BLIND CHANNEL ESTIMATION IN MULTICODE CDMA SYSTEM WITH ANTENNA ARRAY

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ABSTRACT

This paper addresses the problem of blind channel estimation in a multicode DS-CDMA system in multipath environment using antenna array at the receiver. Long code CDMA system model is employed. Multicode transmission is used to provide a higher data rate to a mobile user and receiver antenna array improves the signal to interference and noise ratio (SINR). First we extend blind multicode channel estimation methods to CDMA receiver with multiple antennas. Subspace channel estimation is employed with two different pre-processing schemes. We also extend the principal component channel estimation method used for antenna array CDMA receiver to a multicode model. The performances of the proposed methods are compared in simulation using mean square error measure and bit error rates. Different numbers of interfering users are used in simulations and the impact of code selection is considered.

1. INTRODUCTION

Multipath propagation might cause severe degradation in wireless communication systems. If the multipath components are more than one chip apart from each other they can be coherently combined with a RAKE receiver to obtain multipath diversity. When mutliple antennas are available 2D RAKE receiver can be used to provide also antenna diversity. In this paper we will consider a synchronous DS-CDMA system with aperiodic spreading codes and antenna arrays at the receiver. Such model is valid for uplink multicode transmission but with one single antenna element it is valid also for downlink transmission. Multicode transmission can be used to provide higher data rates to a user. We'll use low spreading factors and assume the codes are known at the receiver. These assumptions are commonly used for the multicode transmission.

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In this paper we extend the blind channel estimation methods considered in [1] to CDMA receivers with multiple antennas. Receiver antenna arrays are used to improve the signal to interference and noise ratio (SINR). The first method performs partial despreading operation followed by a subspace method for channel identification. The method stems from subspace method of [2] for downlink CDMA. However, the proposed method is less sensitive to code selection and it has significantly lower computational complexity, and is more robust in the face of interference than [2]. The second proposed technique extends the principal component method used for antenna array CDMA receivers [3] to multicode system. The extension may be found in a straight forward manner. The performance of the proposed methods is compared to multi-antenna extension of the method given [2] using bit error rates and mean square error. Different numbers of interfering users are used in simulations. The impact of code selection and different channels to the performance is studied, too.

This paper is organized as follows. In section 2, the system model is presented. In section 3, the proposed blind channel estimation algorithms are described in detail and section 4 considers the principal component method. In section 5, simulation examples are given demonstrating the reliable performance of the proposed methods. Finally, section 5 concludes the paper.

2. SYSTEM MODEL

We consider a multicode DS-CDMA system model with aperiodic spreading codes. In multicode system P codes are assigned to one user to obtain higher data rate. The system model is similar to synchronous transmission with P users. An antenna array of M elements is used at the receiver.

Due to aperiodic (long) codes each symbol is spread with different code during the observation period. The pth code for nth symbol is $\mathbf{c}_{p,n} = [c_{p,n}(1) \quad \ldots \quad c_{p,n}(SF)]^T$

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where SF is the spreading factor. The code matrix $\mathbf{C}_{p,n}$ for the pth code and nth symbol is defined as follows:

$$\begin{bmatrix} c_{p,n}(1) & 0 & \dots & 0 \\ c_{p,n}(2) & c_{p,n}(1) & \dots & 0 \\ \vdots & \vdots & \vdots & \vdots \\ c_{p,n}(L) & c_{p,n}(L-1) & \dots & c_{p,n}(1) \\ \vdots & \vdots & \vdots & \vdots \\ c_{p,n}(SF) & c_{p,n}(SF-1) & \dots & c_{p,n}(SF-L+1) \\ 0 & c_{p,n}(SF) & \dots & c_{p,n}(SF-L+2) \\ \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \dots & c_{p,n}(SF) \end{bmatrix},$$

where L is the channel length. The combined code matrix for P codes for nth symbol is

$$\mathbf{C}_n = [\mathbf{C}_{1,n} \ \dots \ \mathbf{C}_{P,n}].$$

The code matrix for multi-antenna receiver may be modeled as

$$\mathcal{C}_n = \mathbf{C}_n \otimes \mathbf{I}_M,$$

where $\otimes$ denotes Kronecker product. Define the combined multipath multi-antenna channel vector as $LM \times 1$ vector $\mathbf{h} = [\mathbf{h}_1^T \ \dots \ \mathbf{h}_L^T]^T$, where the complex channel coefficients of the lth multipath component for each M antenna elements are combined to one vector $\mathbf{h}_l$. For a multicode system the channel matrix can be written as $\mathbf{H} = \mathbf{I}_P \otimes \mathbf{h}$.

The received signal is sampled once per chip and collected to a vector $\mathbf{y}(n)$ with $SF + L - 1$ chips.

$$\mathbf{y}(n) = \mathcal{C}_n\mathbf{H}\ \mathbf{s}(n) + \mathcal{C}_n^b\mathbf{H}\ \mathbf{s}(n-1) + \mathcal{C}_n^f\mathbf{H}\ \mathbf{s}(n+1) + \mathbf{v}(n),$$

where $\mathbf{v}(n)$ is noise and $\mathbf{s}(n) = [s_1(n) \ \dots \ s_P(n)]^T$ are the nth transmitted symbols for all the P codes. The contribution of the previous symbol to the desired symbol is modeled with code matrix

$$\mathcal{C}_n^b = \begin{bmatrix} \mathbf{0}_{M*SF\times L} \\ (M*SF+1)\text{th row of } \ \mathcal{C}_{n-1} \\ \vdots \\ M*(SF+L-1)\text{th row of } \ \mathcal{C}_{n-1} \end{bmatrix}.$$

Correspondingly, the following chip contributes with $\mathcal{C}_n^f$ which has the $M * (L-1)$ first rows from the code matrix $\mathcal{C}_{n+1}$ followed by $M * SF$ rows of zeros. The size of all the code matrices is $M * (SF + L - 1) \times L$.

3. SUBSPACE CHANNEL ESTIMATION METHOD USING PARTIAL DESPREADING

In the blind subspace channel estimation, we only use the part of received signal that is not deteriorated by previous or following symbols. The part of the received signal that is free from inter symbol interference (ISI) may be model as:

$$\tilde{\mathbf{y}}(n) = \tilde{\mathcal{C}}_n\mathbf{H}\mathbf{s}(n) + \tilde{\mathbf{v}}(n),$$

where $\tilde{\mathcal{C}}_n$ denotes the rows from $(L * M)$ to $(SF * M)$ of the matrix $\mathcal{C}_n$. By partially despreading the received signal, we get the following approximation

$$\mathbf{x}(n) = \tilde{\mathcal{C}}_n^H\tilde{\mathbf{y}}(n) \approx \mathbf{H}\mathbf{s}(n) + \tilde{\mathcal{C}}_n^H\tilde{\mathbf{v}}(n) \tag{1}$$

due to good autocorrelation and crosscorrelation properties of the spreading codes. In [2] the pseudoinverse of $\tilde{\mathcal{C}}_n$ was used instead of partial despreading. It has very good performance in the absence of noise and interference but the high computational complexity involved with pseudoinversion makes the method less appealing. The number of rows in $\tilde{\mathcal{C}}_n$ must be greater or equal to the number of columns and the matrix should be full column rank. With low spreading factors these criteria may not be met. In [2] it was stated that the full rank condition holds in practice, if the first criterion is met. Simulation will show that is may not be the case, and the method becomes sensitive to the code selection and noise.

After pre-processing stage the channel estimate may be obtained using widely used subspace processing. In the absence of noise, $\mathbf{x}(n)$'s lie in a subspace spanned by columns of $\mathbf{H}$ (i.e., the latter term may be neglected in equation (1)). By assuming that the channel is nearly constant during a period of N symbols, we collect the partially despread symbols to a matrix

$$\mathbf{X} = [\mathbf{x}(1) \ \dots \ \mathbf{x}(N)] \approx \mathbf{H}\mathbf{S}_N,$$

where $\mathbf{S}_N = [\mathbf{s}(1) \ \dots \ \mathbf{s}(N)]$ is the matrix of N transmitted data symbols. By performing the singular value decomposition of the preprocessed data matrix

$$\mathbf{X} = [\mathbf{U}_s \ \mathbf{U}_n] \begin{bmatrix} \mathbf{\Sigma}_s & \mathbf{0} \\ \mathbf{0} & \mathbf{0} \end{bmatrix} \begin{bmatrix} \mathbf{V}_s^H \\ \mathbf{V}_n^H \end{bmatrix}$$

the decomposition to signal and noise subspaces may be achieved. The channel matrix $\mathbf{H}$ and $\mathbf{U}_s$ span the same subspace. Hence, the noise subspace spanned by singular vectors $\mathbf{U}_n$ is orthogonal to $\mathbf{H}$, i.e. $\mathbf{U}_n^H\mathbf{H} = 0$. This can be rewritten as $[\mathbf{D}_1\mathbf{h} \ \dots \ \mathbf{D}_P\mathbf{h}]$ where $\mathbf{D}_1$ are the first ML columns of $\mathbf{U}_n^H$, $\mathbf{D}_2$ are the next ML columns, etc. Or equivalently $\bar{\mathbf{D}}\mathbf{h} = 0$, where $\bar{\mathbf{D}} = [\mathbf{D}_1^T \ \dots \ \mathbf{D}_P^T]^T$. At this point we will set the first element of the estimated channel vector to 1 and solve for the other $L * M - 1$ elements. The obtained estimate must be scaled based on some initial knowledge. For example a phase estimate of the first channel tap in the first antenna element can be used. By setting $\mathbf{h} = [1 \ \ \mathbf{g}^H]^H$ we obtain $\bar{\mathbf{D}}\mathbf{e}_1 + \tilde{\mathbf{D}}\mathbf{g} = 0$, where $\tilde{\mathbf{D}}$ is the $\bar{\mathbf{D}}$ with first column deleted. Solving this gives the channel estimate up to a complex scale:

$$\hat{\mathbf{g}} = -\tilde{\mathbf{D}}^\dagger\bar{\mathbf{D}}\mathbf{e}_1$$

where $\mathbf{e}_1$ is vector $[1\ 0\ \dots\ 0]^T$. Obviously, the neglected noise term $\tilde{\mathcal{C}}_n^H \tilde{\mathbf{v}}(n)$ has impact on the performance of the subspace processing. In the simulation section the two different pre-processing methods are tested in different noise conditions.

4. PRINCIPAL COMPONENT METHOD FOR MULTICODE

An alternative blind channel estimation the method proposed for long code CDMA system in [3]. It is based on principal eigenvector of the difference between pre- and post-despreading covariance matrices. We extended it to multicode system in [1]. In this paper it is applied to the antenna array model with multicode transmission.

The difference matrix is obtained by $\hat{\mathbf{R}}_{\mathbf{x}} - \hat{\mathbf{R}}_{\mathbf{Y}}$, where

$$\hat{\mathbf{R}}_{\mathbf{Y}} = \frac{1}{N} \sum_{n=1} \mathbf{Y}(n)\mathbf{Y}^H(n)$$

is the pre-despreading covariance matrix and

$$\hat{\mathbf{R}}_{\mathbf{x}} = \frac{1}{PN} \sum_{p=1}^{P} \sum_{n=1}^{N} \mathbf{Y}(n)\mathbf{c}_{p,n}\mathbf{c}_{p,n}^H \mathbf{Y}^H(n)$$

is the multicode post-despreading covariance matrix. The difference to the original method given in [3] is the averaging performed over all the P codes. The matrix $\mathbf{Y}(n)$ consists of the received signal sampled once per chip for the nth symbol. There are rows for each antenna elements and multipath components, i.e. all together ML rows.

The channel estimate $\hat{\mathbf{h}}$ up to scaling is the principal eigenvector $\mathbf{r}$ of the difference matrix $\hat{\mathbf{R}}_{\mathbf{x}} - \hat{\mathbf{R}}_{\mathbf{Y}}$.

5. SIMULATIONS

In this section the simulations are performed to study the robustness of the proposed methods in the face of interference caused by other users and different multipath channels are described. Also the influence of the code selection is considered. The used performance measures are the mean square error (MSE) of the channel estimate , the error variance of the phase estimate and the bit error rate (BER). The BER values are obtained with a 2D-RAKE receiver using maximal ratio combining and the obtained channel coefficients. As comparison the BER values are also given for a 2D-RAKE receiver using the known true channel tap coefficients. This optimal coherent combining with the known channel is referred as optimal in the simulation results. We denote the extension of the pseudoinverse method given in [2] to antenna arrays as WF, the partial despreading method as PD and the multicode extension of the blind channel estimate in [3] as MP.

The spreading code is a combination of orthogonal variable spreading factor (OVSF) codes and long Gold codes. In this paper the spreading factor of 16 is used for all the signals. The desired user uses $P = 4$ different OVSF codes and one long code. Different long codes have used in simulations, but they had no influence on the results. The $P = 4$ codes can be obtained with 1820 different ways form the possible 16 OVSF codes. In all the figures shown in this paper the transmission powers of all the codes have been equal and each channel estimation is based on $N = 30$ QPSK symbols.

The interference has been simulated with both additive white Gaussian noise (AWGN) and actual interfering users. The multiple access interference caused by the other users is simulated with 0 to 16 interfering signals each having a unique long code and a unique radio channel.

In simulations we have used both random Rayleigh fading channels and the stochastic channel model presented in [4]. For both models we assume time invariant model and one to four receiver antennas equally spaced $d = \lambda/2$ apart. The number of antennas is denoted with M. For desired user the channel length L is set to 3 and for the interfering users $L = 2$. The stochastic model defines a spatial correlation between antennas at the receiver for typical urban environment with the matrix [4]:

$$\begin{bmatrix} 1.00 & 0.91 & 0.73 & 0.46 \\ 0.91 & 1.00 & 0.91 & 0.73 \\ 0.73 & 0.91 & 1.00 & 0.91 \\ 0.46 & 0.73 & 0.91 & 1.00 \end{bmatrix}.$$

The multipath components are generated starting with randomly generated zero mean Gaussian. After spatial correlation also the direction of arrival (DOA) of the signal is included in the model. For the channels used in this paper the DOA has been set to 30° against broad side of the array for the desired user and the interfering users are uniformly distributed in DOA = 30° ± 10°. We assume knowledge of the phase of the first channel tap received in the first antenna element and compare the obtained phase estimates of the other $L * M - 1$ channel taps. Four different randomly generated correlated channels and varying number of random channels have been used in simulations.

In the box plot of Figure 1 a the phase angle error variances are given for the MP, PD and WF methods for all the 1820 different OVSF code combinations. $M = 4$ antennas were used and the correlated stochastic channel model was used. The interference is due to 8 and 12 interfering users and AWGN. The signal to noise ratio for one code was $SNR = 12$ dB. By SNR we mean $= \mathcal{E}_s/N_0$, where $\mathcal{E}_s$ is the energy per symbol and $\frac{1}{2}N_0$ is the noise variance. The variances are only shown for the first channel tap of the third antenna element. In a boxplot the box defines the quintile range (from 25% percentile to 75% percentile). The line inside the box is the median. The 'whiskers' are lines extending from each end of the box to show the extent of the rest of the data. The length of a whisker is defined as 1.5 times the length of the quintile range. The possible outliers, outside the whiskers area, are marked by crosses.

The PD and MP estimators give channel estimates with low variance in all the 11 unknown phases even though the interference level is high. The average bit error rates corresponding to the case with 12 interfering users are given in Table 1 for several different channels. The Figure 1 shows that even though the WF gives results similar to PD with some code combinations, the results in general are much worse. The WF method is very sensitive to the selected code combination which appears to follow from the rank properties of the code matrix $\tilde{\mathbf{C}}_n$. In [2] it was assumed that if the $\tilde{\mathbf{C}}_n$ matrix has more rows than columns (now 14×12), the full rank assumption necessary for this method is met in practice. However with the given parameters, roughly 17% of the matrices are not full rank. If only $P = 3$ codes are used, still 4% of the matrices are not full rank, even though the size of the $\tilde{\mathbf{C}}_n$ matrix is now 14×9. The obtained channel estimate is better with $P = 3$ codes than $P = 4$ codes, but in absence of noise, the utilization of the fourth code improves the channel estimates.

The mean square error, $MSE = \frac{1}{K}\sum_{k=1}^{K} ||\hat{\mathbf{h}}^k - \mathbf{h}||^2$, values are given for the case where 12 interfering users and AWGN are present at the bottom of Figure 1. The results first are averaged over $K = 70$ simulation runs and then over four different channels obtained with the statistical model for $M = 4$ antennas. The average MSE values are plotted against the code selection. The performance of WF seems to depend more on the selected codes than on the channel. The codes for which MSE is small have similar number of rank deficient $\tilde{\mathbf{C}}_n$ matrices as the for higher MSE values. When the number of rank deficient matrices was high, the MSE value are similar to the average MSE.

Average bit error rates obtained with a 2D RAKE receiver corresponding to the MSE values shown in Figure 1 are given in Table 1. The last column (optimal coherent combining) is the BER calculated with the known channel tap values. The BER values for the four channels were also calculated in less noisy situations. The middle rows of Table 1 show the BER values in case where the received signal consist only of the desired four signals and AWGN background noise with $SNR = 6$ dB per signal. In the last rows are the results obtained in the absence of noise. These values indicate clearly that the WF method has very good performance in theory since the interference due to multipath of the desired signals is completely canceled prior to channel estimation with the pseudoinverse of $\tilde{\mathbf{C}}_n$ even though it is not always a full rank matrix. The MP channel estimate gives better results than either of the two subspace methods. The BER values of MP are close to optimal results for all the channels considered.

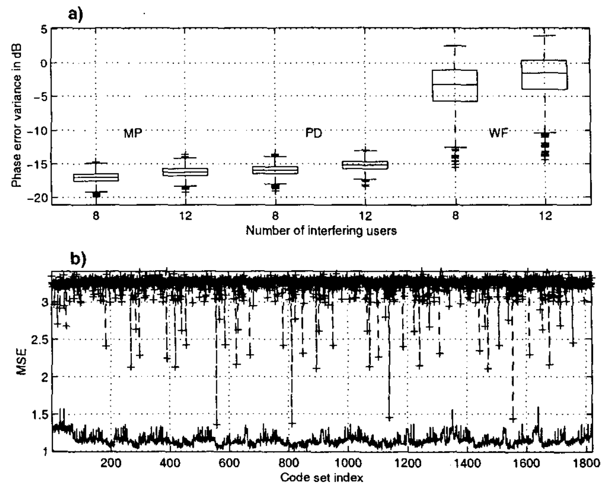

Fig. 1. Sensitivity to code selection. Figure a: Phase angle errors variances for three different channel estimators, for one multipath component of one antenna element. On the X-axis are two different number of interfering users. On the left figure is the principal component method (MP), on the middle is the partial despreading followed by subspace estimator (PD) and on the right is the pseudoinverse followed by subspace (WF). 1820 different code combinations were simulated and each variance is based on averaging over 70 simulation runs. Figure b: Mean square error of the subspace channel estimates. Number of interfering users was 12. On the x-axis are different code sets for desired user. The solid line at bottom is for PD-estimate and the top line ('+–') is for WF-estimate. The MSE values are averaged over four random statistical channels.

The improved bit error rate versus SNR when multiple antennas are employed is seen in Figure 2. In the upper figure the BER for PD is plotted for $M = 1, 2, 4$ antennas. Only the four codes of the desired used where transmitted and so the interference is due to multipath propagation. The corresponding optimal combining 2D RAKE results are also shown. In the middle figure the same results are shown for WF and at the bottom the MP BER curves are shown. For high SNR, the WF gives the best performance no matter how many antennas are used. The BER values are close to optimal 2D RAKE. At lower SNR both the PD and MP outperform WF. When only one antenna is used the performance is quite similar for all the methods at lower SNR. When two or more antennas are available the best performance is obtained with MP. Also BER of PD is improved significantly when more antennas are added, but it's slightly worse than MP which may also bee seen in Ta-

Table 1. Average bit error rates desired user codes

	MP	PD	WF	Optimal
12 interfering users + noise (SNR 12 dB)				
channel 1	0.0400	0.0533	0.1643	0.0313
channel 2	0.0195	0.0520	0.1891	0.0146
channel 3	0.0030	0.0151	0.1068	0.0018
channel 4	0.0312	0.0483	0.1391	0.0264
Only desired user with 4 codes + noise (SNR 6 dB)				
channel 1	0.0215	0.0229	0.0815	0.0154
channel 2	0.0150	0.0227	0.0982	0.0115
Only desired user with 4 codes, no interference, no noise				
channel 1	0.0760	0.0071	0.0043	0.0042
channel 2	0.0036	0.0059	0.0026	0.0026
channel 3	0.0000	0.0000	0.0000	0.0000
channel 4	0.0063	0.0128	0.0107	0.0047

ble 1. The WF seems not to benefit from multiple antennas at lower SNR and the performance is similar to one antenna case.

6. CONCLUSIONS

In this paper we considered pre-processing methods for blind subspace channel estimation algorithms for multicode DS-CDMA receiver with antenna array. Long code system model is used. The first method employs partial despreading as a pre-processing step. The second method extends the pseudoinverse pre-processing used for multicode to antenna array systems. The third method uses the principal eigenvector of the difference between post- and pre-despreading covariance matrices.

When the spreading factor is small compared to number of codes allocated to one user the performance of the pseudoinverse method depends on the selected codes. The partial depsreading method has both lower computational complexity and better performance measured in both channel estimation error and bit error rate comparisons. The principal component method gives better bit error rate values than the two subspace methods when more than one antennas are employed except in very high signal to noise ratios when the pseudoinverse is the best. The antenna gain obtained with multiple antennas is clearly seen for the principal component and partial despreading methods, but not for the pseudoinverse method.

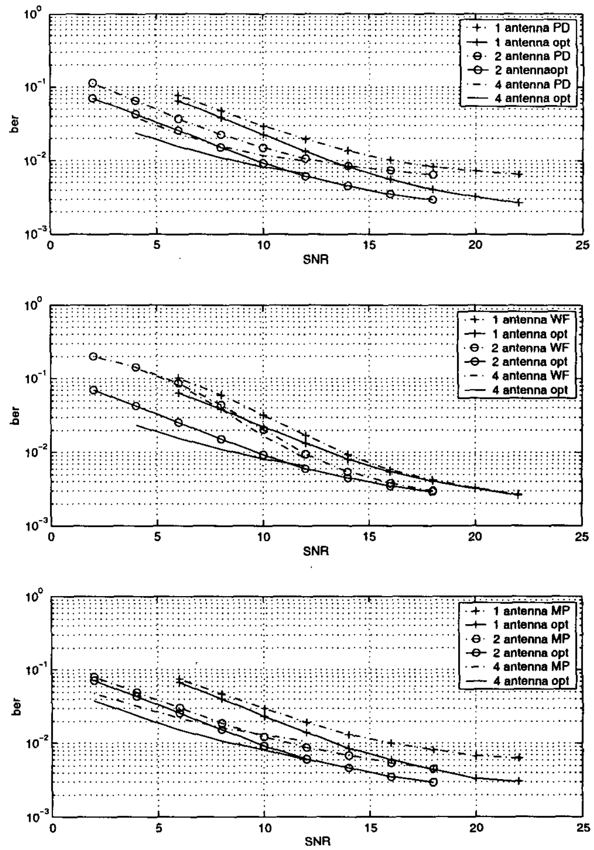

Fig. 2. Bit error rates for PD, WF and MP estimates with different number of antennas.

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