Performance of the Smart Antenna Aided Multicarrier DS-CDMA Uplink

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***Abstract*— In this contribution a generalized MC DS-CDMA system invoking smart antennas for improving the achievable performance of the system is studied, which is capable of suppressing the multiuser interference, while achieving frequency, time and spatial diversity. In the considered MC DS-CDMA system the receiver employs multiple receive antennas and each of the receive antennas consists of several antenna array elements. Four types of optimum linear combining schemes are investigated. In these optimum linear combining schemes the weight vectors are derived based on the optimization criteria of Minimum Variance Distortionless Response (MVDR), of Maximum Signal-to-Interference-plus-Noise Ratio (MSINR), of Minimum Mean-Square Error (MMSE) and of Minimum Power Distortionless Response (MPDR). The paper is concluded with a comparative performance study of various antenna array models employing the above optimization criteria.**

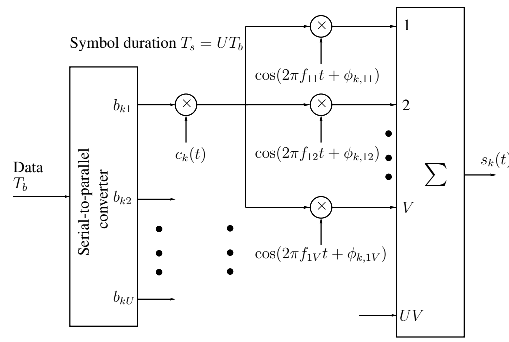

Fig. 1. The kth user's transmitter schematic for the generalized multicarrier DS-CDMA system.

I. Introduction

In recent years numerous research contributions have appeared on the topic of Multi-Carrier Direct Sequence Code Division Multiple Access (MC DS-CDMA), which constitutes an attractive scheme [1],[2],[3],[4], based on a combination of DS-CDMA and OFDM. The multitone DS-CDMA system proposed in [3] and the orthogonal MC DS-CDMA system [1],[4] are capable of efficiently exploiting the transmission bandwidth and mitigating the effects of frequency selective multipath interference, while achieving both frequency and time diversity. On the other hand, smart antennas have been used for improving the performance of wireless systems, since they are capable of radiating and receiving energy in and from the intended directions, which potentially reduces the interference amongst wireless users [5],[6],[7].

In this contribution we discuss the generalized MC DS-CDMA system investigated in [8], [9], which includes the subclasses of multitone DS-CDMA [3] and orthogonal MC DS-CDMA [4] as special cases. The transmitter of the generalized MC DS-CDMA system is portrayed in Fig.1. *The novelty of this paper is that we combine the above-mentioned generalized MC DS-CDMA system [8], [9] with smart antennas for the sake of improving the performance of the system by suppressing the multiuser interference, while achieving frequency, time and spatial diversity.* Specifically, in our considered system the base-station receiver employs multiple receive antennas and each of the receive antennas consists of several antenna array elements. At the receiver the antenna array outputs are combined using one of the four types of optimum linear combining schemes. These optimum linear combining schemes are based on the Minimum Variance Distortionless Response (MVDR), the Maximum Signal-to-Interference-plus-Noise Ratio (MSINR), the Minimum Mean-Square Error (MMSE) and the Minimum Power Distortionless Response (MPDR) optimization criteria, respectively. The achievable performance of the generalized MC DS-CDMA systems using the above-mentioned optimum linear combining schemes is investigated and compared, when operating in various propagation environments. Furthermore, several different antenna array models are employed in our simulations.

The rest of this paper is organized as follows. In Section II the philosophy of the generalized multicarrier DS-CDMA system invoking smart antennas is described and characterized. The statistical analysis of the receiver's decision variable is provided in Section III. In Section IV a range of linear combining based antenna array weight optimization schemes are invoked for deriving the decision variables, while the attainable performance is studied in Section V. Finally, we offer our conclusions in Section VI.

II. System Description

A. Transmitted Signal

In this subsection, the generalized MC DS-CDMA system of Fig. 1 [8], [9], [10] is reviewed. At the transmitter side, the binary data stream having a bit duration of T_b is serial-to-parallel (S/P) converted to U parallel sub-streams. The new bit duration

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of each sub-stream, which we refer to as the symbol duration, becomes $T_s = UT_b$. After S/P conversion, each substream is spread using an N-chip DS spreading sequence waveform $c_k(t)$. Then, the DS spread signal of the uth sub-stream, where $u = 1, 2, \ldots, U$, simultaneously modulates a group of parallel subcarrier frequencies $\{f_{u1}, f_{u2}, \ldots, f_{uV}\}$ using Binary Phase Shift Keying (BPSK). A total of UV number of subcarriers are required in the MC DS-CDMA system considered and the UV number of subcarrier signals are superimposed on each other in order to form the complex modulated signal. Therefore, the transmitted signal of user k can be expressed as

$$s_k(t) = \sum_{u=1}^{U}\sum_{v=1}^{V}\sqrt{\frac{2P}{V}}b_{ku}(t)c_k(t)\cos(2\pi f_{uv}t + \phi_{k,uv}), \quad (1)$$

where P/V represents the transmitted power of each subcarrier and P is the transmitted power corresponding to each bit. Furthermore, $\{b_{ku}(t)\}$, $c_k(t)$, $\{f_{uv}\}$ and $\{\phi_{k,uv}\}$ represent the subcarrier data streams, the kth user's DS spreading waveform, the subcarrier frequency set and the phase angles introduced in the carrier modulation process.

B. Receiver Model

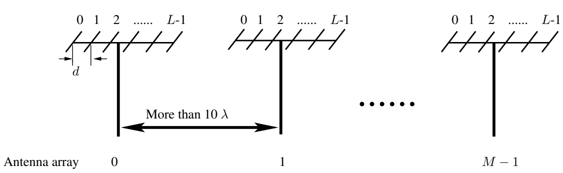

Fig. 2. Multiple antenna configuration used in the generalized MC DS-CDMA system considered.

We assume that at a base-station (BS) there are M number of receiver antenna arrays, as shown in Fig.2, which are located sufficiently far apart so that the corresponding received MC DS-CDMA signals experience independent fading, when they reach the different antennas. Each of the M antenna arrays consists of L number of elements separated by a distance of d, which is usually half a wavelength. For simplicity, we assume that there is no angle spread, then the Spatio-Temporal Channel Impulse Response (ST-CIR) $h_{uv,ml}^{(k)}$ between the uvth subcarrier of the kth user and the lth array element of the mth antenna can be expressed as

$$\begin{aligned} h_{uv,ml}^{(k)} &= \alpha_{uv,ml}^{(k)}(t)\exp(j[2\pi\frac{d}{\lambda}l\sin(\psi_m^{(k)}])\delta(t-\tau_k), \quad (2)\\ & m = 0, 1, \ldots, M-1;\ l = 0, 1, \ldots, L-1;\\ & u = 1, 2, \ldots, U;\ v = 1, 2, \ldots, V;\\ & k = 1, 2, \ldots, K, \end{aligned}$$

where τ_k is the signal's delay, $\alpha_{uv,ml}^{(k)}(t)$ is the Rayleigh faded envelope's amplitude, d is the inter-element spacing of the antenna array, λ is the wavelength and $\psi_m^{(k)}$ is the average Direction-Of-Arrival (DOA).

We assume that K asynchronous MC DS-CDMA users are supported by the system, while perfect power control is assumed. Consequently, the received baseband equivalent K-user signal vector at the antenna arrays' output can be expressed as

$$\mathbf{r}(t) = \sum_{k=1}^{K}\mathbf{r}_k(t) + \mathbf{n}(t), \quad (3)$$

where $\mathbf{n}(t)$ is the $(ML \times 1)$-dimensional additive white Gaussian noise (AWGN) vector having a zero mean and covariance of $2N_0\mathbf{I}\cdot\delta(t_1 - t_2)$, with the superscript H denoting the conjugate transpose operation. In (3) $\mathbf{r}_k(t)$ represents the kth user's received signal arriving over the spatio-temporal channels considered, which can be expressed as $\mathbf{r}_k(t) = s_k(t) \otimes \tilde{\mathbf{h}}_{uv}^{(k)}(t)$, where the $ML \times 1$-dimensional vector of

$$\begin{aligned} \tilde{\mathbf{h}}_{uv}^{(k)}(t) &= [(\mathbf{h}_{uv,0}^{(k)}(t))^T, (\mathbf{h}_{uv,1}^{(k)}(t))^T, \ldots, (\mathbf{h}_{uv,(M-1)}^{(k)}(t))^T]^T \\ &= \tilde{\mathbf{a}}_{uv}^{(k)}(t)\delta(t-\tau_k). \quad (4) \end{aligned}$$

represents the ST-CIRs corresponding to the entire set of ML elements of the M antenna arrays.

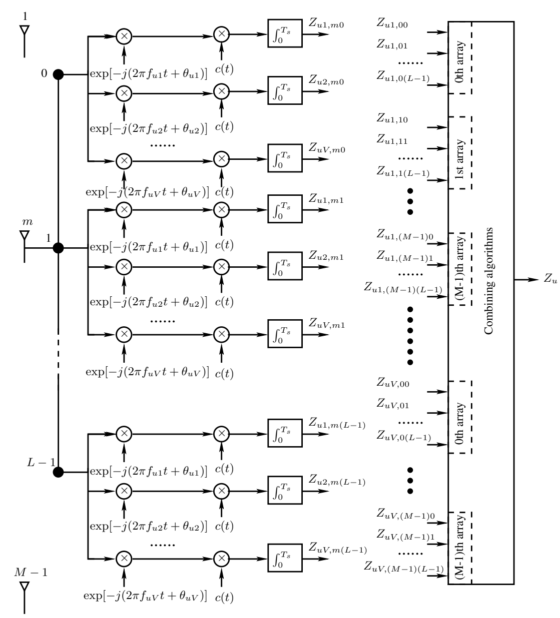

Fig. 3. Receiver block diagram of the generalized MC DS-CDMA system considered. The receiver employs beamforming, receiver diversity combining and multicarrier-spreading-assisted frequency-selective diversity combining.

The block diagram of the receiver designed for detecting the information arriving from the reference user is shown in Fig.3, where the superscript and subscript denoting the reference user of $k = 1$ have been omitted for notational convenience. We assume that the receiver is capable of acquiring perfect time-domain synchronization. The attenuations and phases of the ST CIR taps are assumed to be perfectly estimated. Furthermore, in this paper we assume that the DOA of the reference user's signal is also known at the base-station. Returning to Fig.3, after multicarrier demodulation and DS despreading, the antenna array outputs corresponding to the 0th bit $b_u[0]$ and the vth subcarrier can be expressed as

$$\mathbf{z}_{uv} = \int_0^{T_s} \mathbf{r}(t)c(t)\exp\left(-j[2\pi f_{uv}t + \theta_{uv}]\right) dt, \quad v = 1, 2, \ldots, V, \tag{5}$$

where we assumed that the reference signal's transmission delay was $\tau_1 = 0$ for the sake of simplicity.

III. Statistical Analysis

In this section we characterize the properties of the decision variable $\mathbf{z}_u = [\mathbf{z}_{u1}^T, \mathbf{z}_{u2}^T, \ldots, \mathbf{z}_{uV}^T]^T$, where $\mathbf{z}_{uv}$ is expressed in (5). Let us first derive the components of $\mathbf{z}_u$. We assume that the ST CIR taps given by $\tilde{\mathbf{a}}_{uv}^{(k)}$ of Equation (4) remain constant for a symbol duration of T_s. Then, it can be shown that $\mathbf{z}_{uv}$ expressed in terms of the vth subcarrier of b_u can be written as

$$\mathbf{z}_{uv} = \sqrt{\frac{2P}{V}} T_s [b_u[0]\tilde{\mathbf{a}}_{uv} + \tilde{\mathbf{n}}_{uv} + \underbrace{\sum_{u'=1}^{U}\sum_{v'=1}^{V}}_{v'\neq v, \text{ if } u'=u} \mathbf{i}_{u'v'}^{(s)} + \sum_{k=2}^{K} \mathbf{i}_{uv}^{(k)} + \sum_{k=2}^{K} \underbrace{\sum_{u'=1}^{U}\sum_{v'=1}^{V}}_{v'\neq v, \text{ if } u'=u} \mathbf{i}_{u'v'}^{(k)}], \tag{6}$$

where $b_u[0]\tilde{\mathbf{a}}_{uv}$ represents the desired array outputs, while $\tilde{\mathbf{n}}_{uv}$ is contributed by $\mathbf{n}(t)$ of (3), which is an $(ML\times 1)$-dimensional AWGN vector having zero mean and a covariance of $\frac{VN_0}{E_b}\mathbf{I}$. The term $\mathbf{i}_{u'v'}^{(s)}$ in (6) represents the self-interference contributed by the subcarrier indexed by u', v' of the reference signal. The multiuser interference (MUI) term $\mathbf{i}_{uv}^{(k)}$ in (6) is engendered by the subcarrier signal indexed by u and v of the kth interfering user. Finally, the MUI term $\mathbf{i}_{u'v'}^{(k)}$ in (6) is imposed by the subcarrier identified by u' and v' associated with the kth interfering user.

We have assumed that the MC DS-CDMA signal experiences independent fading, when it reaches different antennas. Hence, after ignoring the common factor of $\sqrt{2P/V}T_s$ in (6), the array outputs expressed in the context of bit $b_u[0]$ can be written as

$$\mathbf{z}_u = \tilde{\mathbf{a}}_u b_u[0] + \tilde{\mathbf{n}}_u + \mathbf{j}_u, \tag{7}$$

where $\tilde{\mathbf{a}}_u = [\mathbf{a}_{u1,0}, \cdot, \mathbf{a}_{u1,M-1}, \cdots, \mathbf{a}_{uV,0}, \cdot, \mathbf{a}_{uV,M-1}]^T$, and $\mathbf{a}_{uv,m}$ can be expressed as

$$\mathbf{a}_{uv,m} = \left[\alpha_{uv,m0}, \alpha_{uv,m1}\exp\left(j\left[2\pi\frac{d}{\lambda}\sin(\psi_m)\right]\right), \ldots, \alpha_{uv,m(L-1)}\exp\left(j\left[2\pi\frac{d}{\lambda}(L-1)\sin(\psi_m)\right]\right)\right]^T. \tag{8}$$

In (7) $\mathbf{z}_{uv,m}$ can be modelled as an independently distributed random vector having a common covariance matrix. Furthermore, $\tilde{\mathbf{a}}_u$, $\tilde{\mathbf{n}}_u$ and $\mathbf{j}_u$ in (7) are assumed to be the mutually independent random vectors.

The correlation matrix $\mathbf{R}_u$ of $\mathbf{z}_u$ can be expressed as

$$\mathbf{R}_u = E\left[\mathbf{z}_u\mathbf{z}_u^H\right] = \mathbf{R}_d + \mathbf{R}_n, \tag{9}$$

where $\mathbf{R}_d$ contains the autocorrelations of the desired signals hosted by the matrix $\tilde{\mathbf{a}}_u$ in (7), which can be expressed as

$$\begin{aligned}\mathbf{R}_d = \text{diag}\{&E\left[\mathbf{a}_{u1,0}\mathbf{a}_{u1,0}^H\right], \ldots, E\left[\mathbf{a}_{u1,M-1}\mathbf{a}_{u1,M-1}^H\right], \\ &E\left[\mathbf{a}_{u2,0}\mathbf{a}_{u2,0}^H\right], \ldots, E\left[\mathbf{a}_{u2,M-1}\mathbf{a}_{u2,M-1}^H\right], \\ &\cdots\cdots\cdots \\ &E\left[\mathbf{a}_{uV,0}\mathbf{a}_{uV,0}^H\right], \ldots, E\left[\mathbf{a}_{uV,M-1}\mathbf{a}_{uV,M-1}^H\right]\},\end{aligned} \tag{10}$$

where diag $\{\ldots\}$ represents a diagonal matrix. Assuming that the fading parameters $\{\alpha_{uv,ml}\}$ of (8) are independent of the index l, then we have

$$E\left[\mathbf{a}_{uv,m}\mathbf{a}_{uv,m}^H\right] = (\alpha_{uv,m})^2 E\left[\mathbf{d}_m\mathbf{d}_m^H\right], \tag{11}$$

where $\mathbf{d}_m = [1, \exp(j2\pi\frac{d}{\lambda}\sin(\psi_m)), \ldots, \exp(j2\pi\frac{d}{\lambda}(L-1)\sin(\psi_m))]^T$ is an L-length vector related to the DOA of the desired user's signal.

In (9) $\mathbf{R}_n$ represents the covariance matrix of the interference-plus-noise, which is given by

$$\mathbf{R}_n = E\left[(\tilde{\mathbf{n}}_u + \mathbf{j}_u)(\tilde{\mathbf{n}}_u + \mathbf{j}_u)^H\right] = \mathbf{R}_N + \mathbf{R}_J, \tag{12}$$

where we have $\mathbf{R}_N = \frac{VN_0}{E_b}\mathbf{I}$, while $\mathbf{R}_J$ represents the covariance matrix of the composite interference given by the sum of the self-interference and multiuser interference. Following the approach of [8], $\mathbf{R}_J$ can be expressed as

$$\mathbf{R}_J = \eta \cdot \sum_{k=2}^{K} \mathbf{R}_J^{(k)}, \tag{13}$$

where we have

$$\begin{aligned}\mathbf{R}_J^{(k)} = \text{diag}\{&E[\mathbf{a}_{u1,0}^{(k)}(\mathbf{a}_{u1,0}^{(k)})^H], \ldots, E[\mathbf{a}_{u1,M-1}^{(k)}(\mathbf{a}_{u1,M-1}^{(k)})^H], \\ &E[\mathbf{a}_{u2,0}^{(k)}(\mathbf{a}_{u2,0}^{(k)})^H], \ldots, E[\mathbf{a}_{u2,M-1}^{(k)}(\mathbf{a}_{u2,M-1}^{(k)})^H], \\ &\cdots\cdots\cdots \\ &E[\mathbf{a}_{uV,0}^{(k)}(\mathbf{a}_{uV,0}^{(k)})^H], \ldots, E[\mathbf{a}_{uV,M-1}^{(k)}(\mathbf{a}_{uV,M-1}^{(k)})^H]\},\end{aligned}$$

and $\eta = \frac{2}{3N} + 2(UV-1)\overline{I}_M$, $\overline{I}_M = \frac{1}{UV(UV-1)}\sum_{v=1}^{UV}\sum_{\substack{u=1\\u\neq v}}^{UV}\frac{N}{2\pi^2(u-v)^2\lambda^2}\left[1-\text{sinc}\left(\frac{2\pi(u-v)\lambda}{N}\right)\right]$, while $E\left[\mathbf{a}_{uv,m}^{(k)}(\mathbf{a}_{uv,m}^{(k)})^H\right]$ was given in (11), when the ST-CIR was independent of the index l.

Above, the received signal of the VML-dimensional vectors of the V subcarriers conveying replicas of the bits and ML array elements was jointly processed and the statistical properties of the antenna array's outputs $\mathbf{z}_u$ were characterized. Thus after deriving the VML-dimensional optimum combiner $\mathbf{w}$, the decision variable may be expressed as $Z_u = \mathbf{w}^H \mathbf{z}_u$. However, the optimum combiner may also be derived by separately processing the V number of received subcarrier signals, each of which is hosted by an $(ML \times 1)$-dimensional vector. The resultant correlation matrix $\mathbf{R}_{u,v}$ of $\mathbf{z}_{uv}$ may be expressed as

$$\mathbf{R}_{u,v} = E\left[\mathbf{z}_{uv}\mathbf{z}_{uv}^H\right] = \mathbf{R}_{d,v} + \mathbf{R}_{n,v}, \tag{14}$$

where $\mathbf{R}_{d,v} = \text{diag}\{E[\mathbf{a}_{uv,0}\mathbf{a}_{uv,0}^H], \ldots, E[\mathbf{a}_{uv,M-1}\mathbf{a}_{uv,M-1}^H]\}$. In (14) $\mathbf{R}_{n,v}$ is given by

$$\mathbf{R}_{n,v} = \mathbf{R}_{N,v} + \mathbf{R}_{J,v}, \tag{15}$$

where we have $\mathbf{R}_{N,v} = \frac{VN_0}{E_b}\mathbf{I}$, while $\mathbf{R}_{J,v}$ can be expressed as $\mathbf{R}_{J,v} = \eta \times \sum_{k=2}^{K} \text{diag}\{E[\mathbf{a}_{uv,0}^{(k)}(\mathbf{a}_{uv,0}^{(k)})^H], \ldots, E[\mathbf{a}_{uv,M-1}^{(k)}(\mathbf{a}_{uv,M-1}^{(k)})^H]\}$. Consequently, after deriving the ML-dimensional optimum combiner $\mathbf{w}_v$ for the vth subcarrier, the optimum combiner's output corresponding to the vth subcarrier is given by $Z_{uv} = \mathbf{w}_v^H \mathbf{z}_{uv}$. Finally, the decision variable Z_u can be obtained by combining the subcarrier signals Z_{uv} according to the MRC principle formulated as $Z_u = \sum_{v=1}^{V} Z_{uv}$.

IV. Combining Criteria for Space-Time MC DS-CDMA Signals

A. Minimum Variance Distortionless Response

The MVDR approach constitutes a well-known beamforming optimization criterion [11], [12], which is capable of providing the minimum-variance unbiased estimate of the transmitted signal. More explicity, this corresponds to minimizing the power of the interference-plus-noise at the array's output, provided that the desired signal is distortionless. In the optimum combiner using the MVDR criterion, the MVDR solution may be expressed as

$$\mathbf{w}_{MVDR} = \frac{\mathbf{R}_n^{-1}\tilde{\mathbf{a}}_u}{\tilde{\mathbf{a}}_u^H \mathbf{R}_n^{-1}\tilde{\mathbf{a}}_u}. \tag{16}$$

B. Maximum Signal-To-Interference-Plus-Noise Ratio

According to the MSINR criterion [11], [12], [13], the weight vector is optimized by maximizing the SINR at the output of the combiner. The optimum weight vector that maximizes the SINR at the output of the combiner is given by [11]:

$$\mathbf{w}_{MSINR} = \alpha \mathbf{R}_n^{-1}\tilde{\mathbf{a}}_u. \tag{17}$$

C. Minimum Mean-Square Error

The MMSE combiner [11], [12], [13] minimizes the mean-square error of $E[|e(\mathbf{w})|^2]$ between the combiner output $\mathbf{w}^H\mathbf{z}_u$ and the desired signal $b_u[0]$. The optimum weight vector based on the MMSE criterion may be expressed as

$$\mathbf{w}_{MMSE} = \alpha \mathbf{R}_u^{-1}\tilde{\mathbf{a}}_u, \tag{18}$$

where α is a constant. From [11], the optimum weight vector of the MMSE combiner may also be expressed as

$$\mathbf{w}_{MMSE} = \frac{\alpha}{1 + \sigma_1^2 \tilde{\mathbf{a}}_u^H \mathbf{R}_n^{-1}\tilde{\mathbf{a}}_u}\mathbf{R}_n^{-1}\tilde{\mathbf{a}}_u. \tag{19}$$

D. Minimum Power Distortionless Response

Finally, let us consider the combiner based on the MPDR criterion [11], which is closely related to the MVDR combiner of Subsection IV-A. According to the MPDR criterion, the optimum array weight vector may be expressed as [11]:

$$\mathbf{w}_{MPDR} = \frac{\mathbf{R}_u^{-1}\mathbf{a}_m}{\mathbf{a}_m^H \mathbf{R}_u^{-1}\mathbf{a}_m}, \tag{20}$$

where $\mathbf{a}_m$ represents the steering vector [11], while $\mathbf{R}_u$ is the auto-correlation matrix of the received signal vector.

V. Performance Results

In this section we quantify the performance of the generalized MC DS-CDMA system communicating over a single-path non-dispersive Rayleigh fading channel contaminated by AWGN employing 31-chip Gold codes as time-domain spreading sequences and 4-chip Walsh codes as the frequency-domain spreading codes. We have four subcarriers, as seen in Figures 4 and 5. Specifically, in Figure 4 we assume that we have $\Gamma_k = 1, i = 2, 3, \ldots, K$, where $\Gamma_k = \frac{\text{mean received signal power of the } k\text{th interferer}}{\text{mean received noise power}}$, and we use a 3×1-dimensional linear antenna array $(M = 1, L = 3)$. Four users are supported in this scenario. It transpires from Figure 4 that the performance of the joint subcarrier processing based optimum MSINR combiner is better than that of the joint subcarrier processing based optimum MMSE combiner. Furthermore, the performance of the individual subcarrier-based optimum MSINR combiner is better than that of the individual subcarrier-based optimum MMSE combiner. Hence the employment of the entire post-decorrelation signal's correlation matrix $\mathbf{R}_u$ seen in (9) may result in a more substantial effect on the array weight vector, than that of the interference-plus-noise correlation matrix $\mathbf{R}_n$ formulated in (12). Figure 4 also demonstrates that as expected, the performance of the joint subcarrier processing based optimum combiner is better, than that of the individual subcarrier-based optimum combiner. However, the matrix $\mathbf{R}_n$ of (12) processed by the joint subcarrier processing based optimum combiner is $(VML \times VML)$-dimensional, while the V number of matrices $\mathbf{R}_{n,v}$ in (15) processed by the individual subcarrier-based optimum combiner are $(ML \times ML)$-dimensional. Hence, the computational complexity imposed by the inversion of the matrix $\mathbf{R}_n$ of (12) on the joint subcarrier processing based optimum combiner is significantly higher, than that of the inversion of the matrix $\mathbf{R}_{n,v}$ of (15) carried out V times by the individual subcarrier-based optimum combiner. In Fig. 5 we support $K = 16$ users. In this scenario we consider three different antenna arrays. The first antenna array is a (1×6)-dimensional linear antenna array associated with $(M = 1, L = 6)$, having an element-spacing of $\lambda/2$. The second is a (6×1)-dimensional linear antenna array $(M = 6, L = 1)$, having an array-spacing of 10λ. The third is a (2×3)-dimensional array $(M = 2, L = 3)$, having an element-spacing of $\lambda/2$ and an array-spacing of 10λ. Figure 5 shows that when the spatial signals arriving at the elements of an antenna array are less correlated, the attainable spatial diversity gain becomes higher, hence the achievable performance improves. Figure 5 also demonstrates that when the correlation between the signals arriving at the different elements of an antenna array decreases, the performance difference between the MRC scheme and the optimum combiner also reduces. By contrast, there is no appreciable difference between the performance of the MRC scheme and the MSINR combiner. Furthermore, the system adopting a (2×3)-dimensional antenna array ($M = 2, L = 3$) achieves a mediocre performance. As seen for the ($M = 6, L = 1$) scenario, the performance of the MRC scheme and the MSINR combiner is similar also in this scenario. Finally, the performance of the system using a (1×6)-dimensional linear antenna array ($M = 1, L = 6$) is the worst and the associated performance difference between the MRC scheme and the MSINR combiner is the highest, although the difference is still less than 2dB.

VI. Conclusions

In this paper, we have proposed four different combiners based on the MVDR, MSINR, MMSE and MPDR array element optimization criteria, which are invoked for a generalized MC DS-CDMA system. Several different antenna array models have been employed in our simulations. We conclude from the simulation results and the accompanying analysis that the performance of the different combiners is different. When the number of users increases, the correlation matrix $\mathbf{R}_n$ of (12) or $\mathbf{R}_u$ of (9) becomes reminiscent of an identity matrix. Accordingly, the weighting process becomes reminiscent of the action of the classic MRC scheme. Having investigated a range of different antenna array models, in harmony with our expectations we found that when the spatial signals arriving at the different elements of the antenna array become less correlated, the spatial diversity gain becomes higher, hence the achievable performance improves. By contrast, when the correlation of the signals arriving at the different elements of the antenna array decreases, the performance difference between the MRC scheme and the optimum combiners reduces, as anticipated.

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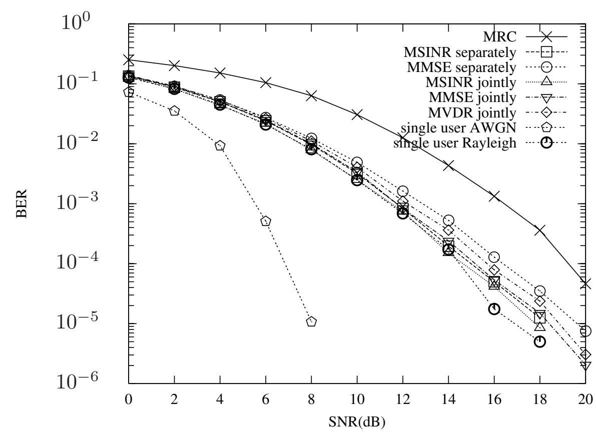

Fig. 4. BER versus SNR performance of the uplink of a generalized MC DS-CDMA wireless system supporting $K = 4$ users and employing 31-chip Gold codes as time-domain spreading sequences and 4-chip Walsh codes as frequency-domain spreading sequences, using a (3×1)-dimensional antenna array ($M = 1, L = 3$), as well as different optimum combiners based on the combining schemes of Section IV. The four users employ the same Gold code as their time-domain spreading sequence, while using different Walsh codes as their frequency-domain spreading sequences.

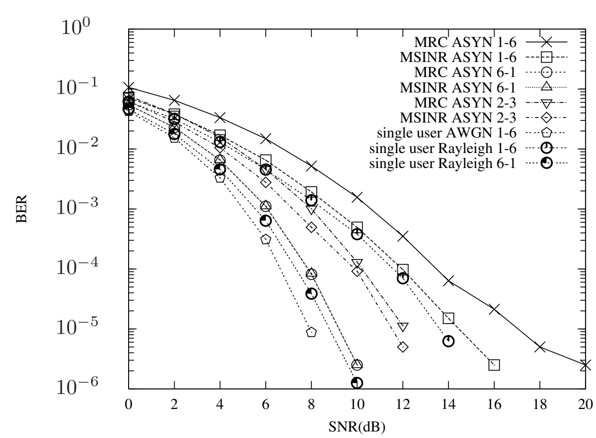

Fig. 5. BER versus SNR performance of the uplink of a generalized MC DS-CDMA wireless system supporting $K = 16$ users and employing 31-chip Gold codes as time-domain spreading sequences and 4-chip Walsh codes as frequency-domain spreading sequences, using different antenna arrays, different optimum combiners based on the combining schemes of section IV.