

# Joint Downlink Power Control and Multicode Receivers for Downlink Transmissions in High Speed UMTS

Bessem Sayadi, Stefan Ataman, and Inbar Fijalkow

*ETIS/ENSEA, University of Clergy-Pontoise/CNRS, 6 Avenue du Ponceau, 95014 Clergy-Pontoise, France*

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We propose to combine the gains of a downlink power control and a joint multicode detection, for an HSDPA link. We propose an iterative algorithm that controls both the transmitted code powers and the joint multicode receiver filter coefficients for the high-speed multicode user. At each iteration, the receiver filter coefficients of the multicode user are first updated (in order to reduce the intercode interferences) and then the transmitted code powers are updated, too. In this way, each spreading code of the multicode scheme creates the minimum possible interference to others while satisfying the quality of service requirement. The main goals of the proposed algorithm are on one hand to decrease intercode interference and on the other hand to increase the system capacity. Analysis for the rake receiver, joint multicode zero forcing (ZF) receiver, and joint multicode MMSE receiver is presented. Simulation is used to show the convergence of the proposed algorithm to a fixed point power vector where the multicode user satisfies its signal-to-interference ratio (SIR) target on each code. The results show the convergence behavior for the different receivers as the number of codes increases. A significant gain in transmitted base station power is obtained.

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## 1. INTRODUCTION

As wireless access to the internet rapidly expands, the need for supporting multirate services (voice, data, multimedia, etc.) over limited spectrum increases. CDMA technologies are being considered for third-generation wireless networks, UMTS. There are hence two channelization schemes for achieving multirate transmissions. The first, known as the variable spreading factor scheme, achieves variable-data rate transmission by assigning the radio link a single variable-length random spreading sequence. However, short codes, when subjected to a large delay-spread multipath channel lose their orthogonality and lead to a significant intersymbol interference (ISI). To circumvent this limitation, we consider the second option called multicode transmission. The high-rate data stream is split into several lower rate data substreams [1]. Each substream is spread by a specific spreading sequence and all the substreams are then transmitted synchronously as virtual users. A future transmission mode such as the high-speed downlink packet access (HSDPA [2]) will make wide use of multicode to considerably increase the data rate in the downlink with a peak-data rate in the range of 10–14 Mbit/s. All the spreading sequences are orthogonal to each other to avoid signal interference between parallel channel

codes in a synchronous multipath free channel. However, multipath propagation partially destroys the orthogonality of the multicode transmission and leads to a significant self-intercode interference which increases with the number of parallel codes for a multicode scheme. Therefore, the quality of the downlink under frequency selective fading environments is interference limited. In this paper, we consider a single cell environment where one or more users employ a multicode downlink transmission.

In order to improve the quality of the downlink which is typically defined in terms of the signal-to-interference ratio (SIR), a joint multicode reception was recently proposed in [3] with the assumption that the different codes have a fixed transmitting power. Based on a description of the signal received over fading code-division multiple-access channel, where many different data rates are considered, it is shown in [3] that the problem of recovering the multicode user can be expressed as a multiuser interference cancellation problem, where each channel code represents a virtual user.

Independently in literature, power control is proposed, classically for the link between the multiusers and the base station (BS), to overcome the near-far problem, to maintain the mobile station power consumption, and to reduce the cochannel interference. The power control approach assumes

that a fixed receiver, usually the conventional (single user) receiver, is being used. It optimizes the communication between the mobiles and the BS by controlling the transmitted powers of the different users [4, 5].

Given the importance of power control, an extensive research is focused on this subject. In [6], two optimization criteria are considered in a single-cell case: minimizing total transmitted power and maximizing throughput. In [7], the optimum power vector is given and also statistics on the received power are considered. A statistical approach of the optimum power solution is developed in [8]. The existence (or feasibility) of this optimal power allocation is also considered in [7, 9]. A distributed and iterative power control algorithm where each user's power converges to the minimum power needed to meet its quality of service (QoS) specification is proposed in [10]. A joint optimization of both receiver filters and user transmit powers has been considered in [11] to find the jointly optimum powers and linear MMSE (minimum mean square error) filter coefficients. A similar approach is proposed in reference [12] where the authors employ a successive interference cancelation scheme. Recently, a unified approach of the uplink power control that is applicable to a large family of multiuser receivers is proposed in [13, 14], based on the large system results published in [15].

Based on the fact that for a fixed base station assignment the feasibilities of uplink and downlink are equivalent (see [16] for more details), the authors in [16] present a joint power control and base station assignment for the downlink. Many others researchers are interested on the study of the downlink power control such as [17–19]. In [17], the authors studied the joint optimal power control and beamforming in wireless networks. In [18], the authors studied the downlink power control allocation for multiclass wireless systems. However, in the case of HSDPA system, the way the base station (BS) must allocate the power on the different codes in the case of multicode transmission is still an open issue. It is indeed desirable for the BS not to use more transmission power than what it needs to. This paper proposes a possible way to solve this problem.

In order to achieve this goal, we propose in this paper to combine the downlink power control approach and the joint multicode detection, presented in [3], for the multicode user. We propose an algorithm which controls both the transmitted code powers at the BS and the joint multicode receiver filters implemented in the mobile. The resulted algorithm adapts the transmitted code's powers taking into account a multicode reception strategy at the mobile which aims to reduce the intercode interference. Mathematically, the strategy involves two alternate optimization problems which are resolved iteratively in the proposed algorithm. At each iteration first the receiver filter coefficients of the multicode user are updated to reduce the intercode interference and then the transmitted code powers are updated and assigned. So that, each spreading code of the multicode scheme creates the minimum possible interference to others while satisfying the quality of service requirement. This algorithm has as main goals to decrease intercode interference and to increase the system capacity. Using downlink power

control, the BS output power is adapted to the radio link conditions.

The implementation of this approach, in the HSDPA mobile, requires interference measurements for each code. These measurements are envisaged in HSDPA standard [20]. We show, using simulations, that the resulting algorithm converges to a fixed point power vector where the multicode user satisfies its signal-to-interference ratio (SIR) target on each code. The feasibility of the proposed approach is based on the transmission of the requested code power via a feedback link in order to update the BS output powers. Such a feedback is considered in the HSDPA standard where the mobile transmits the channel quality indicator to the base station [2]. In this study, we consider the case of the joint zero forcing and the joint minimum mean square error (MMSE) multicode linear receivers for various scenarios where we compare their performance to those obtained by considering a bank of rake receivers considered, here, as the conventional power control strategy.

The paper is organized as follows. Section 2 introduces the proposed linear algebraic model which describes the signal received over time-dispersive fading channel including a hybrid multicode/variable spreading factor transmissions. Section 3 gives the problem statement. The proposed strategy is introduced in Sections 4 and 5, and its performance in a simplified HSDPA environment is assessed by means of numerical simulations in Section 6. Finally, Section 7 presents our conclusions.

Throughout this paper scalars, vectors, and matrices are lower case, lower-case bold and upper-case bold characters, respectively.  $(\cdot)^T$ ,  $(\cdot)^{-1}$  denote transposition and inversion, respectively. Moreover,  $E(\cdot)$  denotes the expected value operator.

## 2. SYSTEM MODEL

We assume a multicode CDMA frequency division duplex cellular system. In each cell,  $K$  mobile users, each employing a different rate, communicate with a base station. Each user receives a frame with a standardized number of chips denoted by  $N_{\text{chip}}$ . Based on the quality of service required by user  $k$ , the base station assigns  $M_k$  spreading codes, the processing gain is denoted by  $G_k$ , at the condition that  $N_{\text{chip}} = G_k N_{\text{bit}}^{(k)}$  where  $N_{\text{bit}}^{(k)}$  is the number of transmitted symbols for user  $k$ . Under the constraint that a constant chip rate,  $1/T_c$ , where  $T_c$  denotes the chip period, must be maintained, the symbol period, denoted here by  $T_{s,k} = G_k T_c$ , varies with the requested rate by user  $k$ . The index  $s$  is related to the symbol period and the index  $k$  is related to the  $k$ th user. In order to facilitate the description, the terminologies defined in Table 1 are used in the rest of this paper.

The path-loss attenuation between the BS and the  $k$ th user is denoted by  $z_k$ . In the no-shadowing scenario, the path loss (PL) is modeled as a simple distance-dependent loss:

$$z_k^{(\text{PL})} \approx \lambda d_k^{-\zeta} \quad (1)$$

TABLE 1: Terminology description.

Notation	Description
$K$	the number of user
$N_{\text{chip}}$	the number of chips in a one radio block
$G_k$	the spreading factor assigned to the $k$ th user
$M_k$	the number of spreading code assigned to the $k$ th user
$N_{\text{bit}}^{(k)}$	the number of bits or symbols transmitted in a one radio block
$T_c$	the common chip period
$T_{s,k}$	the symbol period related to the $k$ th user, $1 \leq k \leq K$
$z_k$	the attenuation due to the path loss and the shadowing
$L$	the number of paths
$\tau_i$	the delay of the $i$ th path
$p_m^{(k)}$	the power of the $m$ th code, $1 \leq m \leq M_k$ of the $k$ th user
$n$	the symbol index time
$\mathbf{b}^{(k)}$	the transmitted symbol vector by the $k$ th user
$\mathbf{C}^{(k)}$	the spreading coding matrix related to the $k$ th user
$\mathbf{W}^{(k)}$	the code's power matrix related to the $k$ th user
$\mathbf{H}^{(k)}$	the channel matrix related to the $k$ th user
$\mathbf{n}$	the noise vector

or, in dB,

$$z_k^{(\text{PL})} [\text{dB}] \approx 10 \log_{10}(\lambda) - 10 \cdot \zeta \cdot \log_{10}(d_k), \quad (2)$$

where the constants  $\lambda$  usually depend on the frequency used, as well as the height of the base station and the wireless terminal. The  $d_k$  is the distance from user  $k$  to the base station. The attenuation coefficient  $\zeta$  is usually between 2 and 6 for most indoor and outdoor environments. The model presented in (1) is a general form for the most empirical and semiempirical path-loss attenuation model. For more details, the reader can refer to [21].

In the shadowing case (SH), the variation due to shadowing is added to the path-loss value to obtain the variations. Therefore, the path-loss can be modeled as the product of a distance-dependent path-loss attenuation and a random log-normally distributed shadowing effect [21]:

$$z_k^{(\text{PL,SH})} \approx \lambda d_k^{-\zeta} 10^{\xi_k/10}, \quad \xi_k \sim \mathcal{N}(0, \sigma_\xi^2) \quad (3)$$

or, in dB,

$$z_k^{(\text{PL,SH})} [\text{dB}] \approx 10 \log_{10}(\lambda) - 10 \cdot \zeta \cdot \log_{10}(d_k) + \xi_k, \quad (4)$$

where  $\mathcal{N}(0, \sigma_\xi^2)$  is the Gaussian density with mean 0 (in dB) and variance  $\sigma_\xi^2$  (in dB). In the rest of the paper, we denote  $z_k^{(\text{PL,SH})}$  by  $z_k$ .

The effect of the downlink multipath channel is represented by a vector with  $L$  paths denoted, here, by

$$\mathbf{h} = [\alpha_0, \alpha_1, \dots, \alpha_{L-1}]^T \quad (5)$$

with corresponding delays  $[\tau_0, \dots, \tau_{L-1}]$ . Therefore, the

channel, corresponding to user  $k$ , is described as the following:

$$\mathbf{h}_k = z_k \mathbf{h}. \quad (6)$$

The transmit power towards the  $k$ th user on  $m$ th code will be denoted by  $p_m^{(k)}$ . The transmitted signal for the  $k$ th user can be written as

$$y_k(t) = \sum_{n=0}^{N_{\text{bit},k}-1} \sum_{m=1}^{M_k} \sqrt{p_m^{(k)}} b_m^{(k)}(n) c_m^{(k)}(t - nT_{s,k}), \quad (7)$$

where

$$c_m^{(k)}(t) = \sum_{q=0}^{G_k-1} c_m^{(k),(q)} \psi(t - qT_c) \quad (8)$$

with  $G_k$  the spreading factor for the  $k$ th user and  $b_m^{(k)}(n)$  is the transmitted symbol at time  $n$  for the  $k$ th user on the  $m$ th channel-code denoted by  $c_m^{(k)}(t) \cdot \psi$  is a normalized chip waveform of duration  $T_c$ . The base-band received signal at the desired user can be written as

$$r(t) = \sum_{k=1}^K z_k \sum_{l=0}^{L-1} \alpha_l \sum_{n=0}^{N_{\text{bit},k}-1} \sum_{m=1}^{M_k} \sqrt{p_m^{(k)}} b_m^{(k)}(n) c_m^{(k)}(t - nT_{s,k} - \tau_l) + n(t), \quad (9)$$

where  $n(t)$  is a zero-mean additive white Gaussian noise (AWGN) process.

The received signal is time-discretized at the rate of  $1/T_c$ , leading to a chip-rate discrete-time model that can be written as

$$r_l = r(lT_c) = \sum_{k=1}^K z_k \sum_{l=0}^{L-1} \alpha_l \sum_{n=0}^{N_{\text{bit},k}-1} \sum_{m=1}^{M_k} \sqrt{p_m^{(k)}} b_m^{(k)}(n) c_m^{(k)}((l - nG_k - t_{l,k})T_c) + n(lT_c), \quad (10)$$

where  $t_{l,k} = \lfloor \tau_l/G_k \rfloor$  is the time-discretized path delay in sample intervals (chip period).

Throughout the paper, we employ a block model. The blocks of transmitted symbols for each user,  $k = 1, \dots, K$ , are concatenated in a vector:

$$\mathbf{b}^{(k)} = [b_1^{(k)}(0), \dots, b_{M_k}^{(k)}(0), \dots, b_{M_k}^{(k)}(N_{\text{bit}}^{(k)} - 1)]^T \quad (11)$$

containing  $N_{\text{bit}}^{(k)}$  bits transmitted with the different codes for a given user,  $k$ .

The transmission of the data sequence over the CDMA channel can be expressed by the received sequence  $\mathbf{r}$  [3]:

$$\mathbf{r} = [r_1, \dots, r_{N_{\text{chip}}+L-1}]^T = \sum_{k=1}^K \mathbf{C}^{(k)} \tilde{\mathbf{H}}^{(k)} \mathbf{W}^{(k)} \mathbf{b}^{(k)} + \mathbf{n}, \quad (12)$$

where  $\tilde{\mathbf{H}}^{(k)} = \text{diag}(\mathbf{h}_k, \dots, \mathbf{h}_k)$  is of size  $(N_{\text{bit}}^{(k)} M_k L, N_{\text{bit}}^{(k)} M_k)$  and  $\mathbf{W}^{(k)} = \text{diag}(\mathbf{P}^{(k)}, \mathbf{P}^{(k)}, \dots, \mathbf{P}^{(k)})$  of size  $N_{\text{bit}}^{(k)} M_k$  where  $\mathbf{P}^{(k)} = \text{diag}(\sqrt{p_1^{(k)}}, \sqrt{p_2^{(k)}}, \dots, \sqrt{p_{M_k}^{(k)}})$  and  $\text{diag}(\mathbf{X})$  represents the diagonal matrix containing only the diagonal elements of the matrix  $\mathbf{X}$ . The matrix  $\mathbf{C}^{(k)}$  represents the code matrix of size  $((N_{\text{chip}} + L - 1), N_{\text{bit}}^{(k)} M_k L)$  built as follows:

$$\begin{aligned} \mathbf{C}^{(k)} &= [\mathbf{v}_{0,0,0}^k, \dots, \mathbf{v}_{N_{\text{bit},k}-1, M_k-1, L-1}^k], \\ \mathbf{v}_{n,m,l}^k &= [\mathbf{0}_{nG_k}^T, \mathbf{u}_{m,l}^{kT}, \mathbf{0}_{(N_{\text{bit},k}-n-1)G_k}^T]^T, \\ \mathbf{u}_{m,l}^k &= [\mathbf{0}_{t_l}^T, \mathbf{c}_m^{kT}, \mathbf{0}_{L-t_l-1}^T]^T, \\ \mathbf{c}_m^k &= [c_m^k(1), \dots, c_m^k(G_k)]^T, \end{aligned} \quad (13)$$

where  $n = 0, \dots, N_{\text{bit},k}-1$ ,  $m = 0, \dots, M_k-1$ , and  $l = 0, \dots, L-1$ .

$\mathbf{0}_n$  denotes the null vector of size  $n$ . The vector  $\mathbf{n}$ , of length  $N_{\text{chip}} + L - 1$ , represents the channel noise vector with  $N_0$  as a power spectral density.

The vector  $\mathbf{c}_m^k = [c_m^k(1), \dots, c_m^k(G_k)]^T$  denotes the spreading code vector of length  $G_k$  related to the  $k$ th user. It is obtained by the discretization at the chip rate of the function  $c_m^k(t)$  given by (8). The index  $m$  denotes the index of the spreading code in the multicode scheme containing  $M_k$  codes.

The model just proposed for a multirate and multicode DS-CDMA system follows the structural principles of practical downlink UMTS and leads to a convenient algebraic form which allows for a powerful receiver design for a multicode multirate CDMA system.

For the sake of simplicity, the propagation channel is assumed to be time invariant during the transmission of  $N_{\text{chip}}$  chips. We also assume that the interferences due to symbols before and after  $N_{\text{chip}}$  data block can be completely cancelled. This is possible when those interfering symbols are known by the receiver via a training sequence. The model presented in (12) can be generalized to incorporate scrambling codes and multiple antenna transmissions.

### 3. PROBLEM STATEMENT

Without loss of generality, the user 1 is chosen as the user of interest. By denoting  $\mathbf{A}^{(k)} = \mathbf{C}^{(k)} \tilde{\mathbf{H}}^{(k)}$ , the received signal can be expressed as

$$\mathbf{r} = \underbrace{\mathbf{A}^{(1)} \mathbf{W}^{(1)} \mathbf{b}^{(1)}}_{\text{desired signal + intercode interference}} + \underbrace{\sum_{k=2}^K \mathbf{A}^{(k)} \mathbf{W}^{(k)} \mathbf{b}^{(k)}}_{\text{MAI + ISI}} + \underbrace{\mathbf{n}}_{\text{noise}}, \quad (14)$$

where we separate the user of interest's signal, the multiple access interference (MAI), and intersymbol interference (ISI) caused by the other users and the noise. The first term in (14) contains the useful signal and the intercode interference caused by the multicode scheme.

Let  $\mathbf{F}$  denote the joint multicode receiver filter employed by the receiver of user 1, user of interest. From the output of the joint multicode receiver,  $\mathbf{y} = \mathbf{F}^T \mathbf{r}$ , the SIR of virtual user of interest can be written for code  $m$  and symbol  $n$  as the following:

$$\text{SIR}(m, n) = \frac{p_m E(\beta(\mathbf{F}, \mathbf{h}_k, \mathbf{C}^{(k)}) |\hat{b}_m^{(1)}(n)|^2)}{E(|\Omega(p_{m' \neq m})|^2)} \quad (15)$$

for  $m=1, \dots, M_1$ ,  $m'=1, \dots, M_1$ , and  $n=1, \dots, N_{\text{bit},1} \cdot \Omega(p_{m' \neq m})$  is the sum of the intercode interferences, the multiple access interference, the intersymbols interference, and the noise.  $\beta(\mathbf{F}, \mathbf{h}_k, \mathbf{C}^{(k)})$  denotes the term depending on the multicode receiver filter coefficients, the spreading code and the channel coefficients.  $p_m$  denotes the power assigned to the  $m$ th code. In the sequel, we present the expression of the terms  $\beta(\mathbf{F}, \mathbf{h}_k, \mathbf{C}^{(k)})$  and  $\Omega(p_{m' \neq m})$  in the case of the rake, the zero forcing, and the MMSE multicode receivers.

The aim of the power control algorithm in CDMA system is to assign the mobile the minimum power necessary to achieve a certain QoS which is typically defined in terms of SIR. In this context, the most employed power control algorithm was proposed by Foschini and Miljanic in [10] and it is known as distributed power control (DPC). The optimum transmission power of user  $k$ , supposed monocode user, is computed iteratively in order to achieve an SIR target denoted here by  $\text{SIR}_{\text{target}}$ .

$$p_k(n+1) = \frac{\text{SIR}_{\text{target}}}{\text{SIR}(n)} p_k(n). \quad (16)$$

When the target SIR is achieved, the power's updating stops. This approach assumes a fixed receiver, usually a single receiver. To overcome this limitation, Ulukus and Yates in [11] proposes to optimize jointly the multiuser receiver and the user's power in the uplink. As the main result, it is shown that the same performance as the DPC algorithm is achieved with less transmitted power. In continuation of Yates' idea of a combined power control and receiver adaptation in a CDMA uplink, we develop, here, a joint power control and multicode receiver adaptation algorithm suitable for a high-speed UMTS downlink.

So, the problem is to determine the different code powers,  $p_m$ , and multicode receiver filter coefficients, such that the allocated power to the multicode user is minimized while satisfying the quality of service requirement on each code,  $\text{SIR}_m \geq \text{SIR}_{\text{target}}$ , where  $\text{SIR}_m = E_n((\text{SIR}(m, n)))$ ,  $m = 1, \dots, M_1$ , and  $\text{SIR}_{\text{target}}$  is the minimum acceptable level of SIR for each code.  $E_n$  denotes the expectation over the symbol index. Therefore, the problem can be stated mathematically as follows:

$$\min_{\mathbf{P}} \sum_{m=1}^{M_1} p_m \quad (17)$$

constrained to

$$p_m \geq \text{SIR}_{\text{target}} \frac{E\left(|\Omega(p_{m' \neq m})|^2\right)}{E\left(\beta(\mathbf{F}, \mathbf{h}_k, \mathbf{C}^{(k)}) |\hat{b}_m^{(1)}(n)|^2\right)} \quad (18)$$

$$p_m \leq p_{\max}, \quad m = 1, \dots, M_1,$$

where  $p_{\max}$  denoted the maximum allowed transmitted user's power.

The following optimization problem is difficult since the constraints denominators are also power dependent. The solution is to consider a double optimization problem where an inner optimization is inserted in the constraint set as the following:

$$\min_{\mathbf{p}} \sum_{m=1}^{M_1} p_m \quad (19)$$

constrained to

$$p_m \geq \text{SIR}_{\text{target}} \min_{\mathbf{F}} \frac{E\left(|\Omega(p_{m' \neq m})|^2\right)}{E\left(\beta(\mathbf{F}, \mathbf{h}_k, \mathbf{C}^{(k)}) |\hat{b}_m^{(1)}(n)|^2\right)}, \quad (20)$$

$$p_m \leq p_{\max}, \quad m = 1, \dots, M_1.$$

In [11], the equivalence between the optimization formulation given by (17) and the formulation given by (19) is demonstrated.

The second optimization formulation is a two alternate optimization problem. The first optimization problem involved in (19), and called the outer optimization, is defined over the code power. Whereas the second one, called the inner optimization, which is involved in (20), assumes a fixed power vector. It is defined over the filter coefficients of the multicode receiver. In this stage, we optimize the multicode filter coefficients to maximally suppress the intercode interference. The implementation of these two alternate optimization problems are realized iteratively in the algorithm described in the next section.

#### 4. COMBINED DOWNLINK POWER CONTROL AND JOINT MULTICODE RECEIVERS

In this section, we propose to combine the downlink power control and the joint multicode receivers. The objective of the algorithm is to achieve an output SIR equal to a target  $\text{SIR}_{\text{target}}$  for each assigned code to the multicode user. To do this, we exploit the linear relationship between the output SIR and transmit code power as is seen in (15). The proposed algorithm is a two-stage algorithm. First, we adjust the filter coefficients for a fixed code power vector, the inner optimization. Second, we update the transmitted code powers to meet the SIR constraints on each code for the chosen filter coefficients using (16). The description of the proposed algorithm is as follows:

The subscript 1 marks out the considered multicode user.

If we consider also a maximum transmit power limitation  $p_m^{\max}$ , for  $m = 1, \dots, M_1$ , step (3) from the above algorithm is

- (1)  $i = 0$ , start with initial powers  $p_0^{(1)}, \dots, p_{M_1}^{(1)}$ .
- (2) Receiver parameter calculation and receiver output SIR calculation.
- (3) Update the code powers using
 
$$p_m^{(1)}(i+1) = (\text{SIR}_{\text{target}} / E_n[\text{SIR}(m, n)]) p_m^{(1)}(i), \text{ for } m = 1, \dots, M_1.$$
- (4)  $[\mathbf{W}(i+1)]_{j,j} = \sqrt{p_m^{(1)}(i+1)}$ , with  $j = m + (n-1)M_1$  where  $m = 1, \dots, M_1$  and  $n = 1, \dots, N_{\text{bit},1}$ .
- (5)  $i = i + 1$ , stop if convergence is reached; otherwise, go to step (2).

#### ALGORITHM 1

modified according to

$$p_m^{(1)}(i+1) = \min \left\{ \frac{\text{SIR}_{\text{target}}}{E_n[\text{SIR}(m, n)]} p_m^{(1)}(i), p_m^{\max} \right\}. \quad (21)$$

The new code power calculated in step (3) are transmitted via a feedback link to the BS.

In the sequel, we present the SIR derivation in the case of the zero forcing and the MMSE multicode joint receivers.

### 5. JOINT MULTICODE RECEIVER STRUCTURES

In this section, we derive the expression of the output SIR on each code by considering the joint multicode receivers: ZF and MMSE.

The received signal given by (14) can be written as

$$\mathbf{r} = \mathbf{A}\mathbf{W}\mathbf{b} + \tilde{\mathbf{n}} \quad (22)$$

by denoting  $\tilde{\mathbf{n}} = \sum_{k=2}^K \mathbf{A}^{(k)} \mathbf{W}^{(k)} \mathbf{b}^{(k)} + \mathbf{n}$ .

#### 5.1. Rake receiver

The conventional data estimator consists of a bank of rake receivers. In this case, the output signal is

$$\mathbf{y}_{\text{Rake}} = \mathbf{A}^H \mathbf{r} = \mathbf{\Gamma}\mathbf{W}\mathbf{b} + \mathbf{A}^H \tilde{\mathbf{n}}, \quad (23)$$

where  $\mathbf{\Gamma} = \mathbf{A}^H \mathbf{A}$ .

We separate the desired user's symbols, the intercode interference generated by the multicode transmission and the MAI + ISI + noise generated by the noise and the other users,

$$\mathbf{y}_{\text{Rake}} = \underbrace{\text{diag}\{\mathbf{\Gamma}\mathbf{W}\mathbf{b}\}}_{\text{desired symbols}} + \underbrace{\overline{\text{diag}\{\mathbf{\Gamma}\mathbf{W}\mathbf{b}\}}}_{\text{intercode interference}} + \underbrace{\mathbf{A}^H \tilde{\mathbf{n}}}_{\text{MAI + ISI + noise}}, \quad (24)$$

where  $\overline{\text{diag}}(\mathbf{X}) = \mathbf{X} - \text{diag}(\mathbf{X})$  represents a matrix with zero diagonal elements containing all but the diagonal elements of  $\mathbf{X}$ .

The useful signal for the  $n$ th transmitted symbol on the  $m$ th code is given by

$$E \left\{ \left( [\mathbf{\Gamma}\mathbf{W}]_{j,j} b_1^{(1)}(n) \right)^2 \right\} = ([\mathbf{\Gamma}\mathbf{W}]_{j,j})^2 E \left\{ \left| b_1^{(1)}(n) \right|^2 \right\}, \quad (25)$$



where  $[\mathbf{X}]_{j,j}$  denotes the element in the  $j$ th row and  $j$ th column of the matrix  $\mathbf{X}$ .

The interference and the noise are given by

$$I = E \left\{ (\Gamma \mathbf{W} \mathbf{b} - \text{diag}\{\Gamma \mathbf{W} \mathbf{b}\} + \mathbf{A}^H \tilde{\mathbf{n}})^2 \right\}. \quad (26)$$

We consider in the sequel that  $E\{|b_1^{(1)}(n)|^2\} = 1$ .

After developing the term  $I$  and taking the  $j$ th diagonal element, the SIR at the output of the rake receiver related to the  $n$ th transmitted symbol on the  $m$ th code can be expressed as follows by denoting  $\Gamma' = \Gamma \mathbf{W}$  and  $\mathbf{R}'_{\tilde{\mathbf{n}}} = E\{\tilde{\mathbf{n}}\tilde{\mathbf{n}}^T\}$  as the covariance matrix of the MAI, ISI and noise,

$$\text{SIR}_{\text{Rake}}(m, n) = \frac{([\Gamma']_{j,j})^2}{[(\Gamma')^2]_{j,j} - [(\Gamma')_{j,j}]^2 + [\Gamma' \mathbf{R}'_{\tilde{\mathbf{n}}} \Gamma']_{j,j}} \quad (27)$$

for  $j = m + (n-1)M_1$  where  $m = 1, \dots, M_1$  and  $n = 1, \dots, N_{\text{bit},1}$ .

### 5.2. Joint multicode zero forcing receiver

In the case of the joint ZF receiver, the output signal is

$$y_{\text{ZF}} = \Gamma^{-1} \mathbf{y}_{\text{Rake}} = \mathbf{W} \mathbf{b} + \Gamma^{-1} \mathbf{A}^H \tilde{\mathbf{n}}. \quad (28)$$

The joint ZF receiver leading to the estimate of the desired symbols,  $\mathbf{b}$ , is called zero forcing since it tries to force the residual intercode interference to zero.

Therefore, the SIR at the output of the joint ZF receiver relating to the  $n$ th transmitted symbol on the  $m$ th code can be expressed as follows:

$$\text{SIR}_{\text{ZF}}(m, n) = \frac{[\mathbf{W}]_{j,j}^2}{[\Gamma^{-1} \mathbf{A}^H \mathbf{R}'_{\tilde{\mathbf{n}}} \mathbf{A} \Gamma^{-H}]_{j,j}} \quad (29)$$

for  $j = m + (n-1)M_1$  where  $m = 1, \dots, M_1$  and  $n = 1, \dots, N_{\text{bit},1}$ .

### 5.3. Joint multicode MMSE receiver

The joint multicode MMSE linear receiver minimizes the output mean squared error

$$E \left\{ \|\mathbf{F} \mathbf{y}_{\text{Rake}} - \mathbf{W} \mathbf{b}\|^2 \right\} \quad (30)$$

with respect to  $\mathbf{F}$  which yields

$$\mathbf{F} = \mathbf{W}^2 \Gamma^H [\Gamma \mathbf{W}^2 \Gamma^H + \mathbf{A}^H \mathbf{R}'_{\tilde{\mathbf{n}}} \mathbf{A}]^{-1}. \quad (31)$$

Therefore, the output signal from the MMSE receiver yields, by denoting  $\mathbf{W}_0 = \mathbf{F} \Gamma$ ,

$$\mathbf{y}_{\text{MMSE}} = \mathbf{F} \mathbf{y}_{\text{Rake}} = \mathbf{W}_0 \mathbf{W} \mathbf{b} + \mathbf{W}_0^{-1} \Gamma \mathbf{A}^H \tilde{\mathbf{n}}. \quad (32)$$

Now, we can separate the desired user's symbols, the intercode interference generated by the multicode transmission and the MAI + ISI + noise generated by the noise and the other users,

$$\mathbf{y}_{\text{MMSE}} = \text{diag}\{\mathbf{W}_0 \mathbf{W} \mathbf{b}\} + \overline{\text{diag}\{\mathbf{W}_0 \mathbf{W} \mathbf{b}\}} + \mathbf{W}_0 \Gamma^{-1} \mathbf{A}^H \tilde{\mathbf{n}}. \quad (33)$$

The SIR at the output of the MMSE receiver relating to the  $n$ th transmitted symbol on the  $m$ th code can be expressed as follows by denoting  $\mathbf{W}' = \mathbf{W}_0 \mathbf{W}$  as

$$\begin{aligned} \text{SIR}_{\text{MMSE}}(m, n) &= \frac{([\mathbf{W}']_{j,j})^2}{[\mathbf{W}' \mathbf{W}'^H]_{j,j} - ([\mathbf{W}']_{j,j})^2 + [\mathbf{W}_0^{-1} \Gamma \mathbf{A}^H \mathbf{R}'_{\tilde{\mathbf{n}}} \mathbf{A} \Gamma^{-1} \mathbf{W}_0^H]_{j,j}} \\ &= \frac{([\mathbf{W}']_{j,j})^2}{[\mathbf{W}' \mathbf{W}'^H]_{j,j} - ([\mathbf{W}']_{j,j})^2 + [\mathbf{W}_0^{-1} \Gamma \mathbf{A}^H \mathbf{R}'_{\tilde{\mathbf{n}}} \mathbf{A} \Gamma^{-1} \mathbf{W}_0^H]_{j,j}} \end{aligned} \quad (34)$$

for  $j = m + (n-1)M_1$  where  $m = 1, \dots, M_1$  and  $n = 1, \dots, N_{\text{bit},1}$ .

The proposed approach involves complex matrix inverse computations due to the employment of instantaneous MMSE filtering. This drawback can be recovered by replacing instantaneous MMSE filtering with adaptive filtering. As is suggested in [22], the least mean square and the minimum output energy algorithms present an ease implementation and analysis. As a future work, we suggest to focus on the complexity reduction of the proposed approach.

## 6. SIMULATION RESULTS

Simulation results analyze the performance of the proposed strategy considering the joint multicode MMSE and the joint ZF receivers, and the performance obtained from the conventional power control which assumes a bank of fixed rake receivers. We compare the different solutions by evaluating the total transmit (or mean transmit) power and the SIR (or mean SIR) at the mobile receiver.

Users are placed randomly in a hexagonal cell with radius  $R = 1000$  m around the BS. The path-loss exponent is taken  $\zeta = 4$  and no shadowing is assumed. We consider a 6-path downlink channel. The target SIR is fixed at  $\text{SIR}_{\text{target}} = 4$  (around 6 dB) for all simulations. We consider a number of  $K = 20$  users, among them we have  $K'$ ,  $K' < K$  multicode users. The spreading factor for the single-code users is  $G_k = 128$  for any  $k = K', \dots, K$ . The multicode users has a spreading gain  $G_{k'} = 64$ ,  $k' = 1, \dots, K'$ . We fix the user 1 as user of interest. We vary its number of allocated codes between  $M_1 = 4$  and  $M_1 = 64$ .

In Figure 1, we plot the mean SIR,  $(1/M_1) \sum_{m=1}^{M_1} \text{SIR}(m)$ , versus iteration index in the case of  $M_1 = 4$  for the conventional power control algorithm (fixed rake receiver) and the proposed strategy which optimizes the joint MMSE and ZF multicode receiver coefficients. We note the one-iteration convergence of the multicode ZF receiver, the fast convergence of the multicode MMSE receiver, and the much slower convergence of the rake receiver.

In the case of  $M_1 = 16$ , the conventional rake receiver cannot meet the target SIR anymore, as shown in Figure 2, where we plot the variation of the  $\text{SIR}(m)$  on each code. However, the multicode receivers (ZF and MMSE) show good performance. Adding more virtual users brings the conventional receiver to even worse performance as is shown in Figure 3.

For  $M_1 = 64$ , the different lines for each receiver type correspond to the variation of the SIR on each code,  $\text{SIR}(m)$ , versus iteration index.

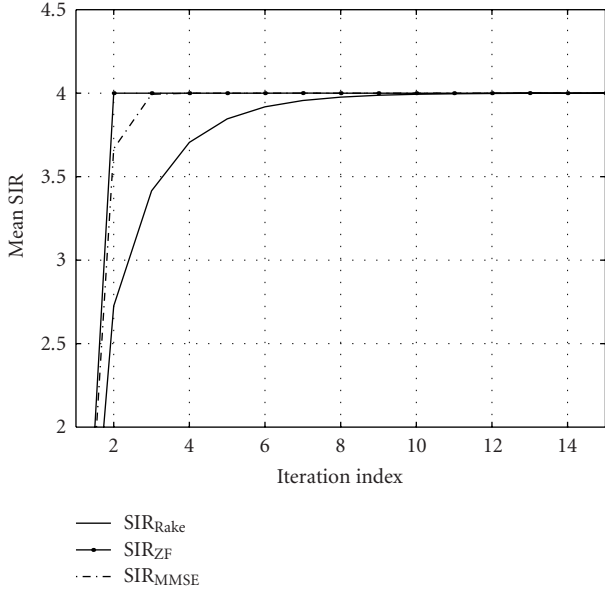


FIGURE 1: The SIR convergence for the rake, ZF, and MMSE receivers in the case  $M_1 = 4$  multicode.

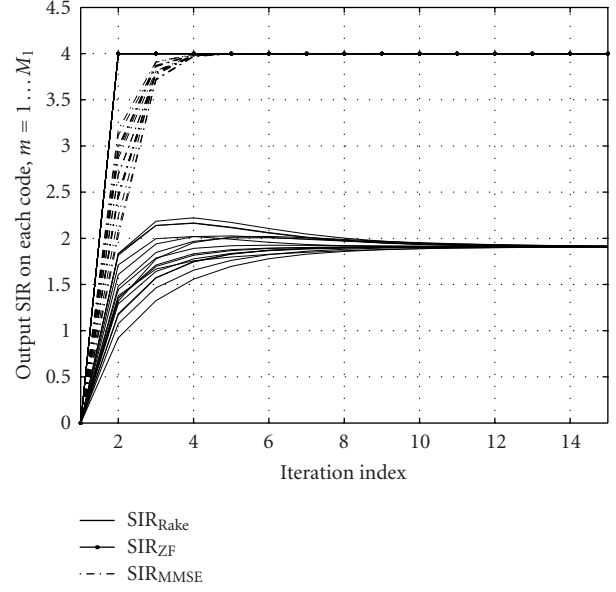


FIGURE 2: The SIR convergence for the rake, ZF, and MMSE receivers in the case  $M_1 = 16$  multicode.

From Figures 2 and 3, we observe the difficulty of the conventional power control to reach the target SIR because of the MAI, ISI, and the intercode interferences. In the case of low load in the cell (few users), the conventional power control reaches the SIR target; see Figure 1. However, in this case, our proposed strategy presents a faster convergence.

The variation of the base station transmit power ratios  $p_{ZF}/p_{Rake}$  and  $p_{MMSE}/p_{Rake}$  versus the iteration index is shown in Figure 4 in the case of a number of codes  $M_1 = 16$  codes of the multicode user. We note a decrease of about 20% of the transmitted BS power.

However, a much significant gain in transmitted BS power is noted in the case of  $M_1 = 64$ , as we can deduce from the results of Figure 5. The MMSE shows its optimality with significantly improved results with respect to the ZF receiver: the MMSE always gains power with respect to the rake receiver (the ratio is smaller than 1) where the ZF increases first the required power to achieve the required SIR.

We observe from Figures 4 and 5 that the proposed strategy of joint downlink power control and multicode receivers outperforms the conventional downlink power control in terms of total transmitted power of the multicode user.

In all simulations, we note the very fast (1 iteration) convergence of the ZF receiver, the fast convergence of the MMSE receiver, and the much slower convergence of the conventional power control. The fast convergence of the ZF receiver is easy to explain: since this receiver performs an orthogonal projection into the subspace formed by the interfering signals, the output desired signal does not depend on the interfering signals' amplitudes. There is only one update of (21). In the case of the joint multicode MMSE receiver, at each iteration the receiver is updated since it depends on the received powers of each code. Finally, the rake receiver is a

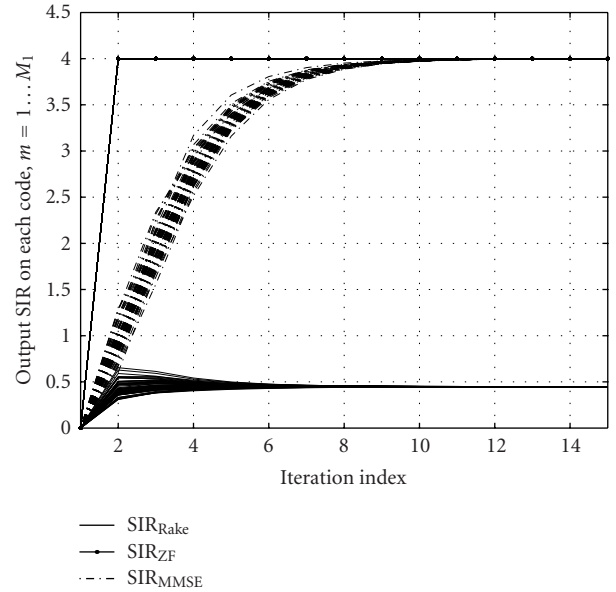


FIGURE 3: The SIR convergence for the rake, ZF, and MMSE receivers in the case  $M_1 = 64$  multicode.

fixed receiver that takes into account only the desired signal processing the MAI, ISI, and intercode interferences as noise, therefore yielding the worst performance.

The best performance in minimizing transmit powers and maximizing the cell capacity is obtained by the MMSE receiver. The ZF receiver shows slightly lower performance, in terms of total transmit power, at high-cell loads (case of  $M_1 = 64$ , see Figure 5).

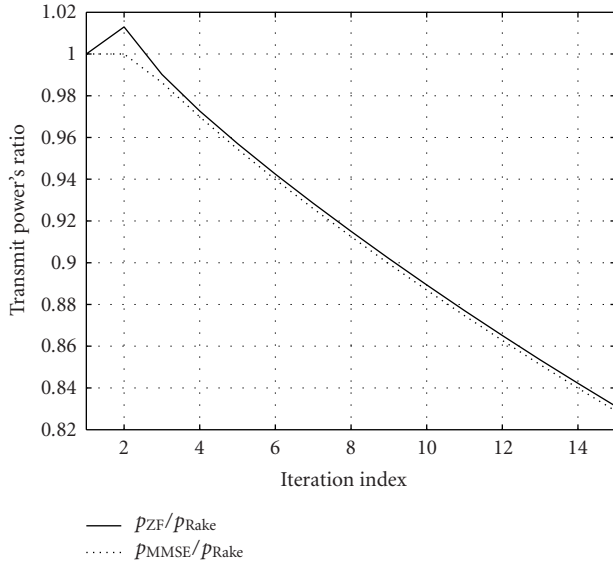


FIGURE 4: The mean total transmit powers ratio  $p_{ZF}/p_{Rake}$  and  $p_{MMSE}/p_{Rake}$  versus the iteration index for  $M_1 = 16$ .

It should be noticed that at very low-cell loads (i.e., few interfering single-code users and few codes for the multicode user (case of  $M_1 = 4$ )) the three receivers show similar performance, a result that is expected.

After the convergence of the proposed strategy using a joint multicode MMSE receiver, the codes' power allocation is shown in Figure 6. As one can notice, it is not the same power per code. This confirms the interest of this power allocation-strategy for the downlink of the multicode user.

## 7. CONCLUSION

In this paper, we have analyzed the benefits of combining the downlink power control and the joint multicode detection for a multicode user. The proposed algorithm updates iteratively the transmitted code powers of the multicode users and the joint multicode receiver filter coefficients. We have used simulations to show the convergence and performance of the proposed algorithm in a system of practical interest. An important gain in transmit power reduction is obtained by implementing joint multicode detection. The performance of the ZF receiver allows an important reduction in computations (step 4 is avoided). The study of theoretical convergence of the proposed algorithm is under investigation based on the analysis proposed in [23].

In order to overcome the limitation of power control due to temporal filtering only, a joint power control and beamforming for wireless network is proposed in [17] where it is shown that a capacity increase is possible if array observations are combined in the MMSE sense. Therefore, as a direction for further research, the combination of the three basic interference cancellation approaches (transmit power control, multiuser detection, and beamforming) represents an

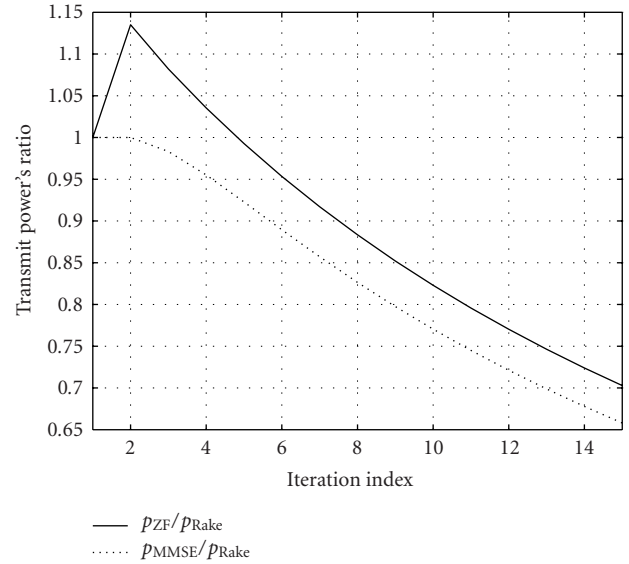


FIGURE 5: The mean total transmit power ratio  $p_{ZF}/p_{Rake}$  and  $p_{MMSE}/p_{Rake}$  versus the iteration index for  $M_1 = 64$ .

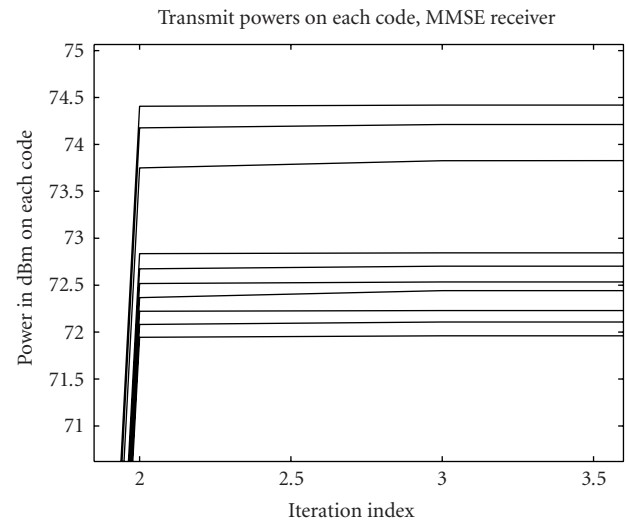


FIGURE 6: The code power allocation in the case of  $M_1 = 10$  codes after convergence.

ambitious challenge to be met by third-generation systems in order to provide high-capacity flexible services.

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**Bessem Sayadi** received the B.S. Engineering degree in signal processing from the Ecole Supérieure des Télécommunications de Tunis (Sup'Com Tunis), Tunisia, in 1999, and both the M.Phil. (2000) and the Ph.D. (2003) degrees from the Signals and Systems Laboratory (LSS) at Supélec, Gif-sur-Yvette, the Paris XI University, Orsay, France. In 1999, he joined France Telecom where he was engaged in research on echo cancellation and adaptive filtering. He has also served as a Teaching Assistant in several courses on digital communications, signal processing, and electronics in the Department of Electronic and Electrical Engineering, SUPÉLEC, ENSEA, and University Paris IX, since September 2000. From 2003 to 2005, he was an Associate Researcher in the Image and Signal Processing Team (ETIS), at ENSEA, Cergy-Pontoise. In 2006, he joined France Telecom as a Research Engineer. His current research interests include Bayesian method, multiuser detection, video coding, radio resource management, IP-mobility, and cross-layer design.



**Stefan Ataman** received the B.S. and M.S. degrees from the Polytechnic University of Bucharest, Romania, in 1999 and 2000, respectively, and the Ph.D. degree from Université Paris-Sud, France, in 2004, all in electrical engineering. Currently, he is working as a Research Associate with University Cergy-Pontoise/ETIS laboratory, France. His research interests are in the areas of digital communications and signal processing with applications to CDMA wireless communications, power control, and multiuser receivers in CDMA cellular systems.



**Inbar Fijalkow** received her Engineering and Ph.D. degrees from Ecole Nationale Supérieure des Télécommunications (ENST), Paris, France, in 1990 and 1993, respectively. In 1993–1994, she was a Research Associate at Cornell University, NY, USA. In 1994, she joined ETIS, UMR 8051 (ENSEA - Cergy-Pontoise University - CNRS) in Cergy-Pontoise, France. Since 2004, she is the head of ETIS. Her current research interests are in signal processing applied to digital communications: iterative (turbo) processing (in particular turbo-equalization), analysis of communication systems (including



MIMO, OFDM, CDMA, etc.) and cross-layer optimization. Until 2005, she has been Member of the board of the GDR ISIS, which is the CNRS French national research group on signal, image, and vision processing. She has been an Associate Editor of the IEEE Transactions on Signal Processing 2000–2003.