad (\text{C.1})$$

$$\begin{aligned} & \sum_{\ell=0}^{L-1} \sum_{\ell' \neq \ell} |\eta(\ell)|^2 |\eta(\ell')|^2 \\ &= \left( \sum_{\ell=0}^{L-1} |\eta(\ell)|^2 \right)^2 + \sum_{\ell=0}^{L-1} \sum_{\ell' \neq \ell} |\eta(\ell)|^2 |\eta(\ell')|^2 \\ & - \left( \sum_{\ell=0}^{L-1} |\eta(\ell)|^2 \right)^2 \rightarrow \mathbb{E}[|h|^4] - (\mathbb{E}[|h|^2])^2, \end{aligned} \quad (\text{C.2})$$

and  $\sum_{\ell' \neq \ell} \eta(\ell)\eta(\ell')^* \eta_q(\ell)^* \eta_q(\ell') \rightarrow 0$ .

The rest of the proof is mainly an application of Proposition 1 where we consider a path loss of the form  $P e^{-\gamma(|x-qa|)}$  ( $\gamma$  is a decaying factor) between the user  $x$  ( $x \in [-a/2, a/2]$ ) and base station  $q$  ( $q \in \mathbb{Z}$ ) with coordinates  $m_q = qa$ . In this case, the intercell interference has an explicit

form:

$$\begin{aligned} \sum_{q \neq 0} P_q(x) &= P \sum_{q=-\infty}^{+\infty} e^{-\gamma|x-qa|} \\ &= P e^{-\gamma x} \sum_{q=1}^{+\infty} e^{-\gamma qar} + P e^{\gamma x} \sum_{q=1}^{+\infty} e^{-\gamma qar} \\ &= \frac{2P e^{-\gamma ar}}{1 - e^{-\gamma ar}} \cosh(\gamma x). \end{aligned} \quad (\text{C.3})$$

### D. PROOF OF (27)

Let  $\gamma \rightarrow 0$  in (24). In this case,

$$\begin{aligned} & \lim_{\gamma \rightarrow 0} C(a) \\ &= \frac{\alpha}{a} \int_{-a/2}^{a/2} \log_2 \left( 1 + \frac{P(1 - \gamma|x| - (\gamma^2/2)|x|^2)(\mathbb{E}[|h|^2])^2}{I + \sigma^2 \mathbb{E}[|h|^2]} \right) dx \\ & + O(\gamma^3), \end{aligned} \quad (\text{D.1})$$

where

$$\begin{aligned} I &= \alpha a P (\mathbb{E}[|h|^2])^2 \frac{2e^{-\gamma a}}{1 - e^{-\gamma a}} + \alpha a P (\mathbb{E}[|h|^4] - (\mathbb{E}[|h|^2])^2) \\ &= \alpha a P (\mathbb{E}[|h|^4] - 2(\mathbb{E}[|h|^2])^2) + \frac{2\alpha P}{\gamma} (\mathbb{E}[|h|^2])^2 \end{aligned} \quad (\text{D.2})$$

since  $\lim_{\gamma \rightarrow 0} 2/e^{\gamma a} - 1 = 2/\gamma a - 1$ . We have therefore

$$\begin{aligned} \lim_{\gamma \rightarrow 0} C(a) &= \frac{(\alpha/\ln 2) P (\mathbb{E}[|h|^2])^2 (1 - \gamma a/4 - \gamma^2 a^2/24)}{(2\alpha P/\gamma) (\mathbb{E}[|h|^2])^2 (1 + \gamma((a/2)(\mathbb{E}[|h|^4]/(\mathbb{E}[|h|^2])^2 - 2) + \sigma^2/2\alpha P \mathbb{E}[|h|^2]))} + O(\gamma^3) \\ &= \frac{\gamma}{2 \ln 2} \left( 1 - \frac{\gamma a}{4} - \frac{\gamma^2 a^2}{24} \right) \left( 1 - \gamma \left( \frac{a}{2} \left( \frac{\mathbb{E}[|h|^4]}{(\mathbb{E}[|h|^2])^2} - 2 \right) + \frac{\sigma^2}{2\alpha P \mathbb{E}[|h|^2]} \right) \right) + O(\gamma^3) \end{aligned} \quad (\text{D.3})$$

which gives

$$\begin{aligned} \lim_{\gamma \rightarrow 0} C(a) &= \frac{\gamma}{2 \ln 2} \left( 1 - \gamma \left( \frac{a}{2} \left( \frac{\mathbb{E}[|h|^4]}{(\mathbb{E}[|h|^2])^2} - \frac{3}{2} \right) + \frac{\sigma^2}{2\alpha P \mathbb{E}[|h|^2]} \right) \right) + O(\gamma^3), \\ \frac{\partial C}{\partial a} &= -\frac{\gamma^2}{4 \ln 2} \left( \frac{\mathbb{E}[|h|^4]}{(\mathbb{E}[|h|^2])^2} - \frac{3}{2} \right) + O(\gamma^3). \end{aligned} \quad (\text{D.4})$$

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**Nicolas Bonneau** graduated from École Polytechnique in 2002 and obtained an Engineer degree from École Nationale Supérieure des Télécommunications in 2004. He received the M.S. degree in algorithmics from Université Paris VI, Paris, in 2003. He is currently working towards Ph.D. degree in wireless communications. His research interests include information theory, ad hoc networks, as well as the applications of random matrix theory and game theory to the analysis of wireless communication systems.



**Mérouane Debbah** was born in Madrid, Spain. He entered the École Normale Supérieure de Cachan (France) in 1996 where he received the M.S. and the Ph.D. degrees, respectively, in 1999 and 2002. From 1999 to 2002, he worked for Motorola Labs on Wireless Local Area Networks and prospective 4G systems. From October 2002, he was appointed Senior Researcher at the Vienna Research Center for Telecommunications (ftw.), Vienna, Austria working on MIMO wireless channel modeling issues. He is presently an Assistant Professor with the Department of Mobile Communications of the Institute Eurécom. His research interests are in information theory, signal processing, and wireless communications.



**Eitan Altman** received the B.S. degree in electrical engineering (1984), the B.A. degree in physics (1984), and the Ph.D. degree in electrical engineering (1990), all from the Technion—Israel Institute, Haifa. In 1990, he further received his B.Mus. degree in music composition in Tel-Aviv university. Since 1990, he has been with INRIA (National Research Institute in Informatics and Control) in Sophia-Antipolis, France. His current research interests include performance evaluation and control of telecommunication networks and in particular congestion control, wireless communications, and networking games. He is in the editorial board of several scientific journals: *Stochastic Models*, *JEDC*, *COMNET*, *SIAM SICON*, and *WINET*. He has been the general chairman and the (co-)chairman of the program committee of several international conferences and workshops (on game theory, networking games, and mobile networks). More information can be found at <http://www.inria.fr/mistral/personnel/Eitan.Altman/me.html>.



# Special Issue on Wireless Cooperative Networks

## Call for Papers

### Motivations

Cooperative networks are gaining increasing interest from ICT society due to the capability of improving the performance of communication systems as well as providing a fertile environment for the development of context-aware services.

Cooperative communications and networking is a new communication paradigm involving both transmission and distributed processing promising significant increase of capacity and diversity gain in wireless networks, by counteracting faded channels with *cooperative diversity*.

On one hand, the integration of long-range and short-range wireless communication networks (e.g., infrastructured networks such as 3G, wireless ad hoc networks, and wireless sensor networks) improves the performance in terms of both area coverage and quality of service (indeed representing a form of diversity reflected in a greater capacity and number of potential users). On the other hand, the cooperation among heterogeneous nodes, as in the case of wireless sensor networks, allows a distributed space-time signal processing supporting, with a reduced complexity or energy consumption per node, environmental monitoring, localization techniques, distributed measurements, and so forth.

The relevance of this topic is also reflected by numerous sessions in current international conferences on the field as well as by the increasing number of national and international projects worldwide financed on these aspects.

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- Device constraints (power, energy, multiple access, synchronization) and resource management
- Distributed processing and resource management for cooperative networks, for example, distributed

compression in wireless sensor networks, channel and network codes design

- Performance metrics, for example, capacity, cost, outage, delay, energy, scaling laws
- Cross-layer issues, for example, PHY/MAC/NET interactions, joint source-channel coding, separation theorems
- Multiterminal information theory
- Multihop communications
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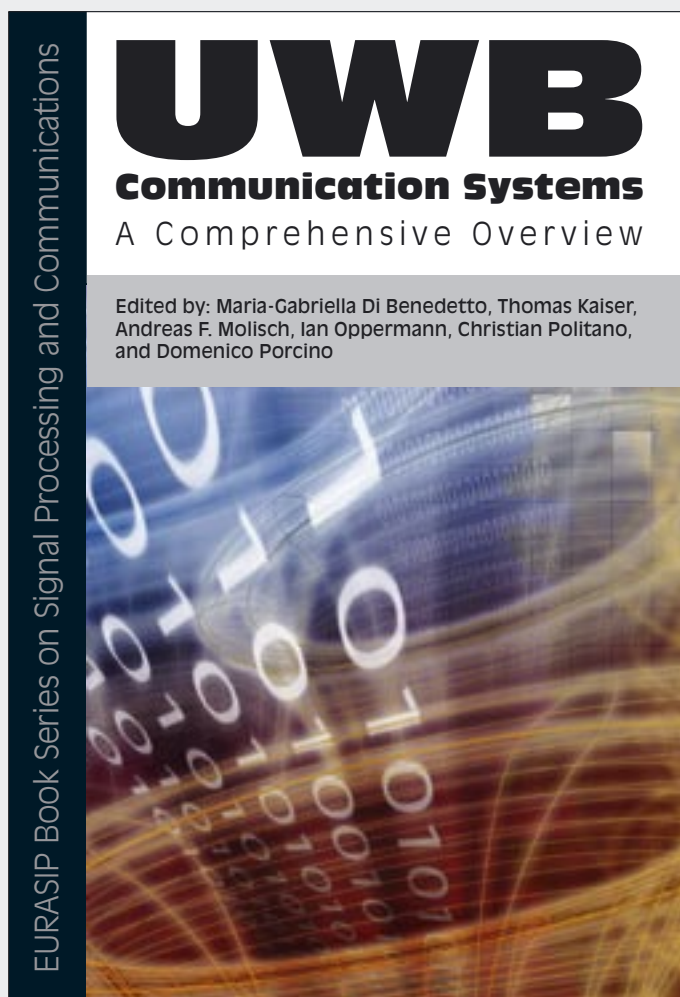
**Hyundong Shin**, Department of Radio Communication Engineering, School of Electronics and Information, Kyung Hee University, Gueonggi-Do 449-701, South Korea, Korea; [hshin@khu.ac.kr](mailto:hshin@khu.ac.kr)

**Ramesh Annavajjala**, ArrayComm Inc., San Jose, CA 95131-1014, USA; [ramesh.annavajjala@gmail.com](mailto:ramesh.annavajjala@gmail.com)

**Moe Win**, Department of Aeronautics and Astronautics, Massachusetts Institute of Technology, Cambridge, MA 02139, USA; [moewin@mit.edu](mailto:moewin@mit.edu)

# UWB Communication Systems—A Comprehensive Overview

Edited by: Maria-Gabriella Di Benedetto,  
Thomas Kaiser, Andreas F. Molisch, Ian Oppermann,  
Christian Politano, and Domenico Porcino



Ultrawideband (UWB) communication systems offer an unprecedented opportunity to impact the future communication world.

The enormous available bandwidth, the wide scope of the data rate / range trade-off, as well as the potential for very low-cost operation leading to pervasive usage, all present a unique opportunity for UWB systems to impact the way people and intelligent machines communicate and interact with their environment.

The aim of this book is to provide an overview of the state of the art of UWB systems from theory to applications. Due to the rapid progress of multidisciplinary UWB research, such an overview can only be achieved by combining the areas of expertise of several scientists in the field. More than 30 leading UWB researchers and practitioners have contributed to this book covering the major topics relevant to UWB. These topics include UWB signal processing, UWB channel measurement and modeling, higher-layer protocol issues, spatial aspects of UWB signaling, UWB regulation and standardization, implementation issues, and UWB applications as well as positioning.

The book is targeted at advanced academic researchers, wireless designers, and graduate students wishing to greatly enhance their knowledge of all aspects of UWB systems.

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