BLIND MULTI-USER COMBINING AT THE BASE STATION FOR ASYNCHRONOUS CDMA SYSTEMS[1]

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ABSTRACT

This paper studies the potential benefits of antenna arrays in cellular CDMA communications and proposes a powerful scheme to undertake the array processing at the base station in CDMA mobile systems.

The proposed technique exploits the temporal structure of CDMA signals. The necessary information is extracted directly from the received signals, thus no training signal or "a priory" spatial information is required. Making use of the overall information present at the receiver after the despreading, the technique here presented, jointly with a Singular Value Decomposition (SVD), estimates the generalized steering vector for each user. Then, an optimal combining of the signal received at each antenna is undertaken, achieving substantial robustness against fading and near-far problems.

1. INTRODUCTION

In the mobile environment, the emitter or/and receiver motion along with the multipath propagation results in multiplicative (fading) and/or dispersive distortion. The dispersive distortion holds any time the spread bandwidth exceeds the coherence bandwidth of the channel, that is, when the channel appears frequency-selective to the spread spectrum signal. The use of spread spectrum can be useful in combating this kind of degradation, since Direct Sequence (DS) waveforms can be used to either reject multipath returns that fall outside of the correlation interval of the spreading waveform, or enhance overall performance by diversity combining multipath returns in a RAKE receiver. The temporal diversity introduced by the RAKE receiver may be useful in the presence of deep fades. Nevertheless, when there is a single fading path (frequency non-selective channels), there is no means of diversity to overcome fading. In such a situation, the use of an antenna array may reduce significantly the probability of error [1]. On the other hand, in the presence of several resolvable multipaths, the wide band array scheme may combine the temporal diversity of the RAKE receiver and the spatial diversity of the narrow band array [2].

The received signal is also corrupted by noise and possible interferences. The ability of spread spectrum signals to combat interferences from other users results in the potential of higher performance and system capacity. Nevertheless, CDMA systems are strongly affected by the "near-far" problem : if the processing gain is not high enough, the higher power (near) sources overwhelm the lower power sources and so system capacity will drop spectacularly. Despite of the name, the difference among the powers of different users depends not only on the distances to the base station but on the Rayleigh fading and changes in shadowing too. A solution to the near-far problem is the use of power control. Another alternative is the use of an antenna array. The antenna array adds the wanted signals vectors in a maximum ratio sense, while interferences are weighted so that their resultant is in a permanent deep fade. Ideally, if the array works properly and total cancellation occurs, system performance should be even better than in the case of having equal power reception. Thus, combining spread spectrum signals with adaptive array beamforming results in a more powerful system that rejects interferences and combats the multipath. Consequently the performance and capacity of the system may be increased.

Some previous work has been carried out using antenna arrays for synchronous CDMA systems in stationary mobile frequency non-selective channels to increase system capacity [3]. Actually, for a single channel multiuser receiver the decorrelating detector completely eliminates the multiple access interference channel in a synchronous system by means of a linear combination of the despreaded signals [4]. Then, the only reason to use the antenna array would be only to gain robustness against fading. Nevertheless, for the asynchronous case the linear combination of the despreaded signals is not enough to eliminate the multiple access interference. The required detector to eliminate completely the multiple access interference presents excessive complexity [5],[6]. Working with antenna arrays and with the spatial correlation matrix of the despreaded signals, we find that a similar relationship to that shown in [4] for the received CDMA signals exists for their corresponding spatial correlation matrices for both synchronous and asynchronous case. Exploiting this property, we apply the same procedure to carry out the beamforming for both cases. This is a very important feature of this method. There is no additional complexity in using this method for asynchronous systems. Although for simplicity the method in this work is presented for the synchronous case first and extended to the asynchronous case later, the method is much more interesting for the later case. There are several reasons for it. The first one is that this method is conceived for the base station receiver and the up-link (mobile to base station) is clearly asynchronous. Another reason to consider is that an asynchronous CDMA system is vulnerable to the "near-far" effect, due to the non-zero cross-correlation of the modulating signals assigned to individual users in CDMA. When having signals with very dissimilar power, the lack of orthogonality among the modulating signals, due to their own nature or provocated by the asynchronism between users, may unability the whole system. In this case, the use of an array to null those interfering signals with higher power can improve

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significantly the system performance in terms of bit error rate. Using the method here proposed, the desired steering vector may be estimated even with the near-far problem.

Our method has another important property to remark. The beamforming is carried out exploiting the particular properties of the CDMA signals. Usually, when the location of the emitter is unknown, as is the case in mobile communications the beamforming is normally realized by the use of a temporal reference instead of spatial reference as is the angle of arrival of the desired source. Nevertheless, algorithms to control the array reception pattern based on temporal references decrease the efficiency of the communication systems. The technique proposed here exploits the temporal structure of the CDMA signal but let extract the necessary information directly from the received signals.

The paper is organized as follows. In section 2, we describe the signal model. In section 3, the proposed method is presented for both the synchronous and asynchronous case. Finally, in section 4 simulation results and conclusions are drawn.

2. SIGNAL MODEL

The system under consideration is a K-user asynchronous DS-CDMA system using BPSK modulation and operating over a frequency non-selective fading channel.

The received analytical signal may be written

$$r(t)=\sum_{k=1}^{K}\sqrt{2p_k}\,\alpha_k(t)s_k(t-\tau_k)e^{j\omega_c t}+n(t) \qquad (1)$$

where n(t) is white Gaussian noise with variance σ^2 and α_k, τ_k and p_k are respectively the complex channel gain, the propagation delay and the transmitted power. All for the k-th user.

The baseband signal for the k-th user

$$s_k(t)=\sum_m d_k(m)b_k(t-mT) \qquad (2)$$

the data stream $d_k(m)\in\{+1,-1\}$ is pulse amplitude modulated by a period of the code waveform $b_k(t)$, and $b_k(t)=0$ for $t\notin[0,T)$ with T the bit time.

From eq.(1), the baseband received signal for the multichannel case:

$$\boldsymbol{x}(t)=\sum_{k=1}^{K}\sqrt{p_k}\,s_k(t-t_k)\boldsymbol{a}_k(t)+\boldsymbol{n}(t) \qquad (3)$$

where $\mathbf{a}_k$ is the generalized steering vector for the k-th user with dimension equal to the number of sensors (N).

Here, it is assumed that the fading processes are wide-sense stationary and slowly varying compared to the symbol time, i.e.

$$\boldsymbol{a}_k(t)\cong\boldsymbol{a}_k(t+T) \qquad (4)$$

The noise is assumed to be uncorrelated among different sensors and with equal power for all of them.

3. DESCRIPTION OF THE PROPOSED METHOD

The received signal vector is fed to a bank of matched filters, each one matched to each one of the active K users. The outputs are sampled every T seconds, each synchronized to the corresponding source emission. For each user, a different array covariance matrix is computed after despreading the signal of each sensor. By means of a linear combination of these matrixes a new set of rank-one matrixes is obtained, each one for each user. The eigenvector associated with the only non-zero eigenvalue for every matrix is exactly the generalized steering vector of the corresponding user. With this information , the optimum weight vector may be calculated for every user. This is the basical idea of the method we proposed, which is described in more detail in next subsections for both the synchronous and asynchronous case.

3.1 Synchronous case

Attention here will be focused first at the symbol-synchronous channel case, where the symbol epochs of all users coincide at the receiver. Thus, $\tau_1=\tau_2=...=\tau_N=\tau$. Without loss of generality it can be assumed $\tau=0$.

Let be r_{lk} the cross correlation at zero offset, in time of the two modulating signals $b_l(t)$ and $b_k(t)$.

$$r_{lk}=\int_0^T b_l(t)b_k(t)dt \qquad (5)$$

The output of the matched filter for $s_k(t)$:

$$z_k(n)=\int_{nT}^{(n+1)T}\boldsymbol{x}(t)b_k(t-nT)dt \qquad (6)$$

$$z_k(n)=\sqrt{p_k}\,d_k(n)\boldsymbol{a}_k+\sum_{l\neq k}r_{lk}\sqrt{p_l}\,d_l(n)\boldsymbol{a}_l+n_k(n) \qquad (7)$$

The spatial covariance matrix for the output of the k-th matched filter is

$$\boldsymbol{R}_{zz,k}=p_k\boldsymbol{a}_k\boldsymbol{a}_k^H+\sum_{l\neq k}r_{lk}^2p_l\boldsymbol{a}_l\boldsymbol{a}_l^H+\sigma_k^2\boldsymbol{I} \qquad (8)$$

I is the identity matrix of dimension NxN, with N the number of sensors and σ_k^2 the noise variance at the output of the k-th matched filter.

Defining matrix **S** as follows,

$$\boldsymbol{S}=\begin{bmatrix} 1 & r_{12}^2 & \cdots & r_{1K}^2 \\ r_{12}^2 & 1 & \cdots & r_{2K}^2 \\ \vdots & & \ddots & \\ r_{1K}^2 & r_{2K}^2 & \cdots & 1 \end{bmatrix} \qquad (9)$$

we can write:

$$\begin{bmatrix} \boldsymbol{R}_{zz,1}-\sigma_1^2\boldsymbol{I} \\ \boldsymbol{R}_{zz,2}-\sigma_2^2\boldsymbol{I} \\ \vdots \\ \boldsymbol{R}_{zz,K}-\sigma_K^2\boldsymbol{I} \end{bmatrix}=(\boldsymbol{S}\otimes\boldsymbol{I})\begin{bmatrix} p_1\boldsymbol{a}_1\boldsymbol{a}_1^H \\ p_2\boldsymbol{a}_2\boldsymbol{a}_2^H \\ \vdots \\ p_K\boldsymbol{a}_K\boldsymbol{a}_K^H \end{bmatrix} \qquad (10)$$

where the symbol $\otimes$ denotes the Kronecker product.

Let be

$$\boldsymbol{A}=\begin{bmatrix} p_1\boldsymbol{a}_1\boldsymbol{a}_1^H \\ p_2\boldsymbol{a}_2\boldsymbol{a}_2^H \\ \vdots \\ p_K\boldsymbol{a}_K\boldsymbol{a}_K^H \end{bmatrix} \qquad (11)$$

$$\boldsymbol{R}_{zz}=\left[\boldsymbol{R}_{zz,1}^T\ \boldsymbol{R}_{zz,2}^T\ \cdots\ \boldsymbol{R}_{zz,K}^T\right]^T \qquad (12)$$

$$\boldsymbol{N}=\left[\sigma_1^2\ \sigma_2^2\ \ldots\ \sigma_K^2\right]^T\otimes\boldsymbol{I} \qquad (13)$$

Assuming the noise variance at the output of the matched filters is equal for all of them, matrix **N** can be written as:

$$\boldsymbol{N}=\sigma^2\boldsymbol{1}\otimes\boldsymbol{I},\qquad \sigma^2=\sigma_1^2=\sigma_2^2=\ldots=\sigma_K^2 \qquad (14)$$

where **1** is an all ones column vector with dimension K.

From eq.(11) to (14), eq.(10) may be written in a more compact way as:

$$R_{zz} - \sigma^2 I \otimes I = (S \otimes I)A \tag{15}$$

Using the properties of the Kronecker product and operating with previous expression we arrive:

$$M = (S^{-1} \otimes I)R_{zz} = A + \sigma^2 u \otimes I \tag{16}$$

where

$$u = S^{-1} 1 \tag{17}$$

Considerer **M** partitioned into NxN blocks as follows:

$$M = \begin{bmatrix} M_1 \\ M_2 \\ \vdots \\ M_K \end{bmatrix} \tag{18}$$

The part of the matrix **M** corresponding to the k-th user is

$$M_k = p_k a_k a_k^H + u_k \sigma^2 I \tag{19}$$

where u_k is the k-th element of vector **u**.

By means of a linear operation over the global "spatial-covariance matrix" R_{zz} a set of matrices M_k is obtained. These matrices have N-1 eigenvalues equal to $u_k\sigma^2$ and one different eigenvalue equal to $p_k + u_k\sigma^2 >> u_k\sigma^2$. The important issue to remark is that the eigenvector associated with this eigenvalue is exactly the GSV (generalized steering vector) of the k-th user.

Once we have the estimations v_k of the generalized steering vectors for all the active users, different beamformers may be computed.

In this communication we have considered only two beamformers. The first one (eq. 20) maximizes the signal to noise plus interference ratio (SNIR) at the array output using the information of the desired user only.

$$w_{k,1} = \frac{R_{zz,k}^{-1} v_k}{v_k{}^H R_{zz,k}^{-1} v_k} \tag{20}$$

The second one (eq. 21) set nulls in the direction of the interferences and gain equal 1 in the direction of the desired signal, minimizing the white noise at the output with these constraints.

$$w_{k,2} = V(V^H V)f \tag{21}$$

$$\text{with } V = [v_1 \ldots v_k \ldots v_K] \text{ and } f = [0 \ldots 1 \ldots 0]^T \tag{22}$$

3.2 Asynchronous case

In this section, the general asynchronous case eq.(1) is considered. For clarity in the exposition, the formulation is carried out just for two users, i.e. K=2.

$$z_k(n) = \int_{nT}^{(n+1)T} x(t) b_k(t - nT - \tau_k) dt \tag{23}$$

It can be assumed without loss of generality $\tau_1 < \tau_2$. Then, it can be shown [5]

$$\begin{aligned} z_1(n) = {} & \sqrt{p_1} d_1(n) a_1 + \\ & + \sqrt{p_2}\left[r_{12,a} d_2(n-1) + r_{12,b} d_2(n)\right] a_2 + n_1(n) \end{aligned} \tag{24}$$

$$\begin{aligned} z_2(n) = {} & \sqrt{p_1}\left[r_{12,a} d_1(n+1) + r_{12,b} d_1(n)\right] a_1 + \\ & + \sqrt{p_2} d_2(n) a_2 + n_2(n) \end{aligned} \tag{25}$$

$$\text{with } \quad r_{12,a} = \int_0^T b_1(t) b_2(t + \tau_1 - \tau_2) dt \tag{26}$$

$$r_{12,b} = \int_0^T b_1(t) b_2(t + T + \tau_1 - \tau_2) dt \tag{27}$$

Assuming that different symbols are uncorrelated

$$R_{zz,1} = p_1 a_1 a_1^H + p_2 a_2 a_2^H (r_{12,a}^2 + r_{12,b}^2) + \sigma_1^2 I \tag{28}$$

$$R_{zz,2} = p_1 a_1 a_1^H (r_{12,a}^2 + r_{12,b}^2) + p_2 a_2 a_2^H + \sigma_2^2 I \tag{29}$$

$$S = \begin{bmatrix} 1 & r_{12,a}^2 + r_{12,b}^2 & \cdots & r_{1K,a}^2 + r_{1K,b}^2 \\ r_{12,a}^2 + r_{12,b}^2 & 1 & \cdots & r_{2K,a}^2 + r_{2K,b}^2 \\ \vdots & & \ddots & \\ r_{1K,a}^2 + r_{1K,b}^2 & r_{2K,a}^2 + r_{2K,b}^2 & \cdots & 1 \end{bmatrix} \tag{30}$$

Eq. 24 to 25 may be generalized for any number of users. Then defining matrix **S** as in eq. 30 the procedure described in previous section may be applied to obtain the generalized steering vector for each user.

4. SIMULATIONS

In order to determine the performance of the previously described receiver, a CDMA system with three active users (K=3) was simulated. The modulating signals in this system are maximal length sequences with lenght L=31. We considerer a $\lambda/2$ array with 6 sensors. In a beginning we considerer for the simulations a non-multipath scenario with different situations relative to synchronism and near-far ratio. The reason to considerer first a non-multipath scenario is because is easier to extract conclusions from the array beam pattern having only one direction of arrival. The direction of arrival of the signals are 38°, 10° and 45° for the first, second and third user respectively. The power considered for each one are 0, 15 and 10 dB respectively. The relative propagation delays are (randomly) set to $\tau_1 = 0$, $\tau_2 = 0.4286T_c$, $\tau_3 = 0.7143T_c$, with T_c the chip time of the maximal length sequence, i.e. T/31. The considered noise power at the input of one single sensor is 0 dB (before the despreading). The processing gain of the code with respect to interferences is reduced from $r_{11}^2/r_{12}^2 = r_{11}^2/r_{13}^2 = 31^2$ (which is the processing gain in a synchronous system for L=31) to $r_{11}^2/(r_{12,a}^2 + r_{12,b}^2) = 31^2/151.1224$ and $r_{11}^2/(r_{13,a}^2 + r_{13,b}^2) = 31^2/408.6735$ due to the asynchronism between users for the time delays considered.

The method proposed in section 3 requires knowledge of the data covariance matrix at the output of the matched filters. In practice, it is impossible to dispose of the spatial autocorrelation matrix because it is defined as the expected value over the realizations. However, assuming ergodicity of the involved signals it is possible to estimate this matrix from the data that are being receiving by means of the covariance method, that is, the temporal averaging of the despreaded signal vector. Using a block size of 100 symbols to estimate the correlation matrix, the beamformers for the first user are:

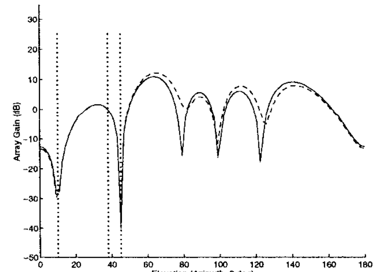

Fig. 1. Radiation pattern. (Dashed line: First beamformer, Solid line: Second beamformer).

The estimation of the steering vectors is sensitive to the near-far ratio. As we are using an estimation of the correlation matrices instead the real ones, the estimation of the steering vectors is degraded. Despite of this, a clear improvement may be achieved using the antenna array with respect to the classical receiver. In order to illustrate the performance and sensibility of our scheme to the near-far ratio, figures 2 and 3 plots the SNIR (signal to noise plus interference ratio) at the output of the array versus the near-far ratio or ISR (interference to signal ratio) when having two users. The time delay are 0 and $0.4286T_c$ respectively. A frequency non-selective Rayleigh fading channel has been considered. The fading rate is determined by the carrier frequency that is 1.9 GHz, the velocity of the mobile which has been assumed 1 m/s (walking speed) and the transmission rate (1152 Kbps). We are not considering here frequency selective fading but just flat fading, which affects only to the signal amplitude. Thus, the relative difference in power between both users depends on the different transmission power (which is the one represented in axis x) and by the random fading channels which affect to both users in a different way. Two different situations of noise have been considered (0 and -15 dB with respect to the desired signal before the despreading). The results, obtained averaging 50 independent runs with random channels and noise, show that, in constrat to the classical one-sensor receiver, the use of an antenna array allows the full rejection of the interference when an exact knowledge of the correlation matrix (or at least a good one) is available (Fig. 2). The performance is quite similar for the two studied beamformers (eq. 20-21). In this case the final SNIR is the input SNR augmented by the array gain independently of the near-far ratio.

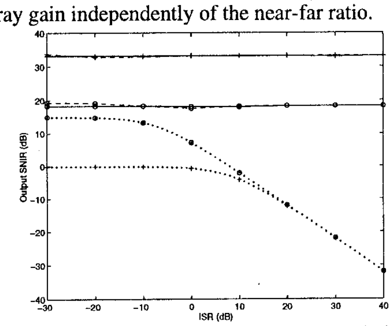

Fig. 2: Array gain vs near-far ratio (Exact correlation matrix). First beamformer (dashed line) and second beamformer (solid line) with Rayleigh fading channel. Classical receiver with non-fading channel (dotted line). SNR=0 dB (-o-) and SNR=15 dB(-+-) before the despreading.

Nevertheless, the first beamformer fails for high near-far ratios when an accurate estimation of the correlation matrix is not available. This can be observed in fig. 3 which shows the case of using the estimated correlation matrix using a block size of 100 symbols. This beamformer presents the very well known drawback of the beamformers operating only with the direction of arrival of the desired user. If this direction is not correctly estimated, the pointing error causes signal cancellation since the array always tries to minimize the overall power at the output including the desired signal. As it can be observed the degradation depends on the near-far ratio. Nevertheless, there is also some improvement with respect to the classical receiver. Below a given value of the interference power, the performance is near identical to the case of using the exact correlation matrix and quite better than for the classical receiver.

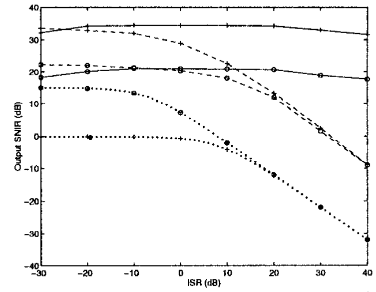

Fig. 3.: Array gain vs near-far ratio (Estimated correlation matrix). First beamformer (dashed line) and second beamformer (solid line) with Rayleigh fading channel. Classical receiver with non-fading channel (dotted line). SNR=0 dB (-o-) and SNR=15 dB(-+-) before despreading

Obviously, errors in the estimations of the correlation matrices affect also to the second beamformer, but these errors are not so critical as for the first one. As we can see in fig. 3 the margin of near-far ratio wherein the second approach can be work is excellent. We can observe also how much better is the performance of this detector in front of the performance of the classical receiver. We must take into account that the code processing gain is reduced from $r_{11}^2/r_{12}^2=31^2$ to $r_{11}^2/(r_{12,a}^2+r_{12,b}^2)=31^2/151.1224$ due to the asynchronism between users. The processing gain of the code with respect to noise is also augmented by the use of the antenna array. In the absence of interferences the use of the array implies an increase of $10\log_{10}N$ dBs over the gain of $10\log_{10}L$ dBs due to the code. N is the number of array elements and L the lenght of the code.

In the case of frequency selective fading, i.e. dispersive distortion along with the multiplicative distortion (fading) the method may be employed in conjunction with the wideband array scheme to isolate the resolved multipaths from one another via beamforming. Once the generalized steering vectors of the resolved paths are known, the spatial and temporal versions of each signal may be optimally combined after the alignment of their respective delays as in a RAKE receiver. In this case the theoretical gain achieved by the detector is given by the number of array elements and the power of the isolated paths optimally combined.

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