BLIND CHANNEL IDENTIFICATION AND PROJECTION RECEIVER DETERMINATION FOR MULTICODE AND MULTIRATE SITUATIONS IN DS-CDMA SYSTEMS

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ABSTRACT

We consider multicode and multirate transmission scenarios in a DS-CDMA system operating in an asynchronous fashion in a multipath environment. Oversampling w.r.t. the chip rate is applied to the cyclostationary received signal and multisensor reception is considered, leading to a linear multichannel model. Channels for different users are considered to be finite-impulse response (FIR) and of possibly different lengths, depending upon their processing gains. We consider an individualized linear MMSE-ZF or projection receiver for a given user, exploiting its spreading sequence and timing information. In the multicode case, each user is considered to use several spreading codes in order to transmit at a higher rate. Considering different code sequences to be issued from different virtual users, the propagation channel impulse responses of all these users are the same. However, the total channel impulse response which includes spreading sequences is different for all users. On the other hand, in the multirate case, a periodically varying set of periodic spreading codes spread successive symbols of a certain user. Symbols spread by different codes can therefore be considered to be issued from different virtual users. The problem therefore boils down to classical multiuser detection with time-invariant interference canceling filters for each virtual user. A blind channel estimate is also obtainable through Capon's method (first used by Tsatsanis) as a by-product of the MMSE-ZF receiver algorithm.

1. INTRODUCTION

Multicode transmission is an alternative to multirate transmission for achieving high data rates in third generation wireless networks based upon DS-CDMA. The former technique employs several channelization codes from a given set of spreading sequences of the same length to transmit data. Hence the input symbol stream at a high rate is demultiplexed and is spread by different spreading sequences to achieve greater capacity through parallel transmission. Multirate transmission, on the other hand, achieves faster data rates by sacrificing the processing gain. Each symbol is thus spread by a shorter code. The choice between the two modes of transmission is essentially open and depends on several parameters like the dynamic range of the receiver power amplifier, channel delay spread, number of actual users in the system, and the receiver algorithm. One advantage of multicode communications is that in large delay spreads, the intersymbol interference (ISI) is still small since the delay spread, even though significant in some cases, is much shorter than the symbol duration.

The purpose of this paper is to give improved insight into the multicode and multirate problems from a signal processing standpoint while presenting the blind MMSE-ZF [1] or projection receiver [2], taking into account the very particular structure of the problem in terms of the synchrony of the virtual multicode sub-users and the fact that the propagation channel seen by all is the same. It is further shown that Capon's method can be employed to blindly estimate the channel impulse response for the user of interest.

The receiver presented in this paper is a per user (decentralized) receiver in the sense that its estimation requires the knowledge of the desired user's spreading sequence and timing information along with the second order statistics of the received signal.

2. CONVERTING MULTIRATE TO MULTICODE

Fig. 1 shows the multirate scenario and depicts the discretized signal model for a ${u_k} = 2$ times faster user in the system ($P_k = P/2$), with $P$ the basic processing gain of the system. $h_k$ represents the discrete time chip-rate channel for the $k$th user, and $T(h_k)$ is the corresponding channel convolution matrix. The overall system therefore depicts spreading of successively symbols $a_k(n)$ by spreading sequences $e_k$, which are later passed through $h_k$, the discrete-time chip rate channel. It can be seen that for a user transmitting $u_k$ times faster than the slowest rate, the block diagonal spreading matrices $T(e_k)$ have periodically varying (with period $u_k$) vector elements on the diagonal. Due to the i.i.d. nature of the input data sequence, $a_k(n)$, this user can be viewed as $u_k$ cyclostationary users with modified spreading sequences shown in the figure with zeros padded either at the top or at the bottom.

This representation of faster users in a multirate system motivates the design of periodically time variant filters or yet better,
independent parallel receivers for the successive symbols of a certain faster user. The advantage of the split-up approach is that the estimation of these filters is independent of each other, exactly as if they were independent users. If the system has $K$ concurrent users transmitting at different rates, then let us denote by $K$, where, $K = \sum_{k=1}^{K} u_k$, the modified number of the basic rate users in the system. In the following, we shall consider this slowest rate representation of a multitrate system with $K$ effective users, and a processing gain, $P$, and concentrate on the detection of data symbols for the $k$th user, which might be a sub-user of a higher rate.

It is clear, therefore, that the case of multitrate communications can be cast into a multicode framework by considering the $u_k$ sub-users of a high rate user $k$ to be multicode users with modified spreading sequences (appended by zeros to keep a common processing gain, $P$). We shall henceforth consider a multirate description of the model and emphasize on its properties.

3. GENERAL MULTIUSER DATA MODEL

Fig. 2 shows the baseband signal model. The $K$ users are assumed to transmit linearly modulated signals over a linear multipath channel with additive Gaussian noise, $K$ is the number of virtual users. Any multitrate user, $k$, using $u_k$ spreading sequences will simply contribute $u_k$ virtual users to $K$. It is assumed that the receiver employs $M$ sensors to receive the mixture of signals from all users. The receiver front-end is an anti-aliasing low-pass filter. The continuous-time signal received at the $m$th sensor can be written in baseband notation as

$$y_m(t) = \sum_{k=1}^{K} \sum_{n} a_k(n) g_{k,m}(t-nT) + v_m(t), \quad (1)$$

where the $a_k(n)$ are the transmitted symbols from user $k$, $T$ is the common symbol period, $g_{k,m}(t)$ is the overall channel impulse response (including the spreading sequence, and the transmit and receive filters) for the $k$th user's signal at the $m$th sensor, and $\{v_m(t)\}$ is the complex circularly symmetric AWGN with power spectral density $N_0$. Assuming the $\{a_k(n)\}$ and $\{v_m(t)\}$ to be jointly wide-sense stationary, the process $\{y_m(t)\}$ is wide-sense cyclostationary with period $T$. The overall channel impulse response $g_{k,m}(t)$, is the convolution of the spreading code $c_k$ and $h_{k,m}(t)$ itself, the convolution of the chip pulse shape, the receiver filter, and the actual channel representing the multipath environment. This can be expressed as

$$g_{k,m}(t) = \sum_{p=0}^{P-1} c_k(p) h_{k,m}(t-pT_c), \quad (2)$$

where $T_c$ is the chip duration. The symbol and chip periods are related through the processing gain/spreading factor $P$: $T = PT_c$.

S/P in fig. 2 denotes serial-to-parallel conversion (vectorization) with down-sampling of a factor $J$. Sampling the received signal at $J$ (oversampling factor) times the chip rate, we obtain the wide-sense stationary $P J \times 1$ vector signal $y_{m}(n)$ at the symbol rate. It is to be noted that the oversampling aspect (with respect to the symbol rate) is inherent to DS-CDMA systems by their very nature, due to the large (extra) bandwidth and the need to acquire chip-level resolution. This aspect directly translates into temporal diversity and explains the interference cancelation capability of these systems.

We consider the channel delay spread between the $k$th user and all of the $M$ sensors to be of length $l_k T$. Let $n_k \in \{0, 1, \ldots, P-1\}$ be the chip-delay index for the $k$th user: $h_{k,m}(n_k)$ is the first non-zero $J \times 1$ chip-rate sample of $h_{k,m}(p)$. Let us denote by $N_k$, the FIR duration of $g_{k,m}(t)$ in symbol periods. It is a function of $l_k$, $n_k$, and $P$. We nominate the user 1 as the user of interest and assume that $n_1 = 0$ (synchronization to user 1). The symbol sequence $\{s_m\}$ for other users are relabeled (delayed or advanced), so that their relative delay with respect to user 1 falls in $[0, T]$.

Let $N = \sum_{k=1}^{K} N_k$. The vectorized oversampled signals at $M$ sensors lead to a discrete-time $PM J \times 1$ vector signal at the symbol rate that can be expressed as

$$y(n) = \sum_{i=0}^{K-1} \sum_{j=0}^{N_k-1} g_j(i) a_k(n-i) + v(n)$$

$$= \sum_{i=0}^{K-1} G_{i,N_k} a_k(n) + v(n) = \mathbf{G}_N \mathbf{A}_N(n) + v(n), \quad (3)$$

$$y(n) = \left[ y_1(n) \right] \cdots \left[ y_M(n) \right] = \left[ \begin{array}{c} y_{11}(n) \\ \vdots \\ y_{1M}(n) \\ \vdots \\ y_{M1}(n) \\ \vdots \\ y_{MM}(n) \end{array} \right]$$

$$G_{i,N_k} = [g_i(N_k-1) \ldots g_i(0)]$$

$$\mathbf{G}_N = [\mathbf{G}_1 \mathbf{N}_1 \ldots \mathbf{G}_K \mathbf{N}_K]$$

$$\mathbf{A}_N(n) = [a_k(n-N_k+1) \ldots a_k(n)]^T$$

$$\mathbf{X}_N(n) = [A_{i1}^T(n) \ldots A_{iK}^T(n)]^T$$

$$\mathbf{y}(n) = \left[ \begin{array}{c} y_{11}(n) \\ \vdots \\ y_{1M}(n) \\ \vdots \\ y_{M1}(n) \\ \vdots \\ y_{MM}(n) \end{array} \right]$$

$$\mathbf{y}_m(n) = \left[ \begin{array}{c} y_{m1}(n) \\ \vdots \\ y_{mM}(n) \end{array} \right]$$

$$\mathbf{h}(i) = \left[ \begin{array}{c} h_{11}(i) \\ \vdots \\ h_{1M}(i) \\ \vdots \\ h_{M1}(i) \\ \vdots \\ h_{MM}(i) \end{array} \right]$$

$$\mathbf{h}_m(i) = \left[ \begin{array}{c} h_{m1}(i) \\ \vdots \\ h_{mM}(i) \end{array} \right]$$

Fig. 2. Signal model in continuous and discrete time for the $m$th sensor, showing only the contribution from one user.
or as a suboptimal processing or initialization stage at the base station. To this end, let us stack $L$ successive $y[n]$ vectors in a super vector

$$Y_L(n) = T_L(G_{\infty}) A_{\infty}(n-d) + V_L(n), \quad (4)$$

where, $T_L(G_{\infty}) = [T_L(G_{1,\infty}), \ldots, T_L(G_{K,\infty})]$, and $T_L(x)$ is a banded block Toeplitz matrix with $L$ block rows and $[x \; 0_{p(L-x)}]$ as first block row ($p$ is the number of rows in $x$), and $A_{\infty}(n-d)$ is the concatenation of user data vectors ordered as $[A_{\infty}^T(n-d), A_{\infty}^T(n-2-d), \ldots, A_{\infty}^T(n-K-d)]^T$. We shall refer to $T_L(G_{\infty})$ as the channel convolution matrix for the $k$th user.

Consider the noiseless received signal shown in fig. 3 for the contribution of user 1, from which the following observations can be made. Due to the limited delay spread, the effect of a particular symbol, $a_1(n-d)$, influences $N_1$ symbol periods, rendering the channel a moving average (MA) process of order $N_1-1$. We are interested in estimating the symbol $a_1(n-d)$ from the received data vector $Y_L(n)$. One can notice that $a_1(n-d)$ appears in the portion $Y_{N_1}$ of $Y_L(n)$. The shaded triangles constitute the ISI, i.e., the effect of neighboring symbols on $Y_{N_1}$. The contributions from the other (interfering) users to the received data vector have a similar structure. Note that to handle ISI and MAI, it may be advantageous to consider the longer received data vector $Y_L(n)$.

4. THE MULTICODE SCENARIO

In fig. 3, $\bar{g}_1 = T^H_1 h_1^1$, with $T_1 = [0 \; C^H_1 \; 0] \otimes I_{M_J}$ is the overall channel impulse response vector for the desired symbol, $a_1(n-d)$ of user 1. Let us consider user 2 to be a multicode co-user of user 1 (a certain user is employing two codes for higher data rates). The symbol of interest for this user is $a_2(n-d)$ and the channel impulse response for this user is

$$\bar{g}_2 = T^H_2 h_2^1, \quad T^1 = [0 \; C^H_2 \; 0] \otimes I_{M_J}.$$ 

Note that since the two users’ signals issue from the same point in space, the propagation channel is the same (given by $h_1$) for both of them. The matrix $C_2$ is the same as $C_1$ except that a different code $c_2$ builds up the band.

In the above setup, the symbols of interest at the $n$th instant are both symbols $a_1(n-d)$ and $a_2(n-d)$, and the matrices $T_1$ and $T^1$ are code correlators matched to all multipath components of these users. The goal of the present discussion is to obtain decorrelating receivers $f_1$ and $f_2$ to obtain, in a blind manner, the estimates of the desired symbols of the multicode user as $\hat{a}_1(n-d) = f_1 Y_L(n)$ and $\hat{a}_2(n-d) = f_2 Y_L(n)$.

5. THE DECENTRALIZED MMSE-ZF RECEIVER

It was shown in [1], that the MMSE-ZF receiver could be obtained by a proper implementation of the unbiased minimum output energy ^2(MOE) criterion. We shall refer to [1] for details while mentioning that the MMSE-ZF receiver can be implemented in the generalized sidelobe canceler (GSC) fashion as in the following.

Let us denote by

$$T_1 = \begin{bmatrix} 0 & C^H_1 & 0 \end{bmatrix} \otimes I_{M_J}; \quad T^1 = \begin{bmatrix} I & 0 & 0 \\ 0 & C^H_1 & 0 \\ 0 & 0 & I \end{bmatrix} \otimes I_{M_J},$$

the partial signature of the desired user and its orthogonal complement employed, respectively, in the upper and lower branches of the GSC, as shown in fig. 4. $C^H_1$ is the orthogonal complement of $C_1$, the tall code matrix given in section 3 ($C^H_1 C_1 = 0$). Then, $C^H_1 Y_{N_1} = T_1 Y_L$ and the matrix $T^1$ acts as a blocking transformation for all components of the signal of interest. Note that $P_{X^H} + P_{X^H} = I$, where, $P_X$ is the projection operator (projection on the column space of $X$). Then the LCMV problem can be written as

$$\min_{f: f^T R_{YY} f = 1} \frac{1}{f^T T^H_1 h_1^1} f^T R_{YY} f^H,$$

where, $[h_1 \; h^1]$ is a square non-singular matrix, and $h^H_1 h^1 = 0$. Note that in the LCMV problem (GSC formulation) there is a number of constraints to be satisfied. However, imposing the second set of constraints, namely $f^T T^H_1 h^1 = 0$ has no consequence because the criterion automatically leads to their satisfaction once, $\text{span} \{ R_{YY} \} \cap \text{span} \{ T^H_1 \} = \text{span} \{ T^H_1 h_1^1 \}$, i.e., when the intersection of the signal subspace and the subspace spanned by the columns of $T^H_1$ is one dimensional.

The matrix $T_1$ is nothing but a bank of correlators matched to the $l_1$ delayed multipath components of user 1’s code sequence. Note that the main branch in fig. 4 by itself gives an unbiased response for the desired symbol, $a_1(n-d)$, and corresponds to the (normalized) coherent RAKE receiver. For the rest, we have an estimation problem, which can be solved in the least squares sense, for some matrix $Q$. This interpretation of the GSC corresponds to the pre-combining (or pathwise) interference (ISI and MAI) canceling approach.

The vector of estimation errors is given by

$$Z(n) = T_1 - QT^1 \quad Y_L(n).$$

Since the goal is to minimize the estimation error variances, or in other words, estimate the interference term in the upper branch as closely as possible from $T^1 Y_L(n)$, the interference cancellation problem settles down to minimization of the trace of the estimation error covariance matrix $R_{ZZ}$ for a matrix filter $Q$, which results in

$$Q = \left( T_1 R^T T^1 \right)^{-1} \left( T^1 R_1^T T_1^{-1} \right)^{-1},$$

and where, $R^T_1$ is the noiseless (denoised) data covariance matrix, $R_{YY}$ with the subscript removed for convenience. The output

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1 also termed as MMSE-ZF receiver and projection receiver

\[^2\text{a derivative of the minimum variance distortionless response (MVDR) method, and a particular instance of the linearly constrained minimum-variance (LCMV) criterion.}\]
ZF constraint expressed as analogous treatment to the single code case. However, this time, the noiseless case, the evidence of which is the rank-1 estimation results in perfect interference cancelation (both ISI and MAI) in the R as the maximum eigenvector of the matrix R

to get the symbol estimate, f

where, R

as before [1] and the problem to be solved is given as

from where we obtain as solution for f,

Note that a noiseless (and regularized) version of the received signal covariance matrix is employed in the derivations to account for the fact that the unbiasedness and zero-distortion constraint (the original constraint in the MVDR criterion) are equivalent only in the noiseless case.

It can be noticed that the channel impulse response, h_1 is yet unknown in the above derivation. As discussed in [1], the MOE which is given by

MOE = tr \{ f R_{YY} f^H \}

(16)
can now be maximized over all possible values of h_1 according to Capon's principle

\[ \max_{h_1 \in \mathbb{C}^L} \text{MOE}(h_1) = \max_{h_1 \in \mathbb{C}^L} \left\{ H^H \left( T^H R_{YY}^{-1} T \right)^{-1} H \right\} \]

(17)
to lead to a blind channel estimate obtained as

\[ \hat{h}_1 = V_{\text{max}} \left\{ \left( T^H R_{YY}^{-1} T \right)^{-1} \right\} + \left( T^H R_{YY}^{-1} T \right)^{-2} \}

(8)

where, \[ \cdot \] stands for the square sub-matrix formed by the i\text{th} block row and j\text{th} block column of the matrix \[ \cdot \].

In a GSC implementation as shown in fig. 4, the main branch will be replaced by T^H, the set of code correlators for both sub-users of the multicode user, and the second branch by T^H, a blocker (T^H T = 0) for the desired signal components of both sub-users. A simulation example is presented in the figure below.

![Graph showing output SINR vs. input SNR for different receivers](image)

5. REFERENCES
