THE USE OF DECORRELATION AND CODE DIVISION MULTIPLEXED PILOTS FOR DS-CDMA SYSTEMS WITH MULTIANTENNA RECEIVERS

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ABSTRACT

In this paper, we investigate the use of code division multiplexed (CDM) pilots for channel estimation for direct-sequence code division multiple access (DS-CDMA) systems with multiple receive antennas. Decorrelation is used on the output of each receive antenna to remove MAI from both the data and pilot streams. This leads to more reliable channel estimation which in turn reduces the combination loss. We look at the bit error rate (BER) performance of this system in flat Rayleigh fading.

1. INTRODUCTION

One of the biggest challenges facing DS-CDMA systems in wireless communications is multiple access interference (MAI) caused by co-channel users. The signature waveforms of other users are usually not orthogonal to the desired user's signature waveforms, leading to interference. Different techniques can be used to overcome these impediments; these include multiuser detection and antenna array processing. Multiuser detectors (MUDs) [1] reduce the effects of MAI by exploiting its well-known structure. Alternatively, space domain techniques such as antenna array processing [2] can also reduce the effects of MAI, while providing diversity combining leading to improved signal to interference and noise ratio (SINR) by coherently combining the signals from different antennas.

Multiple input multiple output (MIMO) systems [3] employ multiple antennas at both the transmitter and the receiver. High system capacities are realized using MIMO systems. However, most research into MIMO systems assume that the receiver knows the channel gains on each transmitter-receiver pair.

Pilot symbols [4]-[5] have been used to estimate channel gains so as to provide pseudo-coherent detection. We can also use the channel gain estimates to weight the outputs of the different receive antennas for maximal ratio combining [5]. In this paper, we propose a DS-CDMA system in which the different users transmit code division multiplexed data and pilots to a multiantenna receiver. MAI from different users is present in both the data and the pilot streams, which can affect the receiver's ability to reliably estimate the channel gains. To combat this, we employ a decorrelator on the output of each antenna to remove the MAI from both the pilots and the data. The decorrelator completely eliminates the MAI, however, the noise variance is increased slightly. Since the data no longer contains MAI, the optimum combination scheme is the maximal ratio combination, thus the channel weight estimates obtained from the decorrelated pilot streams can be used to weight the outputs of the different antennas.

2. SYSTEM MODEL

The system block diagram for N users is shown in Figure 1. Each transmitter has 1 transmit antenna, and each user transmits to an N antenna receiver. The channel between user i and receive antenna j is a slowly varying flat Rayleigh fading channel. The lowpass equivalent of the received signal is multiplied by a complex channel gain α_{ij}. For ease of implementation of the simulations, we consider a symbol synchronous DS-CDMA system.

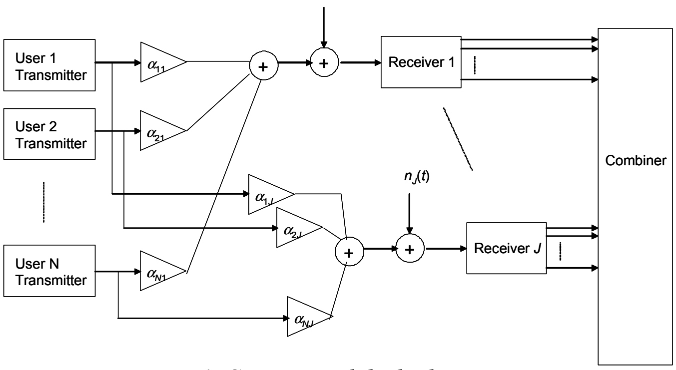

Figure 1: System Model Block Diagram.

We assume that BPSK modulation is used and the pilot is a spreading waveform $c_{pi}(t)$, that is orthogonal to the ith user's data spreading waveform, $c_{di}(t)$. The ith user's transmitted signal on the interval $kT_b \leq t \leq (k+1)T_b$ is given by:

$$s_i(t) = b_i^{(k)} A_{di} c_{di}(t)\cos(2\pi f_c t) + A_{pi} c_{pi}(t)\cos(2\pi f_c t)$$

where $b_i^{(k)}$ is the ith user's bit on the kth signalling interval, A_{di} is the data carrier amplitude, A_{pi} is the pilot carrier amplitude and f_c is the carrier frequency.

On the interval $kT_b \leq t \leq (k+1)T_b$, the received signal at the jth antenna is given by:

$$r_j(t) = \sum_{i=1}^{N} [b_i^{(k)} G_{ij}^{(k)} A_{di} c_{di}(t)\cos(2\pi f_c t + \theta_{ij}^{(k)}) + G_{ij}^{(k)} A_{pi} c_{pi}(t)\cos(2\pi f_c t + \theta_{ij}^{(k)})] + n_j(t)$$

where $G_{ij}^{(k)}$ is the amplitude gain of the ith user's signal to the jth receiver on the kth signalling interval, $\theta_{ij}^{(k)}$ is the phase shift associated with the channel between the ith transmitter and the jth receiver on the kth signalling interval, ad $n_j(t)$ is the white Gaussian noise at the jth receiver. We consider only slow Rayleigh fading in this paper, thus $G_{ij}(k)$ is a Rayleigh distributed random variable that is highly correlated to $G_{ij}^{(k+1)}$. Also,

$$G_{ij}^{(k)} = \sqrt{\mathrm{Re}\{\alpha_{ij}^{(k)}\} + \mathrm{Im}\{\alpha_{ij}^{(k)}\}}$$

where $\mathrm{Re}\{a_{ij}(k)\}$ and $\mathrm{Im}\{\alpha_{ij}(k)\}$ are independent Gaussian distributed random variables with 0 mean and variance ½.

The jth receiver is shown in Figure 2. It is made up of $2N$ matched filters, one for each user's data spreading waveform and one for each user's pilot spreading waveform. On the kth signalling interval, the output of the receiver can be viewed as a vector $\mathbf{U_j}^{(k)}$, with real part $\mathrm{Re}\{\mathbf{U_j}^{(k)}\}$ and imaginary part $\mathrm{Im}\{\mathbf{U_j}^{(k)}\}$ which are given by:

$$\begin{aligned}
\mathrm{Re}\{\mathbf{U_j}^{(k)}\} &= [U_{1jdI}^{(k)} \quad U_{1jpI}^{(k)} \quad \cdots \quad U_{NjdI}^{(k)} \quad U_{NjpI}^{(k)}]^T \\
\mathrm{Im}\{\mathbf{U_j}^{(k)}\} &= [U_{1jdQ}^{(k)} \quad U_{1jpQ}^{(k)} \quad \cdots \quad U_{NjdQ}^{(k)} \quad U_{NjpQ}^{(k)}]^T
\end{aligned}$$

where $U_{ijdI}^{(k)}$ is the ith user's inphase data decision variable on the kth signalling interval on the jth receiver, $U_{ijdQ}^{(k)}$ is the ith user's quadrature data decision variable on the kth signalling interval on the jth receiver, $U_{ijpI}^{(k)}$ is the ith user's unfiltered inphase channel gain times pilot estimate on the kth signalling interval on the jth receiver and $U_{ijpQ}^{(k)}$ is the ith user's unfiltered quadrature channel gain times pilot estimate on the kth signalling interval on the jth receiver. It can be shown that $U_{ijdI}^{(k)}$ is given by:

$$U_{ijdI}^{(k)} = b_i^{(k)} A_{di} T_b \,\mathrm{Re}\{\alpha_{ij}^{(k)}\} + N_{ijdI}^{(k)}$$
$$\sum_{l=1, l\neq i}^{N} [b_l^{(k)} A_{dl} T_b \,\mathrm{Re}\{\alpha_{lj}^{(k)}\}\rho_{idld} + A_{pl} T_b \,\mathrm{Re}\{\alpha_{lj}^{(k)}\}\rho_{idlp}]$$

where the first term is the desired information multiplied by the fading process' inphase component, the second term is the inphase noise component and the summation represents the MAI due to the other users' data and pilot signals. The terms ρ_{idld} and ρ_{idlp} are the cross-correlations between the desired user's data spreading waveform and the interfering users' data spreading waveforms and their pilot spreading waveforms respectively.

We can find similar expressions for $U_{ijdQ}^{(k)}$, $U_{ijpI}^{(k)}$, and $U_{ijpQ}^{(k)}$. We define $\mathbf{A_{jI}}^{(k)}$ as the inphase information vector on signalling interval k and $\mathbf{A_{jQ}}^{(k)}$ as the quadrature information vector on signalling interval k below:

$$\begin{aligned}
\mathbf{A_{jI}}^{(k)} &= [X_{d1I}^{(k)} \quad X_{p1I}^{(k)} \quad \cdots \quad X_{dNI}^{(k)} \quad X_{pNI}^{(k)}]^T \\
\mathbf{A_{jQ}}^{(k)} &= [X_{d1Q}^{(k)} \quad X_{p1Q}^{(k)} \quad \cdots \quad X_{dNQ}^{(k)} \quad X_{pNQ}^{(k)}]^T
\end{aligned}$$

where $X_{diI}^{(k)} = b_i^{(k)} A_{di} T_b \mathrm{Re}\{\alpha_{ij}^{(k)}\}$, $X_{piI}^{(k)} = A_{pi} T_b \mathrm{Re}\{\alpha_{ij}^{(k)}\}$, $X_{diQ}^{(k)} = b_i^{(k)} A_{di} T_b \mathrm{Im}\{a_{ij}^{(k)}\}$, and $X_{piQ}^{(k)} = A_{pi} T_b \mathrm{Im}\{a_{ij}^{(k)}\}$.

It can be shown that:

$$\begin{aligned}
\mathrm{Re}\{\mathbf{U_j}^{(k)}\} &= \mathbf{R}\mathbf{A_{jI}}^{(k)} + \mathbf{N_{jI}}^{(k)} \\
\mathrm{Im}\{\mathbf{U_j}^{(k)}\} &= \mathbf{R}\mathbf{A_{jQ}}^{(k)} + \mathbf{N_{jQ}}^{(k)}
\end{aligned}$$

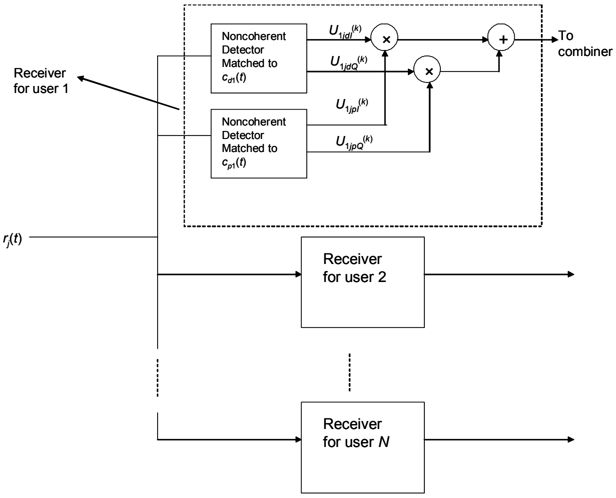

Figure 2: Block diagram of the jth receiver.

where $\mathbf{R}$ is the cross-correlation matrix, and $\mathbf{N_{jI}}^{(k)}$ and $\mathbf{N_{jQ}}^{(k)}$ are output noise vectors. They are given by:

$$\mathbf{R} = \begin{bmatrix}
\rho_{1d1d} & \rho_{1d1p} & \cdots & \rho_{1dNd} & \rho_{1dNp} \\
\rho_{1p1d} & \rho_{1pNp} & \cdots & \rho_{1pNd} & \rho_{1pNp} \\
\vdots & \vdots & \ddots & \vdots & \vdots \\
\rho_{Nd1d} & \rho_{Nd1p} & \cdots & \rho_{NdNd} & \rho_{NdNp} \\
\rho_{Np1d} & \rho_{Np1p} & \cdots & \rho_{NpNd} & \rho_{NpNp}
\end{bmatrix}$$

$$\begin{aligned}
\mathbf{N_{jI}}^{(k)} &= [N_{1jdI}^{(k)} \quad N_{1jpI}^{(k)} \quad \cdots \quad N_{NjdI}^{(k)} \quad N_{NjdI}^{(k)}]^T \\
\mathbf{N_{jQ}}^{(k)} &= [N_{1jdQ}^{(k)} \quad N_{1jpQ}^{(k)} \quad \cdots \quad N_{NjdQ}^{(k)} \quad N_{NjdQ}^{(k)}]^T
\end{aligned}$$

Also, $\rho_{idid} = \rho_{ipip} = 1$ and $\rho_{idip} = \rho_{ipid} = 0$.

By multiplying $\mathrm{Re}\{\mathbf{U_j}^{(k)}\}$ and $\mathrm{Im}\{\mathbf{U_j}^{(k)}\}$ by $\mathbf{R}^{-1}$, we get:

$$\begin{aligned}
\mathbf{R}^{-1}\mathrm{Re}\{\mathbf{U_j}^{(k)}\} &= \mathbf{A_{jI}}^{(k)} + \mathbf{R}^{-1}\mathbf{N_{jI}}^{(k)} \\
\mathbf{R}^{-1}\mathrm{Im}\{\mathbf{U_j}^{(k)}\} &= \mathbf{A_{jQ}}^{(k)} + \mathbf{R}^{-1}\mathbf{N_{jQ}}^{(k)}
\end{aligned}$$

This is the output of the decorrelator on each receiver. The data and pilots are MAI-free, however, the variance of the noise vector has increased.

The pilots are then used to estimate the different channel gains, to correct the phase shift introduced by the channel as well as to weight the different outputs for maximal ratio combining [7]. However, in the absence of noise, the received pilot streams would be slowly varying, thus they can be input to a narrow lowpass filter to greatly reduce the effect of noise. This greatly improves the mean square error between the channel estimates and the actual channel gains.

For each user, the complex data components at the output of the decorrelators are multiplied by the complex conjugates of the filtered received pilot streams. This corrects for the phase shifts caused by the channels weights the different signal components for combination. The weighted data components for a given user from each receiver are then added together to produce the final decision variable.

3. SIMULATED RESULTS

The bit error rate performance of our system is determined by Monte Carlo simulation. For each user, the signal is transmitted over M channels to M different antennas. The fading processes over these M different channels may or may not be correlated, depending on the spacing between the receive antennas. In this paper, we assume that the fading on the different transmission channels has a correlation that is either 0 (uncorrelated) or 0.5 (heavily correlated). We assume that the interfering users are randomly located in the cell and, as such, the fading processes encountered by the interfering users are independent of the fading processes encountered by the desired user. The Doppler spread of the fading process used in this paper, normalized to the symbol rate, is $B_d T_b = 0.1$, where B_d is the Doppler spread of the channel. The fading processes are generated by an autoregressive model to provide the proper U-shaped spectrum with the desired Doppler spread [6].

For simplicity, we assume that all users' signals are received with the same power. However, since we are examining Rayleigh fading channels, the received power of the different signals fluctuate greatly.

The cross-correlations between the spreading waveforms of the different users in the simulations are 0.15 or 0.25 which are typical of DS-CDMA systems with spreading factors of DS-CDMA systems with spreading factors of 40, and 16 respectively. The correlation between the spreading codes used for a specific user's data signal and its' pilot signal is always 0.

We wish to optimize the power allocated to the pilot. The total transmitted power for any user is $(A_{di}^2+A_{pi}^2)/2$, and so the average energy per bit of user i is $E_{bi} = (A_{di}^2+A_{pi}^2)T_b/2$, although the actual energy per data bit is $A_{di}^2 T_b/2$. Increasing A_{di} allows the receiver to better estimate the channel gains, but requires that the transmitter transmit more power to achieve the same BER compared to ideal case when the channel phases and gains are known to the receiver. In [6], it is shown that for these parameters, allocating 15%-25% of the total power to the pilot stream provides the lowest BER for a wide range of E_b/N_o for the single user system. In this paper, we allocate 15% of the total transmitted power to the pilot stream.

The lowpass filters used to filter the received pilot streams, to remove the out of band noise, are simple brick wall lowpass filters whose cutoff frequencies are slightly greater than the Doppler spread. This is not the optimum filter, but optimum filtering is impractical in real systems.

Figure 3-4 show the performance of 1, 2 and 4 user systems when the receiver has only one antenna. Figure 3 is for when the signature waveforms have a cross-correlation of 0.15 Figure 4 is for when the cross-correlation between the waveforms is 0.25.

We see from these figures that as we increase the waveform correlation (or decrease the spreading factor, the degradation caused by multiple users is more pronounced. This is due to the fact that the noise enhancement caused by decorrelation increases with the cross-correlations of the signature waveforms.

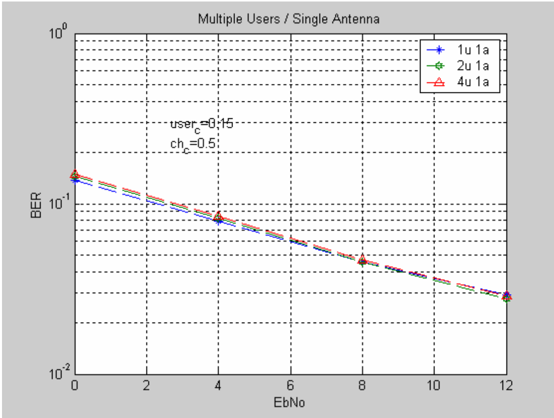

Figure 3: BER Performance of CDMA System with Single Antenna Receiver with Signature Waveform Cross-Correlations of 0.15.

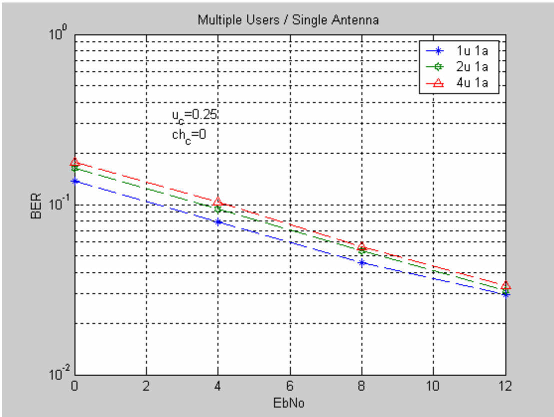

Figure 4: BER Performance of CDMA System with Single Antenna Receiver with Signature Waveform Cross-Correlations of 0.25.

Figures 5-6 show the CDMA system with signature waveforms cross-correlations of 0.15 and 0.25 respectively with a three antenna system for users encountering uncorrelated fading between different antenna elements. We note an increased slope in the BER curves. This is due to the diversity achieved by the multiple antennas.

Figures 7-8 show the BER performance of the CDMA system for signature waveform cross-correlations of 0.15 and 0.25 with a three antenna receiver when the correlation between the fading processes on adjacent antennas is 0.5. We note, that when compared with Figures 5-6 that the slope is less steep in the curves in Figures 7 and 8 since diversity is lost when the fading processes are correlated. In other words, the use of many antennas may not necessarily outperform the use of a few optimally spaced antennas. It should be noted, that for the results shown in this paper, the three antenna receiver with heavily correlated fading still outperforms the single antenna receiver.

Additional results of this study can be found in [6].

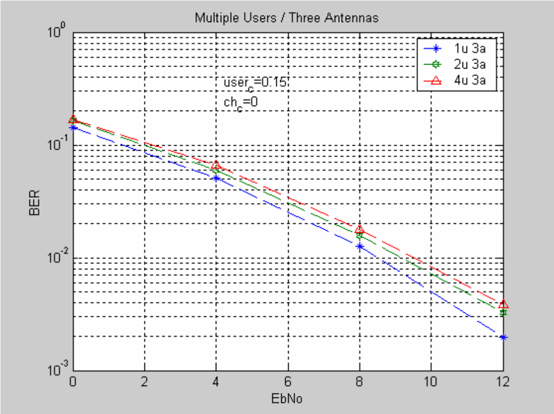

Figure 5: BER Performance of CDMA System with Three Antenna Receiver with Signature Waveform Cross-Correlations of 0.15.

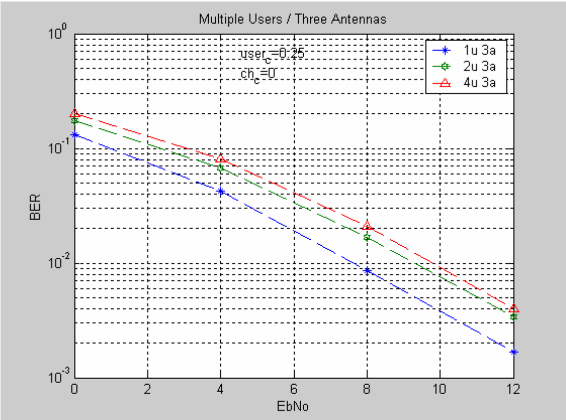

Figure 6: BER Performance of CDMA System with Three Antenna Receiver with Signature Waveform Cross-Correlations of 0.25.

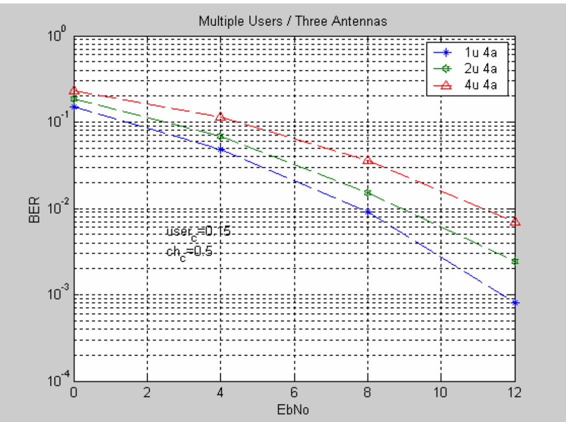

Figure 7: BER Performance of CDMA System with Three Antenna Receiver with Signature Waveform Cross-Correlations of 0.15 and Signals Encountering Correlated Fading Processes.

4. CONCLUSIONS

In this paper, we discuss a possible strategy for channel estimation for multiantenna receivers which eliminates MAI and efficiently combines the signal components from the different antenna elements. We see that the BER performance of the system improves significantly when we use additional antennas, even when the fading is highly correlated.

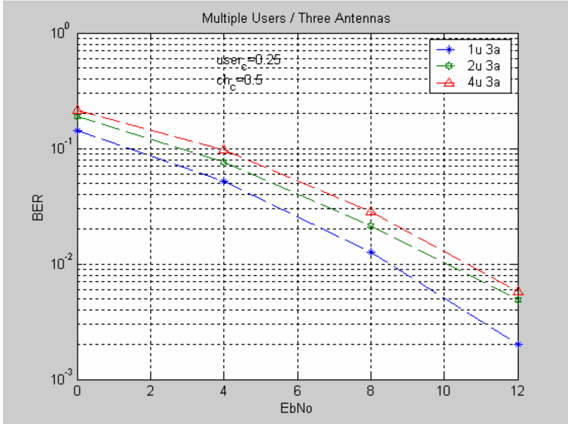

Figure 8: BER Performance of CDMA System with Three Antenna Receiver with Signature Waveform Cross-Correlations of 0.25 and Signals Encountering Correlated Fading Processes.

Decorrelation was used mainly to remove MAI from the pilots to improve channel estimation. This separated the multiple users before any space processing is applied. However, minimum mean square error (MMSE) multiuser detectors perform better than decorrelators, yet residula MAI remains. This residual MAI would be additional suppressed by the multiple antennas due to the spatial separation of the different users. Thus a multiantenna receiver using MMSE multiuser detectors would theoretically outperform our system. However, MMSE receivers require knowledge of the amplitudes of the different received signals, which are found from the pilot signals. Thus rough estimates would need to be made, and then refined iteratively if we were to use MMSE detectors. Thus our system provides a good tradeoff between performance and complexity.

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