LETTER

Coherent Detection for DS/CDMA System with M-Ary Orthogonal Modulation in Multipath Fading Channels

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SUMMARY In this letter, we propose the channel estimation technique in the uplink of a DS/CDMA system with M-ary orthogonal signaling over multipath fading channels. The channel estimation is carried out using the maximum correlator output of RAKE receiver. With the estimated channel parameters, the RAKE receiver can coherently combines the received multipath signals, resulting in a significant performace improvement. The hardware complexity of the proposed detection technique is slightly increased when compared to that of noncoherent detection.
key words: *M-ary orthogonal modulation, DS/CDMA, channel estimation, coherent detection*

1. Introduction

In a direct sequence code division multiple access (DS/CDMA) communication system, a number of DS/CDMA signals share the same frequency spectrum to provide communication between a number of users. In general, in order to increase the system capacity, a coherent demodulation is preferable. However, on the reverse link of a realistic DS/CDMA system (for example, IS-95 system), a coherent demodulation is impractical since it requires the inclusion of a pilot signal for each user. Instead, a noncoherent reception is used. M-ary orthogonal modulation is a spectrum efficient modulation scheme well suited for this application [1], [2].

In this letter, we present a simple method of channel estimation in a DS/CDMA system with M-ary orthogonal modulation over multipath fading channels without the inclusion of a pilot signal. Therefore a coherent demodulation is possible and an average bit error probability is significantly improved. The channel estimation is carried out using the maximum correlator output of RAKE receiver. But, the total hardware complexity of the receiver is slightly increased when compared to that of noncoherent detection. We analyze the performance of the proposed coherent detection technique and compare with that of a conventional noncoherent RAKE receiver. We also present the simulation results to verify the performance of the proposed coherent detection technique.

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2. Coherent Detection for DS/CDMA System with M-Ary Orthogonal Modulation

2.1 System Model

The system under consideration uses a M-ary orthogonal signal set $\{W^j(t)\}$. Assuming the j-th M-ary orthogonal (Walsh) symbol $W^j(t)$ is transmitted, the received signal is the sum of all users' signals plus noise and given by

$$r(t) = \sum_{i=1}^{K}\sum_{n=1}^{N}\sqrt{P_w}\alpha_n^i\Big[\Psi_I^{i,j}(t-\tau_n^i)\cos(w_ct-\theta_n^i) + \Psi_Q^{i,j}(t-T_0-\tau_n^i)\sin(w_ct-\theta_n^i)\Big]+n(t) \quad (1)$$

where K is the number of users and N is the number of multipath components. T_0 is the offset time, $\theta_n^i = \phi_n^i + w_c\tau_n^i$, P_w is the transmitted power per Walsh symbol, and $a_I^i(t)$ and $a_Q^i(t)$ are the spreading waveforms of the in-phase and quadrature channels for the i-th user, respectively. $\Psi_I^{i,j}(t) = a_I^i(t)W^j(t)$, $\Psi_Q^{i,j}(t) = a_Q^i(t)W^j(t)$ and w_c is the carrier angular frequency. α_n^i, ϕ_n^i, and τ_n^i are the amplitude, phase, and delay of the n-th multipath, respectively, and $n(t) = Re[\{n_c(t)+jn_s(t)\}e^{jw_ct}]$ is a zero mean Gaussian noise process with spectral density $N_0/2$.

Figure 1 shows the proposed receiver structure, which includes a channel estimation block and a phase recovery block. The output of the lowpass filter of the I-channel is given by [2]

$$d_I(t) = \sum_{i=1}^{K}\sum_{n=1}^{N}\sqrt{P_w}\alpha_n^i\left[\Psi_I^{i,j}(t-\tau_n^i)\frac{\cos(\theta_n^i)}{2} + \Psi_Q^{i,j}(t-T_0-\tau_n^i)\frac{\sin(\theta_n^i)}{2}\right] + \frac{n_c(t)}{2} \quad (2)$$

Similary, $d_Q(t)$ can be obtained. Let the k-th user

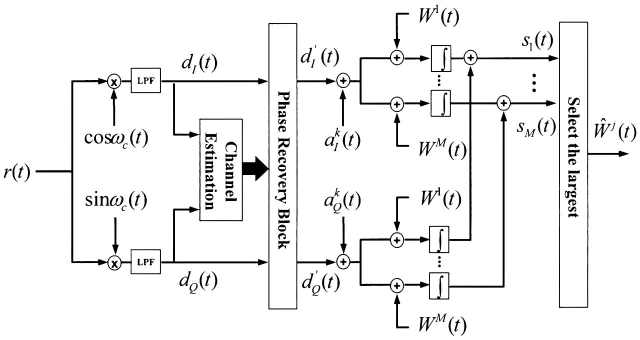

Fig. 1 Receiver block for the l-th RAKE branch with phase recovery.

be the user of interest. The output of the phase recovery block in the l-th RAKE branch is given by

$$d^{'}(t) = [d_I(t) + jd_Q(t)]\left[w_{I,l}^k - jw_{Q,l}^k\right] \quad (3)$$

where $w_{I,l}^k$ and $w_{Q,l}^k$ is the inphase and quadrature componet of channel estimator in the l-th branch of the k-th user's RAKE receiver, respectively. We assume that the estimation of the channel parameter is perfect, that is , $w_l^k = w_{I,l}^k + jw_{Q,l}^k = \alpha_{n_l}^k e^{j\theta_{n_l}^k}$. In contrast to [2], since the phase recovery is performed in our proposed receiver, we do not need to detect the energy of $d_I^{'}(t)$ against the quadrature spreading waveform $a_Q^k(t)$. Thus, the hardware complexity of each l-th branch of the RAKE receiver is reduced to the half of noncoherent one as shown in Fig. 1.

After combining L branches, a decision variable of the k-th user becomes $s^k(m) = \sum_{l=1}^{L} s^k(m,l)$, where $s^k(m,l) = {Z^{'}}_{II}^k(m,l) + {Z^{'}}_{QQ}^k(m,l)$. The output of the m-th correlator and l-th receiver branch, ${Z^{'}}_{II}^k(m,l)$ and ${Z^{'}}_{QQ}^k(m,l)$, are defined as

$${Z^{'}}_{II}^k(m,l) = \frac{1}{\sqrt{T_w}}\int_{\tau_{n_l}^k}^{T_w+\tau_{n_l}^k} d_I^{'}(t)\Psi_I^{k,m}(t-\tau_{n_l}^k)dt \quad (4)$$

$${Z^{'}}_{QQ}^k(m,l) = \frac{1}{\sqrt{T_w}}\int_{\tau_{n_l}^k}^{T_w+\tau_{n_l}^k} d_Q^{'}(t)\Psi_Q^{k,m}(t-\tau_{n_l}^k)dt \quad (5)$$

where T_w is the duration of a Walsh symbol. We assumed that the delay time of the l-th tracked multipath component, $\tau_{n_l}^k$, is estimated perfectly. Therefore, the correlator output is given by

$${Z^{'}}_{II}^k(m,l) = \begin{cases} \frac{(\alpha_{n_l}^k)^2\sqrt{E_w}}{2} + ICI_I^k(l) + MAI_I^{k,i}(l) + N_I^k(l) & m = j \\ ICI_I^k(l) + MAI_I^{k,i}(l) + N_I^k(l) & m \neq j \end{cases} \quad (6)$$

where E_w is the energy of a Walsh symbol, $E_w = P_w T_w$, $ICI_I^k(l)$ is the self-interference to the l-th branch of the RAKE receiver due to multipath, $MAI_I^{k,i}(l)$ is the MAI, and $N_I^k(l)$ is the term due to the presence of AWGN. Since ICI, MAI and $N_I^k(l)$ can be modelled as Gaussian random vaiables, it can be easily shown that ${Z^{'}}_{II}^k(m,l)$ is a Gaussian random variable with variance given by

$$Var[{Z^{'}}_{II}^k(m,l)] = \left(\alpha_{n_l}^k\right)^2\left\{\frac{N_0}{4} + \frac{E_w}{\rho N_c} \times \left[\sum_{n=1,n\neq n_l}^{N} E\left[\left(\alpha_n^k\right)^2\right] + \sum_{i=1,i\neq k}^{K}\sum_{n=1}^{N} E\left[\left(\alpha_n^i\right)^2\right]\right]\right\} \quad (7)$$

where $N_c = T_w/T_c$, T_c is the chip duration, $\rho = 6$ for asynchronous transmission and $\rho = 4$ for synchronous. $E[\cdot]$ denotes expectation. Similarly, we can obtain ${Z^{'}}_{QQ}^k(m,l)$.

2.2 Average Bit Error Probability

For a fixed set of $\{\alpha_n^k\}$, $s^k(m,l)$ becomes a Gaussian random variable with variance $2\cdot Var[{Z^{'}}_{II}^k(m,l)]$ since ${Z^{'}}_{II}^k(m,l)$ and ${Z^{'}}_{QQ}^k(m,l)$ are Gaussian random variables with the same variance. In realistic situations, the random variables $s^k(m,l)$ are correlated due to interchip interference. However, since ICI is relatively small as compared to the MAI, we can assume that the correlator outputs $s^k(m,l)$ are uncorrelated for large K [2]. Thus, the decision variable of the k-th user, $s^k(m)$, becomes a Gaussian random variable with mean

$$E\left[s^k(m)\right] = \begin{cases} \lambda = \sqrt{E_w}\sum_{l=1}^{L}(\alpha_{n_l}^k)^2 & m = j \\ 0 & m \neq j \end{cases} \quad (8)$$

and variance as

$$\sigma^2 = 2\sum_{l=1}^{L} Var[{Z^{'}}_{II}^k(m,l)] \quad (9)$$

To derive the probability of symbol error, without loss of generality, suppose that the signal $W^1(t)$ was transmitted. We assume that $s^k(m)$ $(m = 2, 3, \cdots, M)$ are identical and independent distributed random variables. Thus, the probability that a symbol will be decoded correctly is

$$P_c(\lambda) = \int_0^{\infty}\left[1 - Q\left(\frac{s_1}{\sigma^2}\right)\right]^{M-1}\frac{1}{\sqrt{2\pi\sigma^2}}e^{-\frac{(s_1-\lambda)^2}{2\sigma^2}}ds_1 \quad (10)$$

where $Q(\cdot)$ is the complementary error function. $P_c(\lambda)$ is conditioned on the random variable λ. Finally, the

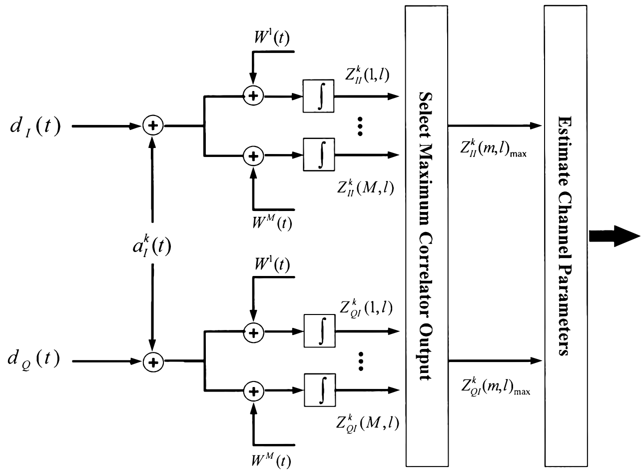

Fig. 2 Noncoherent channel estimation block for the l-th RAKE branch.

symbol error probability becomes $P_M(\lambda) = 1 - P_c(\lambda)$, and the bit error probability becomes [2], [3]

$$P_b(\lambda) = \frac{2^{h-1}}{M-1} P_M(\lambda) \tag{11}$$

where $h = \log_2 M$. The bit error probability $P_b(\lambda)$ is also conditioned on the random variable λ, which in turn depends on the multipath channel amplitudes $\{\alpha_n^k\}$. Hence the average bit error probability depends on the actual probability density function $p(\lambda)$ and is given by

$$\overline{P_b} = \frac{2^{h-1}}{M-1} \int_0^{\infty} P_M(\lambda) p(\lambda) d\lambda \tag{12}$$

However, $p(\lambda)$ is not known in general.

2.3 Channel Estimator

The proposed estimation method is based on the idea that channel parameters can be extracted from the maximum correlator output in a receiver block. Figure 2 shows the proposed estimation block. In Fig. 2, after multiplying $a_I^k(t-\tau_{n_l}^k)$ to $d_I(t)$, and correlating with the m-th Walsh function, $Z_{II}^k(m,l)$ is given by [2]

$$Z_{II}^k(m,l) = \begin{cases} \alpha_{n_l}^k \frac{\sqrt{E_w} cos\theta_{n_l}^k}{2} + I_{Q,II}^{k,k}(l) + I_{II}^{k,k}(l) + I_{II}^{k,i}(l) + N_{II}^k(l) & m = j \\ I_{Q,II}^{k,k}(l) + I_{II}^{k,k}(l) + I_{II}^{k,i}(l) + N_{II}^k(l) & m \neq j \end{cases} \tag{13}$$

where all notations are the same with [2]. $I_{Q,II}^{k,k}(l)$, $I_{II}^{k,k}(l)$, $I_{II}^{k,i}(l)$, and $N_{II}^k(l)$ are all zero-mean Gaussian random variables and each variance is given in [2]. From (13), we can see that $Z_{II}^k(m,l)$ consists of a real term of channel coefficient and AWGN for $m = j$. If we select the largest value, $Z_{II}^k(m,l)_{max}$, out of $Z_{II}^k(m,l)$, $Z_{II}^k(m,l)_{max}$ can be regarded as sum of a real term of channel coefficient and AWGN. In fact, this implies the case $m = j$, which occurs with a high probability at a moderate signal power to noise power ratio (SNR). Therefore, we can design an unbiased estimator based on $Z_{II}^k(m,l)_{max}$. Furthermore, fading channel parameters may be represented as a *first-order Gauss-Markov process* [4], i.e., the channel parameter to be estimated are correlated with the value of previous channel parameter. Therefore, we can estimate the real term of the channel coefficients as follows

$$\mathrm{Re}[w_{l,t}^k] = \beta \mathrm{Re}[w_{l,t-1}^k] + (1-\beta) \frac{2}{\sqrt{E_w}} Z_{II}^k(m,l)_{max} \tag{14}$$

where $w_{l,t}^k$ is the estimated complex channel parameter in the l-th branch of the k-th user's RAKE receiver at time t, and β is the update factor. Similarly, we can estimate the imaginary term of channel coefficient from $Z_{QI}^k(m,l)_{max}$.

$$\mathrm{Im}[w_{l,t}^k] = \beta \mathrm{Im}[w_{l,t-1}^k] + (1-\beta) \frac{2}{\sqrt{E_w}} Z_{QI}^k(m,l)_{max} \tag{15}$$

3. Results and Discussions

To evaluate the performance of the proposed channel estimation technique, the extensive series of Monte-Carlo simulations have been conducted in frequency-selective fading channel using the parameters specified in IS-95 system with (1) Number of diversity branches: $N = L = 3$ and (2) Doppler frequency: $f_d = 30$ Hz.

For the frequency-selective fading channel, we assume that the envelope of the each channel tap coefficient is Rayleigh distributed and each tap has an equal power, resulting in $\sum_{n=1}^{N} (\alpha_n^i)^2 = 1$. For this case, we can numerically evaluate the average bit error probability of (12). Since we assme that $\sum_{n=1}^{N} (\alpha_n^i)^2 = 1$ and α_n^i is identical and independent distributed with the same variance $\sigma_\alpha^2 = 1/2N$, the random variable λ in (8) has a chi-square distribution with L degrees of freedom given by [3]

$$p(\lambda) = \frac{1}{\sigma_\alpha^{2L} 2^L \Gamma(L)} \lambda^{L-1} e^{-\lambda/2\sigma_\alpha^2} \tag{16}$$

The $Var[Z'^k_{II}(m,l)]$ becomes as follows

$$Var[Z'^k_{II}(m,l)] = (\alpha_{n_l}^k)^2 \left\{ \frac{N_0}{4} + \frac{E_w}{4N_0} \times \left[\frac{N-1}{N} + (K-1) \right] \right\} \tag{17}$$

Thus, for $N = L$, the average bit error probability $\overline{P}_b$ can be calculated as follows

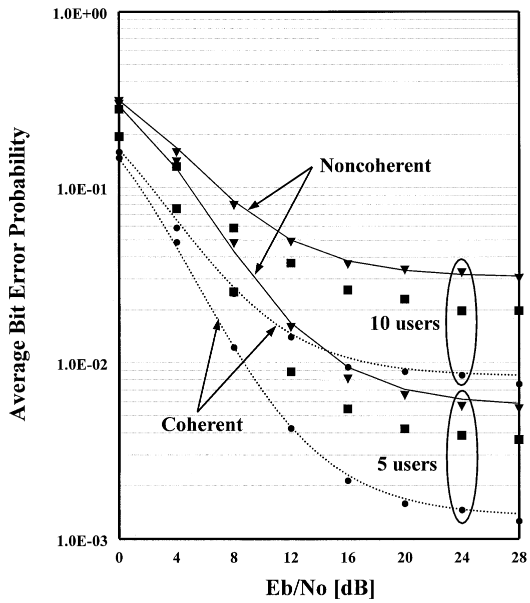

Fig. 3 Average bit error probability for $K = 5$ and 10. (1) solid line: Noncoherent detection (analytical), (2) dotted line: Coherent detection (analytical), (3) triangle-down($\bigtriangledown$): Noncoherent detection (simulation), (4) rectangluar($\square$): Coherent detection with the estimated value (simulation), (5) circle($\circ$): Coherent detection with the no estimation error (simulation).

$$\overline{P}_b = \frac{2^{h-1}}{M-1} \int_0^\infty (1 - P_c(\lambda)) \frac{N^N}{\Gamma(N)} \lambda^{N-1} e^{-\lambda N} d\lambda \tag{18}$$

Some numerical results of average bit error probability are shown in Fig. 3. The solid lines and the dotted lines are the numerical results of noncoherent and coheren detection when 5 and 10 users exist, respectively. We can see that a considerable improvement can be achieved by coherent detection in terms of the average bit error probability.

All points in Fig. 3 are obtained from the computer simulations. The numerical results of the proposed coherent detection are very close to the simulation

Table 1 Variances of estimation error for various values of β ($E_b/N_0 = 20\,\text{dB}$).

β	0.3	0.5	0.7	0.9
Variance	0.006475	0.004509	0.002611	0.008136

results. For the coherent detection with the estimated value, the value of β was set to 0.7, which is choosen in heuristic manner from several simulations. The variances of the estimation error for various value of β are shown in Table 1 and is minimum when $\beta = 0.7$. Since β is sensitive to the channel characteristics, the large value of β is good in a slow fading channel. On the other hand, a small value is appropriate in a fast fading channel. We can insist from Fig. 3 that the proposed detection method with phase recovery outperforms a conventional noncoherent detection.

As a conclusion, it is shown that the estimation of channel parameters in a noncoherent DS/CDMA system with M-ary orthogonal signal is possible by using a maximum correlator output. We can also obtain performance enhancement by coherent detection with phase recovery. The hardware complexity of the proposed detection technique is slightly increased when compared to that of noncoherent detection. However, as the number of simultaneous users increases, the channel estimation may be deteriorated. This can be circumvented by combining the proposed channel estimator with multiuser detector.

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