Constrained MMSE Receivers for CDMA Systems in Frequency-Selective Fading Channels

Seung-Chul Hong, Jinho Choi, Young-Ho Jung, Seong Rag Kim, and Yong H. Lee

***Abstract*—A constrained minimum mean square error (CMMSE)-RAKE receiver for multipath fading channels is developed by extending the CMMSE receiver for flat fading channels. Based on the observation that interpath interference causes a bias of the channel estimator in [8], a receiver that can remove such a bias is proposed, plus a closed-form expression of the bit-error rate of the receiver is derived. Computer simulation is used to demonstrate that the proposed receiver can outperform existing RAKE receivers.**

***Index Terms*—Frequency-selective fading channels, minimum mean square error (MMSE), multiuser detection.**

I. Introduction

ADAPTIVE minimum mean square error (MMSE) receivers for direct-sequence code-division multiple-access (DS/CDMA) systems are considered as a useful alternative to the multiuser detection schemes in [1]. In contrast to multiuser detectors that require spreading sequences for all users, the adaptive MMSE receivers in [2]–[8] only need the spreading sequence of the desired user. Consequently, these receivers are simpler to implement than multiuser detectors, yet they can provide a considerable performance improvement when compared with conventional DS/CDMA receivers.

Various adaptive MMSE receivers have already been proposed for the detection of DS/CDMA systems. For additive white Gaussian noise (AWGN) channels, adaptive MMSE receivers have been developed based on the standard MSE cost function [2]–[4]. In the case of flat-fading channels, a channel estimator has been employed and, in an attempt to improve the tracking capability, the MSE cost is modified [5]–[8]. Furthermore, in [8], a constraint regarding filter coefficients is imposed on the MMSE problem, and it has been shown that the resulting receiver, called the constrained MMSE (CMMSE) receiver, can outperform the other MMSE receivers. For frequency-selective fading channels, use of an adaptive filter for each resolvable transmission path has been suggested, and the receivers in [5]–[8] can be applied to each path. However, in this case, the receiver performance can be degraded due to interpath interference (IPI).

In this letter, we develop a CMMSE receiver for multipath fading channels. This receiver is a RAKE receiver that employs CMMSE receivers at each finger. It was observed that the direct application of the receiver in [8] to CMMSE-RAKE suffers from IPI, which causes a bias of the channel estimator. Therefore, in an attempt to remove such a bias, a CMMSE-RAKE receiver that employs an unbiased maximum-likelihood (ML) channel estimator is proposed, and its performance analyzed.

The organization of this letter is as follows. The system model is presented in Section II. The direct extension of the CMMSE receiver in [8] to multipath channels is considered in Section III. The proposed receiver is developed in Section IV and statistically analyzed in Section V. Finally, Section VI presents analytical and simulation results demonstrating the advantage of the proposed receiver over existing ones.

II. System Model

An asynchronous DS/CDMA system with K users and M propagation paths is considered. Phase-shift keying (PSK) modulated data of the kth user is spread by a normalized spreading code $\{c_k(j)\}$, where $c_k(j) \in \{-1/\sqrt{N}, 1/\sqrt{N}\}$ for $j = 0, 1, \ldots, N-1$ and $c_k(j) = 0$, otherwise, and N is the number of chips per symbol. It is assumed that the attenuation factors (or channel coefficients) are uncorrelated to each other and that the propagation delay of the kth user's mth path, denoted by $\tau_{k,m}$, is a multiple of the chip duration T_c. The received signal is passed through a filter matched to the chip pulse shape and sampled at the chip rate to yield a received sequence denoted by $\{y_j\}$. Suppose, without loss of generality, that the first user is the user of interest and that $\tau_{1,m} = (m-1)T_c$. The MMSE receivers that are considered estimate the nth transmitted symbol $d_1(n)$ from $\{y_{nN}, y_{nN+1}, \ldots, y_{(n+1)N+M-2}\}$, where y_{nN+l} is given by

$$y_{nN+l} = \sum_{m=1}^{M} d_1(n)h_{1,m}(n)c_1(l-(m-1)) + u(nN+l) \tag{1}$$

where M is the number of multipaths, $h_{1,m}(n)$ is the complex attenuation factor of the first user's mth path, and $u(n)$ denotes the sum of the intersymbol interference, multiple access interference, and background noise. In vector form, $\mathbf{y}(n) = [y_{nN}\ y_{nN+1}\ \cdots\ y_{(n+1)N+M-2}]^T = \sum_{m=1}^{M} d_1(n)h_{1,m}(n)\mathbf{c}_{1,m} + \mathbf{u}(n)$ where $\mathbf{c}_{1,m} = [\mathbf{0}_{m-1}^T\ \mathbf{c}_1^T\ \mathbf{0}_{M-m}^T]^T$, $\mathbf{c}_1 = [c_1(0)\ \cdots\ c_1(N-1)]^T, \mathbf{0}_m$ is an $m \times 1$ vector with all zero elements, and

Manuscript received April 24, 2002; revised April 22, 2003; accepted May 16, 2003. The editor coordinating the review of this paper and approving it for publication is A. Yener. This work was supported in part by Korea Science and Engineering Foundation. This paper was presented in part at the IEEE Vehicular Technology Conference, Birmingham, AL, May, 2002.

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Digital Object Identifier 10.1109/TWC.2004.833414

$\mathbf{u}(n) = [u(nN) \cdots u((n+1)N+M-2)]^T$. Dropping the subscript indicating the desired user for notational simplicity, $\mathbf{y}(n)$ can be rewritten as

$$\mathbf{y}(n) = \sum_{m=1}^{M} d(n)h_m(n)\mathbf{c}_m + \mathbf{u}(n) \tag{2}$$

$$= \mathbf{C}\mathbf{h}(n)d(n) + \mathbf{u}(n) \tag{3}$$

where $\mathbf{C}$ is a $(N+M-1)\times M$ code matrix given by $\mathbf{C} = [\mathbf{c}_1\ \mathbf{c}_2\ \cdots\ \mathbf{c}_M]$ and $\mathbf{h}(n) = [h_1(n)\ h_2(n)\ \cdots\ h_M(n)]^T$.

III. Basic CMMSE Receivers in Frequency-Selective Fading Channels

For frequency-selective channels, it is natural to employ an adpative CMMSE filter for each resolvable path and combine the resulting outputs. In this receiver, $d(n)$ is estimated by

$$\hat{d}(n) = \sum_{m=1}^{M} \hat{h}_m^*(n)\mathbf{w}_m^H(n)\mathbf{y}(n) \tag{4}$$

where $\hat{h}_m(n)$ denotes the channel estimates, and the weight of the mth filter $\mathbf{w}_m(n)$ is obtained based on the following optimization:

$$\text{minimize}_{\mathbf{w}_m} E\left[\left|h_m(n)d(n) - \mathbf{w}_m^H\mathbf{y}(n)\right|^2\right]$$
$$\text{subject to } \mathbf{c}_m^H\mathbf{w}_m = 1. \tag{5}$$

The optimal weight, $\mathbf{w}_m^o$, that satisfies (5) is given by

$$\mathbf{w}_m^o = \mathbf{R}_\mathbf{y}^{-1}\mathbf{c}_m\left(\mathbf{c}_m^H\mathbf{R}_\mathbf{y}^{-1}\mathbf{c}_m\right)^{-1} \tag{6}$$

where

$$\mathbf{R}_\mathbf{y} = E[\mathbf{y}(n)\mathbf{y}^H(n)] = \sum_{m=1}^{M} E[|h_m(n)|^2]\mathbf{c}_m\mathbf{c}_m^H + \mathbf{R}_\mathbf{u} \tag{7}$$

and $\mathbf{R}_\mathbf{u} = E[\mathbf{u}(n)\mathbf{u}^H(n)]$. The adaptive receiver based on (5) will be referred to as the basic CMMSE-RAKE receiver[1]. The error signal for the mth path is given by $e_m(n) = \hat{h}_m(n)d(n) - \mathbf{w}_m^H(n)\mathbf{y}(n)$ and the filter weight is updated using

$$\mathbf{w}_m(n+1) = \mathbf{c}_m + \mathbf{P}_{\mathbf{c}_m}^{\perp}(\mathbf{w}_m(n) + \mu(n)e_m^*(n)\mathbf{y}(n)) \tag{8}$$

where $\mathbf{P}_{\mathbf{c}_m}^{\perp} = \mathbf{I} - \mathbf{c}_m(\mathbf{c}_m^H\mathbf{c}_m)^{-1}\mathbf{c}_m^H$ and $\mu(n) = (\rho)/(\mathbf{y}^H(n)\mathbf{y}(n))(0 < \rho < 2)$. In (8), $\mathbf{c}_m$ is fixed and serves as the initial weight vector. The channel is estimated by exploiting pilot symbols under the assumption that $\{h_m(n)\}$ are quasi-constant. If a pilot symbol is inserted at every Q_p symbol, then the channel estimate can be given by

$$\hat{h}_m(n) = \frac{1}{N_p}\sum_{i=0}^{N_p-1} d^*((\nu-i)Q_p)\mathbf{w}_m^H((\nu-i)Q_p) \times\mathbf{y}((\nu-i)Q_p) \tag{9}$$

[1]In [8], a CMMSE receiver for flat-fading channels is proposed. The basic CMMSE-RAKE receiver is a direct extension of the CMMSE receiver for multipath fading channels.

where $\nu = \lfloor(n/Q_p)\rfloor$, $d((\nu-i)Q_p)$ is the pilot and N_p is a positive integer. In Section VI, it will be shown through computer simulation that the basic CMMSE-RAKE receiver can outperform the receiver in [6]. Next, a property of this CMMSE-RAKE receiver is derived.

Property 1: The channel estimate in (9) is biased in multipath-fading channels ($M \geq 2$). It is only unbiased for flat-fading channels ($M = 1$).

Proof: Let $h_m(n) = h_m(n+1) = \cdots = h_m(n+N_p\cdot Q_p - 1) = h_m$. Since $|d(i)|^2 = 1, \mathbf{w}_m^H(i)\mathbf{c}_m = 1$ and the data for each user are uncorrelated, then from (9) we get

$$E[\hat{h}_m\,|\,h_m] = h_m + \frac{1}{N_p}\sum_{i=0}^{N_p-1}\sum_{\substack{m'=1\\m'\neq m}}^{M} h_{m'}\mathbf{w}_m^H((\nu-i)Q_p)\mathbf{c}_{m'}. \tag{10}$$

In the right-hand side (RHS) of this equation, the second term does not vanish unless $M = 1$, because in general $\mathbf{w}_m^H(i)\mathbf{c}_{m'} \neq 0$ for $m \neq m'$. ■

The bias of the channel estimator is caused by IPI and degrades the receiver performance. A technique that can remove this bias and thus improve the performance is presented next.

IV. Proposed Receiver

The proposed receiver is identical to the basic CMMSE-RAKE receiver with the exception that it employs an unbiased channel estimator instead of the one in (9). In the proposed receiver, all channels $\{h_1(n),\ldots,h_m(n)\}$ are estimated simultaneously from a set of signals $\{d^*(iQ_p)\mathbf{w}_m^H(iQ_p)\mathbf{y}(iQ_p), m = 1,\ldots,M\}$ (Fig. 1). Here $d(iQ_p)$ is the pilot symbol that is multiplied with the filter outputs to remove any data dependency. The resulting signal $d^*(iQ_p)\mathbf{w}_m^H(iQ_p)\mathbf{y}(iQ_p)$ can be modeled as a Gaussian process [9]. An $M\times 1$ observation vector is defined by $\mathbf{b}(i) = d^*(iQ_p)[\mathbf{w}_1^H\mathbf{y}(iQ_p)\ \mathbf{w}_2^H\mathbf{y}(iQ_p)\ \ldots\ \mathbf{w}_M^H\mathbf{y}(iQ_p)]^T$. The channels are estimated under the following assumptions:

1) $\{\mathbf{b}(i)\}$ are independent, identically distributed (i.i.d.) Gaussian random vectors;
2) $\{\mathbf{b}(n)\ \mathbf{b}(n-1)\ \cdots\ \mathbf{b}(n-N_p+1)\}$ are given for estimating $\mathbf{h} = [h_1\ h_2\ \cdots\ h_M]^T$;
3) $h_m(nQ_p) = h_m(nQ_p - 1) = \cdots = h_m((n-N_p+1)\cdot Q_p) = h_m$ for $m = 1,\ldots,M$;
4) $\mathbf{w}_m(nQ_p) = \mathbf{w}_m(nQ_p-1) = \cdots = \mathbf{w}_m((n-N_p+1)\cdot Q_p) = \mathbf{w}_m$ for $m = 1,\ldots,M$.

Then, the joint probability density function (pdf) of $\{\mathbf{b}(n)\ \mathbf{b}(n-1)\ \cdots\ \mathbf{b}(n-N_p+1)\}$ conditioned on $\mathbf{h}$ can be written as

$$f(\mathbf{b}\,|\,\mathbf{h}) = \prod_{i=0}^{N_p-1}\frac{1}{\pi^M\cdot\det(\mathbf{C}_\mathbf{b})} \times e^{-(\mathbf{b}(n-i)-\mathbf{m}_\mathbf{b})^H\mathbf{C}_\mathbf{b}^{-1}(\mathbf{b}(n-i)-\mathbf{m}_\mathbf{b})} \tag{11}$$

where $\det(\mathbf{C}_\mathbf{b})$ denotes the determinant of the matrix $\mathbf{C}_\mathbf{b}$; $\mathbf{m}_\mathbf{b}$ and $\mathbf{C}_\mathbf{b}$ are the conditional mean and covariance matrix of $\mathbf{b}$,

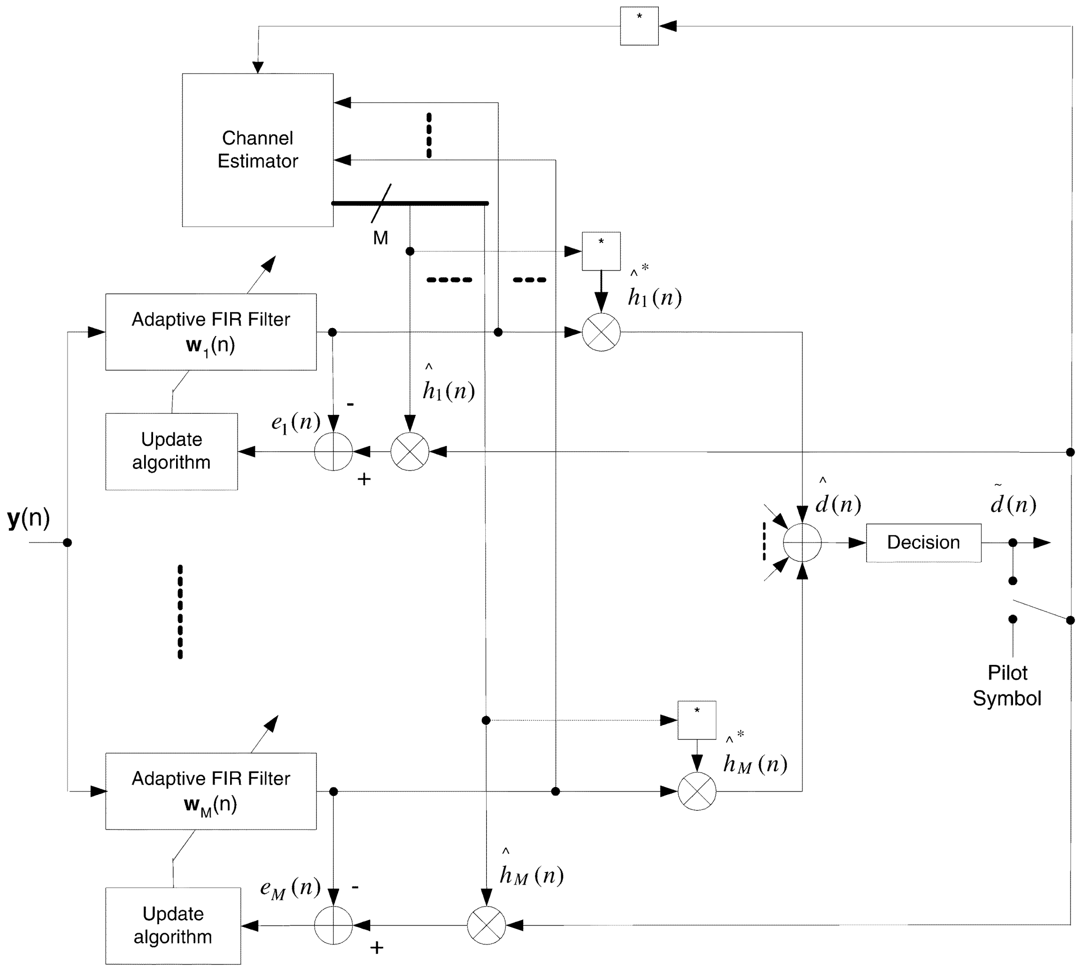

Fig. 1. Structure of proposed receiver.

respectively. They are expressed as $\mathbf{m_b} = E[\mathbf{b}(i) \,|\, \mathbf{h}] = \mathbf{Bh}$, where $\mathbf{B}$ is an $M \times M$ matrix whose (i,j)th entry is given by $\mathbf{w}_i^H \mathbf{c}_j$, and

$$\begin{aligned}\mathbf{C_b} &= E[(\mathbf{b}(i) - \mathbf{m_b})(\mathbf{b}(i) - \mathbf{m_b})^H \,|\, \mathbf{h}] \\ &= [\mathbf{w}_1 \quad \mathbf{w}_2 \quad \cdots \quad \mathbf{w}_M]^H \mathbf{R_u} [\mathbf{w}_1 \quad \mathbf{w}_2 \quad \cdots \quad \mathbf{w}_M]. \end{aligned} \tag{12}$$

The ML estimator that maximizes the pdf in (11) with respect to $\mathbf{h}$ is expressed as

$$\hat{\mathbf{h}} = \mathbf{B}^{-1} \cdot \frac{1}{N_p} \sum_{i=1}^{N_p} \mathbf{b}(i). \tag{13}$$

In general, the matrix $\mathbf{B}$ is nonsingular. This holds because $\{\mathbf{w}_1, \ldots, \mathbf{w}_M\}$, which are determined independently at the fingers, are linearly independent in most cases.

Following from $\mathbf{m_b} = \mathbf{Bh}$, the ML estimate in (13) is unbiased. However, this estimator requires a heavier computation than the one in (9): the number of multiplications for (13) is $(1/3)M^3 + ((3/2) + N)M^2 + ((7/6) + NN_p + N_p)M$, while that for (9) is $(NN_p + N_p)M$.

V. Bit-Error Rate (BER) Analysis

This section derives the BER expression under the following assumptions:

1) $\{h_m(n),\ m = 1, \ldots, M\}$ are time-invariant, that is, $h_m(n) = h_m$ for all m;
2) $\{h_m, m = 1, \ldots, M\}, \{\mathbf{u}(n)\}$ and $\{d(n)\}$ are independent random variables with zero mean, plus mutually independent;
3) $\mathbf{u}(n)$ is complex Gaussian [9], $|h_m(n)|$ is Rayleigh distributed, and $d(n) \in \{1, -1\}$;
4) channel estimation is perfect; in this case, the proposed receiver becomes identical to the basic CMMSE-RAKE receiver.
5) adaptive filters in the receiver are in the steady-state, and the filter weights are identical to the optimal weights in (6).

Referring to Fig. 1, the decision variable of the proposed receiver is given by

$$\hat{d}(n) = \sum_{m=1}^{M} h_m^* \mathbf{w}_m^H \mathbf{y}(n) \tag{14}$$

$$= \sum_{m=1}^{M} |h_m|^2 d(n) + \sum_{m=1}^{M} h_m^* \mathbf{w}_m^H \times \left(\sum_{\substack{j=1 \\ j \neq m}}^{M} \mathbf{c}_j h_j d(n) + \mathbf{u}(n) \right) \tag{15}$$

where the second equality comes from $\mathbf{w}_m^H \mathbf{c}_m = 1$. The first term of (15) represents the signal of interest, while the second term denotes the interference and noise, where $\sum_{j=1,j\neq m}^{M} \mathbf{c}_j h_j d(n)$ represents the IPI. When $\mathbf{h} = [h_1 \cdots h_M]^T$ is fixed, the signal power is given by

$$S_p(\mathbf{h}) = \left| \sum_{m=1}^{M} |h_m|^2 \right|^2 \tag{16}$$

and the interference and noise power can be written as

$$I_p(\mathbf{h}) = \sum_{m=1}^{M} \sum_{m'=1}^{M} \sum_{\substack{j=1\\ j\neq m}}^{M} \sum_{\substack{k=1\\ k\neq m'}}^{M} h_m^* h_{m'} h_j h_k^* \mathbf{w}_m^H \mathbf{c}_j \mathbf{c}_k^H \mathbf{w}_{m'} + \sum_{m=1}^{M} \sum_{m'=1}^{M} h_m^* h_{m'} \mathbf{w}_m^H \mathbf{R_u} \mathbf{w}_{m'}, \tag{17}$$

which results from the independence assumptions. The interference and noise power averaged over $\mathbf{h}$ is given by

$$E_{\mathbf{h}}[I_p(\mathbf{h})] = \sum_{m=1}^{M} E_{\mathbf{h}}[|h_m|^2] \left(\mathbf{w}_m^H \mathbf{R_y} \mathbf{w}_m - E_{\mathbf{h}}[|h_m|^2] \right) \tag{18}$$

where $\mathbf{R_y}$ is defined (7). The average signal-to-noise plus interference ratio (SINR), which is the ratio between $E_{\mathbf{h}}[S_p(\mathbf{h})]$ and $E_{\mathbf{h}}[I_p(\mathbf{h})]$, is expressed as

$$\text{SINR} = \frac{E_{\mathbf{h}}\left[\left| \sum_{m=1}^{M} |h_m|^2 \right|^2 \right]}{\sum_{m=1}^{M} E_{\mathbf{h}}[|h_m|^2] \left(\mathbf{w}_m^H \mathbf{R_y} \mathbf{w}_m - E_{\mathbf{h}}[|h_m|^2] \right)}. \tag{19}$$

The BER of the proposed receiver is derived following the approach in [10, p. 846]. For a fixed $\mathbf{h}$, the decision variable $\hat{d}(n)$ in (15) is Gaussian, and the bit-error probability is simply

$$P(\gamma_b) = Q(\sqrt{2\gamma_b}) \tag{20}$$

where $Q(\cdot)$ is the Q function [10, p. 40] and $\gamma_b = S_p(\mathbf{h})/I_p(\mathbf{h})$. If γ_b can be approximated as

$$\gamma_b \simeq \sum_{m=1}^{M} c_m \cdot |h_m|^2 \tag{21}$$

for c_m a constant, then from the fact that $\{|h_m|^2\}$ has a chi-squared distribution with two degrees of freedom, the pdf of γ_b can be given by

$$p(\gamma_b) = \sum_{m=1}^{M} \frac{\pi_m}{\bar{\gamma}_m} e^{-\gamma_b/\bar{\gamma}_m}, \quad \gamma_b \geq 0 \tag{22}$$

where $\bar{\gamma}_m = c_m \cdot E[|h_m|^2]$ and $\pi_m = \prod_{\substack{i=1\\ i\neq m}}^{M} (\bar{\gamma}_m)/(\bar{\gamma}_m - \bar{\gamma}_i)$. The BER averaged over γ_b is given by

$$P_e = \int_0^{\infty} P(\gamma_b) p(\gamma_b) d\gamma_b = \frac{1}{2} \sum_{m=1}^{M} \pi_m \left[1 - \sqrt{\frac{\bar{\gamma}_m}{1+\bar{\gamma}_m}} \right]. \tag{23}$$

Next, it will be shown that γ_b of the proposed receiver can be approximated as (21).

An approximation of γ_b, which is the SINR for a fixed $\mathbf{h}$, can be obtained by dropping the "expectation" from (19)

$$\gamma_b \simeq \frac{\left| \sum_{m=1}^{M} |h_m|^2 \right|^2}{\sum_{m=1}^{M} |h_m|^2 \left(\mathbf{w}_m^H \mathbf{R_y} \mathbf{w}_m - |h_m|^2 \right)}. \tag{24}$$

To further simplify γ_b, the following relations are assumed:

$$\mathbf{w}_m^H \mathbf{R_y} \mathbf{w}_m - |h_m|^2 = \mathbf{w}_m^H \mathbf{R_u} \mathbf{w}_m \tag{25}$$

$$\frac{\sum_{m=1}^{M} |h_m|^2 \mathbf{w}_m^H \mathbf{R_u} \mathbf{w}_m}{\sum_{m=1}^{M} |h_m|^2} = \frac{1}{M} \sum_{m=1}^{M} \mathbf{w}_m^H \mathbf{R_u} \mathbf{w}_m \triangleq \bar{\epsilon}. \tag{26}$$

The equality in (25) only holds when $M = 1$, while (26) holds when $\{|h_m|^2, m = 1, \ldots, M\}$ are the same. Under the assumptions in (25) and (26), the RHS of (24) can be rewritten as

$$\gamma_b \simeq \sum_{m=1}^{M} \frac{1}{\bar{\epsilon}} \cdot |h_m|^2. \tag{27}$$

Thus, the approximated BER of the proposed receiver can be obtained using (27) in (20)–(23).

VI. Simulation Results

For simulation, an asynchronous CDMA system with binary phase shift keying (BPSK) modulation was considered. Gold code with a length of 31 ($N = 31$) was employed. The data rate ($1/T$) was 64 kbps. A frequency selective Rayleigh-fading channel with an exponentially decaying power delay profile was assumed and the transmission powers of all active users were equal. The carrier frequency was 2.0 GHz [root mean squared (rms) delay spread (τ_{rms}) was 1 μs] and the vehicle speed was fixed at 100 km/h, corresponding to $f_d T = 0.00289$, where f_d was the maximum Doppler shift. The time delay of the kth user's mth path signal was

$$\tau_{k,m} = \tau_{k,1} + (m-1)T_c \quad \text{for } m = 1, \ldots, M. \tag{28}$$

Here, $\tau_{k,1}$ was randomly chosen from $\{0, T_c, 2T_c, \ldots, (N-1)T_c\}$. The variance of the channel parameter was given by $E[|h_{k,m}(n)|^2] = (1/q) \times e^{-(m-1)T_c/\tau_{\text{rms}}}$ where q was the normalization factor defined by $q = \sum_{m=1}^{M} e^{-(m-1)T_c/\tau_{\text{rms}}}$. The channel parameters $\{h_{k,m}(n)\}$ were generated following the approach in [11].

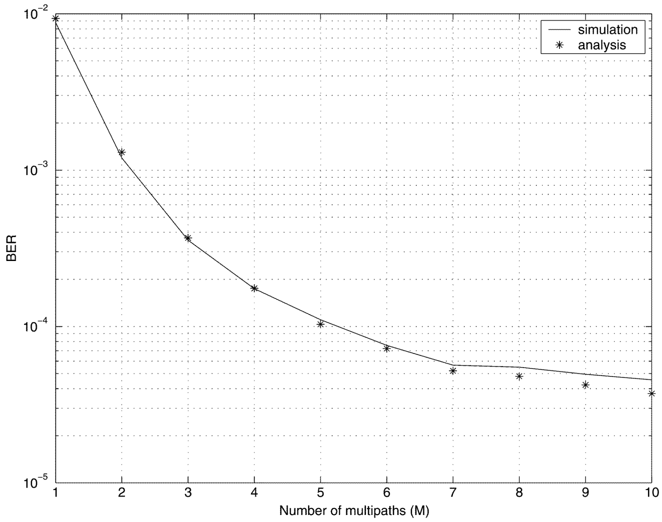

Fig. 2. BER versus number of multipaths when $K = 3$ and $E_b/N_0 = 15$ dB.

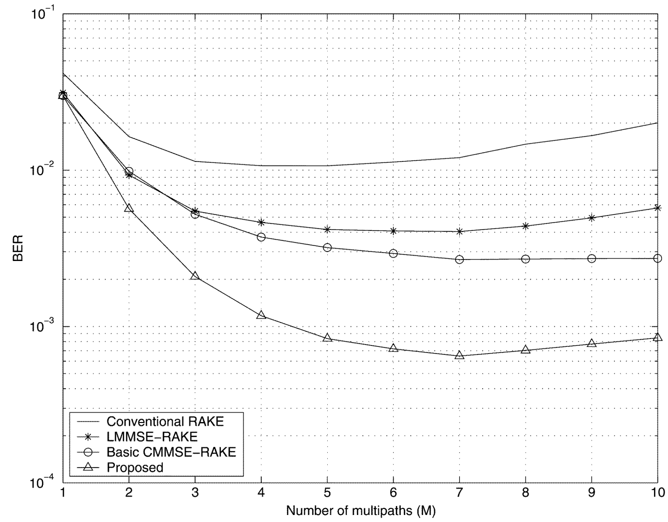

Fig. 4. BER versus number of multipaths when $K = 3$ and $E_b/N_0 = 15$ dB (channel estimation was employed).

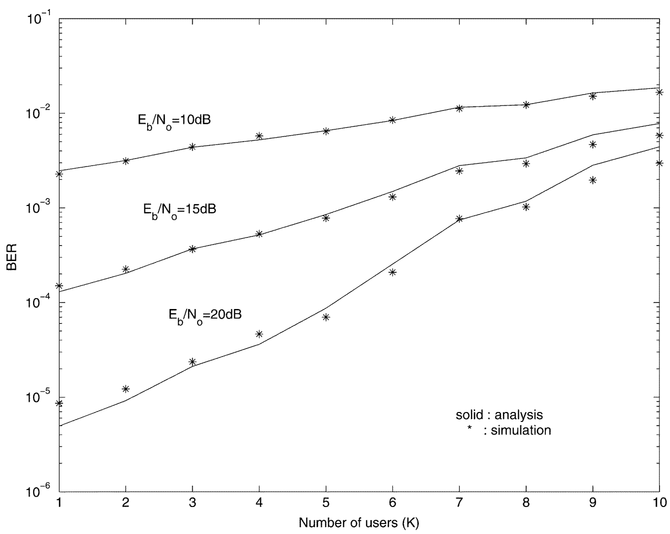

Fig. 3. BER versus number of users when $M = 3$.

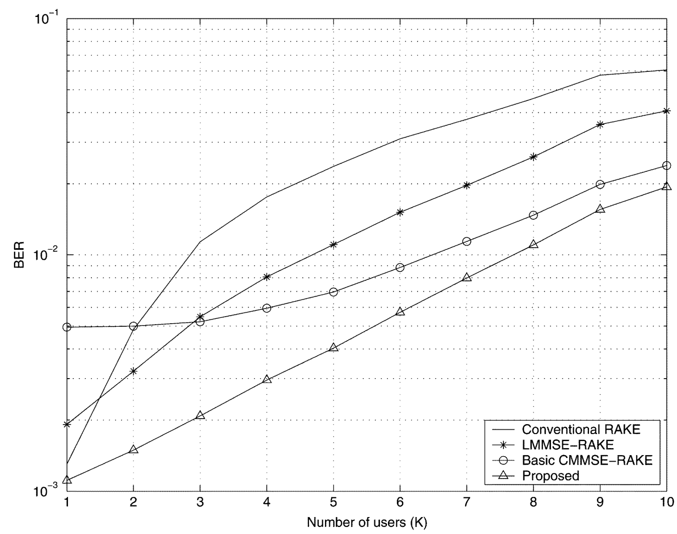

Fig. 5. BER versus number of users when $M = 3$ and $E_b/N_0 = 15$ dB (channel estimation employed).

Figs. 2–3 compare the BERs obtained from the analysis and simulation, under the assumption that the channel coefficients were known. Fig. 2 shows the BER versus the number of multipaths (M) when the number of users (K) was 3 and $E_b/N_0 = 15$ dB. The BERs decreased as M increased because of the diversity gain provided by the multipaths. The analytical and simulation results matched well. The BERs when $M = 3$ and the number of users K varied from 1 to 10 are compared in Fig. 3. Again, remarkably good agreements between the analysis and simulation were observed.

The performances of the proposed receiver were compared with those of the conventional RAKE, linear MMSE (LMMSE)-RAKE [6], and basic CMMSE-RAKE receivers when the channel parameters were unknown. To assist in the channel estimation, one pilot symbol was inserted at every ten data symbols $(Q_p = 10)$. The parameters N_p for the channel estimators and ρ for the weight update were set at 3 and 0.1, respectively, which exhibited the best performance.

Figs. 4–6 show the BERs for various values of M, K, and E_b/N_0. It was observed that the two CMMSE receivers outperformed the conventional and LMMSE-RAKE with a few exceptions: in Fig. 5 the basic CMMSE-RAKE receiver performed the worst when $K = 1$ and 2, due to channel estimation errors caused by IPI; in Fig. 6, the basic CMMSE performed slightly worse than the others at a low E_b/N_0, yet within this range of E_b/N_0 all the receivers exhibited poor performances. In general, the proposed receiver performed the best, however, the performance gain of the proposed receiver was achieved at the expense of additional computation.

Finally, it is interesting to compare the curves in Figs. 2 and 4 which show the BERs against M. The BER curve in Fig. 2 monotonically decreased as M increased, due to the diversity gain provided by the multipaths. On the other hand, the BER values in Fig. 4 tended to increase above a certain value of M, because the performance degradation of the channel estimators caused by IPI became severer as M increased.

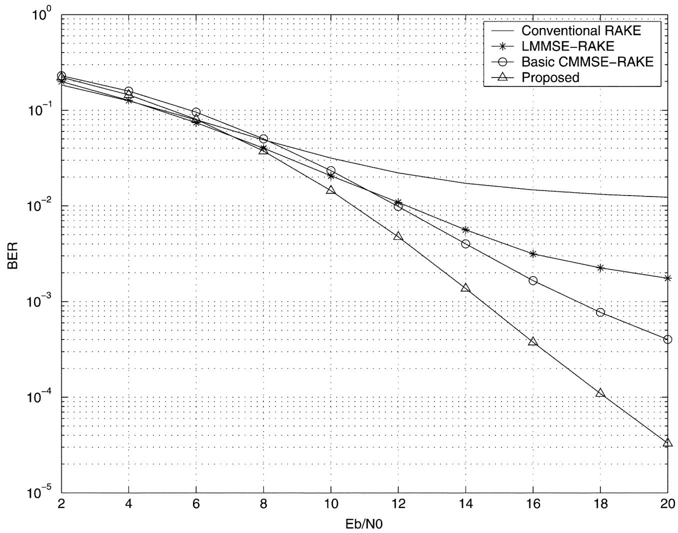

Fig. 6. BER versus E_b/N_0 when $K = 3$ and $M = 9$ (channel estimation employed).

VII. Conclusion

A new CMMSE-RAKE receiver for frequency-selective fading channels was proposed, and the closed-form BER expression derived. Simulation results showed that the proposed receiver provided a performance improvement over the conventional RAKE and currently available adaptive MMSE receivers. Further work will concentrate on the extension of CMMSE-RAKE receivers to multiple antenna systems.

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