Multiuser Channel Estimation for CDMA Systems Over Frequency-Selective Fading Channels

Jingxian Wu, Chengshan Xiao, *Senior Member, IEEE*, and Khaled Ben Letaief, *Fellow, IEEE*

***Abstract*—In this paper, a pilot-assisted minimum mean square error (MMSE) multiuser channel estimation algorithm is proposed for quasi-synchronous code-division multiple-access (CDMA) systems that undergo frequency-selective channel fading. The frequency-selective multiuser fading channel is represented as a symbol-wise time-varying chip-spaced tapped delay line filter with correlated filter taps. The multiuser channel tap coefficients at pilot symbol positions are estimated under the MMSE criterion with the help of the channel intertap correlation matrix, which is determined by the combined effects of the physical fading channel, transmit filter, and receive filter. In the development of the estimation algorithm, the channel intertap correlation matrix is deemed as an essential factor, and a novel iterative method is proposed for the joint estimation of the channel intertap correlation and filter tap timing based on the received pilot samples. Simulations show that the information of the channel intertap correlation is critical to the performance of the channel estimation, and the channel taps at different delays cannot be assumed uncorrelated for CDMA systems experiencing frequency-selective fading.**

***Index Terms*—CDMA, channel intertap correlation, MMSE, multiuser channel estimation, multiuser detection.**

I. Introduction

In a code-division multiple-access (CDMA) communication system, multiple users can access a given frequency bandwidth simultaneously with different preassigned spreading codes, which may lead to multiuser interference due to the nonperfect orthogonality of the spreading codes. Multiuser detectors [1], [2] can mitigate this problem by exploiting the information of all users present in the system, and substantial performance gain can be obtained over single-user detectors. Most of the works on multiuser detectors require knowledge of the multiuser channel states information, and this necessitates the research of multiuser channel estimation.

The topic of channel estimation for CDMA systems has received considerable attention in the literature [3]–[20]. Among them, the subspace-based blind channel estimation methods were proposed in the literature [3]–[5], [13], [19] to obtain channel parameters by exploiting the orthogonality of the signal and noise subspaces. The maximum-likelihood-based techniques were discussed in [7], [10], [11], and [18] for single-user channel and/or multiuser channel estimation with training symbols or pilots. The Kalman filter-based methods were considered in [9] and [17] to estimate and track time-varying channels, where a relatively long training sequence is required to ensure the proper identification of the Kalman filter parameters. The minimum mean square error (MMSE) methods were employed in [16] and [20] to estimate fading channels by using training sequences.

All the aforementioned algorithms appeared to work fine under certain assumptions on the fading channels to be estimated. Specifically, it was assumed in [3]–[7], [10], [11], and [13]–[16] that the fading channels are time invariant frequency selective during the entire estimation block. In [9] and [12], the fading channels were assumed to be symbol-wise time varying (i.e., channel varies from symbol to symbol and keeps constant within one symbol period) but frequency flat. In [20], both time-varying and frequency-selective channel was considered under the assumption that the delay spread and Doppler spread are known to the receiver. Moreover, most of the existing algorithms assumed that the transmit pulse shaping filer and receive matched filter are rectangular waveforms with a single chip duration. This is quite different from the bandwidth-limited root-raised cosine pulses adopted by the current and emerging CDMA wireless systems such as IS-95, cdma2000 [21], and universal mobile telecommunications system (UMTS) [22]. The rectangular shaping pulse assumption certainly leads to simple system models to explore new algorithms for channel estimation, but it implies unlimited bandwidth, which is not the case in practice. Additionally, it is commonly assumed that, for a wide-sense stationary uncorrelated scattering (WSSUS) channel [27], the uncorrelated multipaths are chip spaced [23], which is generally not common in practice due to the random nature of multipaths. This assumption leads the fading channels to be represented as tapped delay line filters whose taps are statistically uncorrelated [16], [17], [23]. However, it was recently shown in [24] when fractionally spaced WSSUS multipaths pass through the bandwidth-limited and/or time-limited receive filter and the chip-rate sampling (or any other rate sampling after the matched filter), the equivalent discrete-time time-varying channel taps are generally correlated. This intertap correlation can affect the system performance dramatically if it is not carefully taken.

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J. Wu was with the Department of Electrical and Computer Engineering, University of Missouri, Columbia, MO 65211 USA. He is now with the Department of Engineering Science, Sonoma State University, Rohnert Park, CA 94928 USA.

C. Xiao is with the Department of Electrical and Computer Engineering, University of Missouri, Columbia, MO 65211 USA (e-mail: xiaoc@missouri.edu; http://www.missouri.edu/~xiaoc/).

K. B. Letaief is with the Department of Electronic and Electrical Engineering, Hong Kong University of Science and Technology, Clear Water Bay, Kowloon, Hong Kong (e-mail: eekhaled@ee.ust.hk; http://www.ee.ust.hk/eekhaled/).

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In this paper, we focus on the multiuser channel estimation for quasi-synchronous CDMA (QS-CDMA)[1] systems under a time-varying and frequency-selective fading environment. The composite fading channel response, which combines the effects of the transmit filter, the physical frequency-selective fading channel, and the receive filter of each user in the QS-CDMA system, is represented as a tapped delay line filter with correlated tap coefficients, which is different from the system models presented in [16], [17], and [23], where uncorrelated tap coefficients were employed. Moreover, we will show that the channel intertap correlation is critical to the performance of the channel estimation. Utilizing the channel intertap correlation information, we propose an MMSE multiuser channel estimation algorithm, with the knowledge of the pilot symbols and spreading code of each user. In the development of the algorithm, the channel intertap correlation is treated as an essential factor, and efforts are put to preserve this information in the estimated channels. An iterative method is proposed for the joint estimation of the channel intertap correlation and the channel tap timing based on the received samples, and these parameters are used to form the MMSE algorithm. Simulation results show that the bit-error-rate (BER) performance of a CDMA system with our proposed multiuser channel estimation algorithm is close to that of a CDMA system with perfect multiuser channel knowledge.

The rest of this paper is organized as follows. Section II presents a discrete-time representation of the multiuser CDMA system with correlated channel taps. In Section III, multiuser channel estimation algorithms are summarized for a CDMA system with pilot-symbol-assisted modulation (PSAM). Section IV discusses the estimation of the channel statistics required by the MMSE algorithm, and a novel iterative method is presented for the joint estimation of the channel intertap correlation and channel tap timing information. Simulation results are given in Section V, and Section VI concludes the paper.

II. Discrete-Time System Model

A discrete-time model of the multiuser CDMA system is described in this section. We consider the uplink of a multiuser CDMA system consisting of M users. The transmitted signal $s_m(t)$ of the mth user is given by

$$s_m(t) = \sqrt{\frac{P_m}{N}} \sum_{i=-\infty}^{+\infty} \sum_{k=0}^{N-1} b_m(i) \cdot c_m(k) \cdot p(t - iT_\mathrm{s} - kT_\mathrm{c}) \tag{1}$$

where P_m is the average transmit power of the mth user, N is the processing gain, T_c is the chip period, $T_\mathrm{s} = NT_\mathrm{c}$ is the symbol period, $\mathbf{c}_m = [c_m(0), c_m(1), \cdots, c_m(N-1)]^\mathrm{T} \in \mathbb{C}^{N\times 1}$ is the mth user's spreading code[2] with $(\cdot)^\mathrm{T}$ denoting matrix transpose, $b_m(i)$ is the ith transmit data (or pilot) symbol, and $p(t)$ is the normalized root raised cosine (RRC) filter with $\int_{-\infty}^{\infty} p(t)p^*(t)\mathrm{d}t = 1$. The spreading code $\mathbf{c}_m$ satisfies $\mathbf{c}_m^\mathrm{H}\mathbf{c}_m = N$, with $(\cdot)^\mathrm{H}$ representing the Hermitian transpose. In a system with PSAM, the transmit symbols $b_m(i)$ of each user are divided into slots, with the pilot symbols being distributed within each slot.

Let $g_m(t,\tau)$ be the time-varying fading channel impulse response (CIR) for the mth user, then at the base station, the received signal $r(t)$ is the superposition of the fading distorted signals from all the M users plus the additive noise. $r(t)$ can be expressed as follows:

$$r(t) = \sum_{m=1}^{M} s_m(t - \Delta_m) \circledast g_m(t,\tau) + v(t) \tag{2}$$

where $\circledast$ denotes the convolution operation, Δ_m is the differential transmission delay experienced by the mth user, and $v(t)$ is the additive white Gaussian noise (AWGN) with variance N_0. For a quasi-synchronous system, the relative delay Δ_m is assumed to be uniformly distributed within $[-DT_\mathrm{c}, +DT_\mathrm{c}]$ with $D \ll N$ [17]. The received signal $r(t)$ is passed through the receive filter $p^*(-t)$, which is matched to the transmit filter $p(t)$, and the output $y(t) = r(t) \circledast p^*(-t)$ can be expressed by

$$y(t) = \sqrt{\frac{P_m}{N}} \sum_{m=1}^{M} \sum_{i=-\infty}^{+\infty} \sum_{k=0}^{N-1} b_m(i)c_m(k) \times \int_{-\infty}^{\infty} R_{pp}(t - iT_\mathrm{s} - kT_\mathrm{c} - \Delta_m - \alpha) g_m(t,\alpha)\mathrm{d}\alpha + z(t) \tag{3}$$

where

$$R_{pp}(t) = \int_{-\infty}^{\infty} p(t+\tau)p^*(\tau)\mathrm{d}\tau$$

$$z(t) = v(t) \circledast p^*(-t).$$

If we define the mth user's composite CIR $h_m(t,\tau)$ as

$$h_m(t,\tau) = \int_{-\infty}^{+\infty} R_{pp}(\tau - \Delta_m - \alpha) g_m(t,\alpha)\mathrm{d}\alpha \tag{4}$$

then the chip-rate sampled output of the matched filter can be written as

$$\begin{aligned} y_j(n) &= \sqrt{\frac{P_m}{N}} \sum_{m=1}^{M} \sum_{i=-\infty}^{+\infty} \sum_{k=0}^{N-1} b_m(i) \cdot c_m(k) \\ &\quad \cdot h_m\left[jN + n, (j-i)N + (n-k)\right] + z_j(n) \\ &= \sqrt{\frac{P_m}{N}} \sum_{m=1}^{M} \sum_{i=-\infty}^{+\infty} \sum_{l=-\infty}^{+\infty} b_m(i) \\ &\quad \cdot c_m\left[(j-i)N + (n-l)\right] \\ &\quad \cdot h_m(jN+n, l) + z_j(n) \\ &\quad \text{for } n = 0,1,\ldots,N-1;\ j = 1,2,3,\ldots \end{aligned} \tag{5}$$

[1]In a QS-CDMA system, the uncertainty of the relative transmission delay of each user is limited to a few chip periods, which can be achieved with the use of a GPS receiver at the base station and mobile stations [17], [25], [26].

[2]The normalized signature waveform of the mth user is given by $w_m(t) = (1/\sqrt{N}) \sum_{k=0}^{N=1} c_m(k)p(t - kT_\mathrm{c})$.

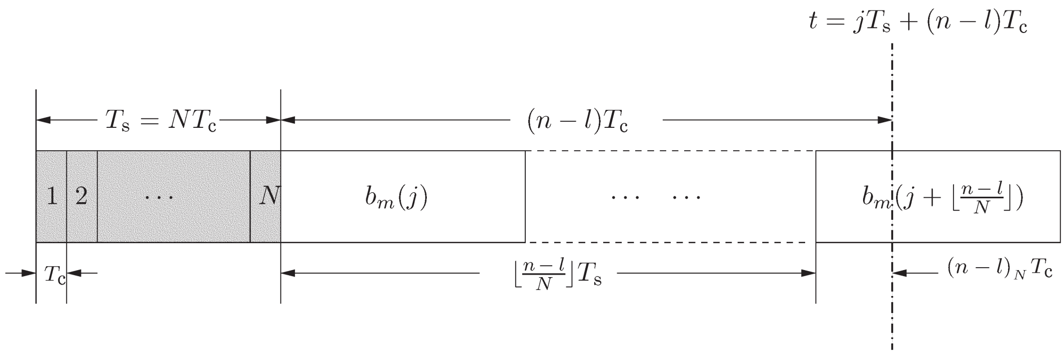

Fig. 1. The lth delayed version of the transmitted symbols.

where $y_j(n)$ is the nth chip-rate sample of the jth data symbol of $y(t)$ with $t = jT_\mathrm{s} + nT_\mathrm{c}$, and $z_j(n)$ is the chip-rate sample of $z(t)$ at the time instant $t = jT_\mathrm{s} + nT_\mathrm{c}$. Likewise, $h_m(jN + n, l) = h_m(jT_\mathrm{s} + nT_\mathrm{c}, lT_\mathrm{c})$ is the discrete-time version of the CIR $h_m(t, \tau)$, and we set $l = (j-i)N + (n-k)$ in the second equality. The noise component $z_j(n)$ is still AWGN with variance N_0 because the chip matched filter $p^*(-t)$ is a normalized RRC filter.

To simplify the representation of (5), we note that the chip index k of $c_m(k)$ satisfies $0 \le k < N$. Combining this inequality with $k = (j-i)N + (n-l)$, we can immediately get

$$j + \frac{n-l}{N} - 1 < i \le j + \frac{n-l}{N} \tag{6}$$

where i is the symbol index of the transmitted symbol $b_m(i)$, and it can only take integer values. In the range of i described in (6), there exists one and only one integer value, which must be $i = j + \lfloor (n-l)/N \rfloor$, where $\lfloor \cdot \rfloor$ denotes rounding to the nearest smaller integer. Substituting the value of i to $k = (j-i)N + (n-l)$, we can get

$$k = -\left\lfloor \frac{n-l}{N} \right\rfloor N + (n-l) = (n-l)_N \tag{7}$$

where $(x)_N$ can be viewed as the residue of x/N with $0 \le (x)_N \le N-1$. The above analysis leads to a simplified representation of (5)

$$\begin{aligned} y_j(n) &= \sqrt{\frac{P_m}{N}} \sum_{m=1}^{M} \sum_{l=-\infty}^{+\infty} b_m\left(j + \left\lfloor \frac{n-l}{N} \right\rfloor\right) \\ &\quad \cdot c_m\left[(n-l)_N\right] \cdot h_m(jN+n, l) + z_j(n) \\ &= \sqrt{\frac{P_m}{N}} \sum_{m=1}^{M} \sum_{l=-\infty}^{+\infty} d_m(j, l, n) \\ &\quad \cdot h_m(jN+n, l) + z_j(n) \end{aligned} \tag{8}$$

where $d_m(j, l, n) = b_m(j + \lfloor (n-l)/N \rfloor) \cdot c_m[(n-l)_N]$. The relationship of j, n, and l is illustrated in Fig. 1, where the lth delayed version of the transmitted symbols is given as an example, and the corresponding sampling time is $t = jT_\mathrm{s} + (n-l)T_\mathrm{c}$.

In the discrete-time system (8), the chip-wise time-varying and frequency-selective fading channel coefficients are represented by $h_m(jN + n, l)$, where l is the channel tap index, j is the symbol index, and n is the chip index within a symbol. Utilizing the same procedure described in [24], we can prove that the channel tap coefficients $h_m(j_1N + n_1, l_1)$ and $h_m(j_2N + n_2, l_2)$ are both temporally correlated and intertap correlated in a Rayleigh fading channel. If the physical channel $g_m(t, \tau)$ experiences WSSUS Rayleigh fading in which $g_m(t, \tau)$ is a zero-mean complex Gaussian random variable with autocorrelation given by $E[g_m(t_1, \tau) \cdot g_m^*(t_2, \tau')] = J_0[2\pi f_\mathrm{d} \cdot (t_1 - t_2)] \cdot G_m(\tau) \cdot \delta(\tau - \tau')$, where $J_0(\cdot)$ is the zero-order Bessel function of the first kind, f_d is the maximum Doppler frequency, $G_m(\mu)$ is normalized power delay profile of the channel with $\int_{-\infty}^{+\infty} G_m(\mu)\mathrm{d}\mu = 1$, then the cross correlation between $h_m(j_1N + n_1, l_1)$ and $h_m(j_2N + n_2, l_2)$ contains the temporal correlation and intertap correlation as follows:

$$\begin{aligned} &E\left[h_m(j_1N + n_1, l_1) \cdot h_m^*(j_2N + n_2, l_2)\right] \\ &\quad = J_0\left[2\pi f_\mathrm{d} \cdot (j_1N - j_2N + n_1 - n_2) \cdot T_\mathrm{c}\right] \cdot \rho_m(l_1, l_2) \end{aligned} \tag{9}$$

where $\rho_m(l_1, l_2)$ is the intertap correlation coefficient given by [24]

$$\begin{aligned} \rho_m(l_1, l_2) = \int_{-\infty}^{+\infty} & R_{pp}(l_1T_\mathrm{c} - \Delta_m - \mu) \\ & \times R_{pp}^*(l_2T_\mathrm{c} - \Delta_m - \mu) G_m(\mu)\mathrm{d}\mu. \end{aligned} \tag{10}$$

We will show in this paper that the intertap correlations can be exploited to enhance channel estimation accuracy. To achieve our objective, we would like to simplify the discrete-time system (8) by taking some practical issues into consideration.

In (8), we assumed that the discrete-time channel $h_m(jN + n, l)$ is a noncausal infinite impulse response (IIR) filter, where the tap index $l \in (-\infty, +\infty)$ for all j and m. The validity of this assumption lies in the fact that the time duration of the RRC filter $p(t)$ is infinite in theory to have a limited frequency bandwidth. In practice, the time-domain tails of RRC filter $p(t)$ falls off rapidly, and the physical CIR $g_m(t, \tau)$ has finite support in the τ domain. Thus, the amplitude of $h_m(k, l)$ will decrease quickly with the increase of $|l|$. When a channel tap coefficient's average power (or squared amplitude) is smaller than a certain threshold, this tap has very little impact on the output signals, and thus it can be discarded. Therefore, the time-varying noncausal IIR channel can be truncated to a finite impulse response channel. Without loss of generality, we use $\mathbf{l}_m = [-L_{m1}, \cdots, L_{m2}]^{\mathrm{T}} \in \mathbb{I}^{\lambda_m \times 1}$ to represent the tap index vector of the mth user,[3] where L_{m1} and L_{m2} are nonnegative integers, and λ_m is the length of the vector. Furthermore, it is pointed out that the chip-wise time-varying frequency-selective Rayleigh fading coefficient $h_m(jN+n, l)$ in (8) can be approximated by the symbol-wise time-varying frequency-selective Rayleigh fading coefficient $h_m(jN, l)$ if $f_{\mathrm{d}} T_{\mathrm{s}} \ll 1$, because in this case, the symbol duration T_{s} is much shorter than the channel coherence time $1/f_{\mathrm{d}}$. It is noted that the condition $f_{\mathrm{d}} T_{\mathrm{s}} \ll 1$ is generally valid for the current third-generation CDMA systems. For example, in a worst case scenario, a UMTS mobile handset with 2-GHz carrier frequency travels at a high speed of 120 km/h; the maximum Doppler frequency $f_{\mathrm{d}} = 222$ Hz. For a high spreading gain $N = 512$, the symbol duration $T_{\mathrm{s}} = 512 T_{\mathrm{c}} = 13.33$ ms, and therefore, $f_{\mathrm{d}} T_{\mathrm{s}} = 0.0296 \ll 1$. Based on these two aforementioned arguments, we can now simplify (8) as follows:

$$y_j(n) = \sqrt{\frac{P_m}{N}} \sum_{m=1}^{M} \sum_{l=-L_{m1}}^{L_{m2}} d_m(j, l, n) \cdot h_m(jN, l) + z_j(n). \tag{11}$$

We define the multiuser channel tap coefficient vector $\mathbf{h}(j)$ as follows:

$$\mathbf{h}(j) = \left[\mathbf{h}_1(j)^{\mathrm{T}}, \mathbf{h}_2(j)^{\mathrm{T}}, \cdots, \mathbf{h}_M(j)^{\mathrm{T}}\right]^{\mathrm{T}} \tag{12}$$

where $\mathbf{h}_m(j) = [h_m(jN, -L_{m1}), \cdots, h_m(jN, L_{m2})]^{\mathrm{T}}$ is the channel tap coefficient vector of the mth user. Then (11) can be written into a matrix format as follows:

$$\mathbf{y}(j) = \mathbf{D}(j) \cdot \mathbf{h}(j) + \mathbf{z}(j) \tag{13}$$

where $\mathbf{y}(j) = [y_j(0), y_j(1), \cdots, y_j(N-1)]^{\mathrm{T}}$ and $\mathbf{z}(j) = [z_j(0), z_j(1), \cdots, z_j(N-1)]^{\mathrm{T}}$ are the received sample vector and the additive noise vector of the jth symbol, respectively, and the matrix $\mathbf{D}(j) = \left[\mathbf{D}_1(j) \vdots \mathbf{D}_2(j) \vdots \cdots \vdots \mathbf{D}_M(j)\right] \in \mathbb{C}^{N \times \lambda}$ is made up of the data and spreading codes of all the users. The submatrix $\mathbf{D}_m(j)$ related to the mth user is defined by

$$\mathbf{D}_m(j) = \sqrt{\frac{P_m}{N}} \cdot [\mathbf{d}_m(j, -L_{m1}), \mathbf{d}_m(j, -L_{m1}+1), \cdots, \mathbf{d}_m(j, L_{m2})] \tag{14}$$

with the nth element of the vector $\mathbf{d}_m(j, l) \in \mathbb{C}^{N \times 1}$ being $d_m(j, l, n)$, for $n = 1, 2, \cdots, N$.

Equation (13) is a discrete-time representation of the multiuser CDMA system, and the time-varying and frequency-selective fading channel is represented as a T_{c}-spaced tapped delay line filter. With this representation, the necessary knowledge of the multiuser channel is the set of symbol-wise time-varying coefficients characterizing each path of the channel, and the problem of multiuser channel estimation is converted to the estimation of the time-varying channel coefficients $h_m(jN, l)$ and the T_{c}-spaced delay index vector $\mathbf{l}_m$. It should be noted from (4) that the relative delay Δ_m of each user is incorporated into the representation of the discrete-time CIR $h_m(jN, l)$. With the estimation of $\mathbf{h}(j)$ and $\mathbf{l}_m$, there is no need to explicitly recover the relative transmission delay Δ_m of all the users.

[3]It should be noted that the noncausality of the discrete-time channel model is due to the effects of transmit filter and receive filter, and the physical fading channel is always causal. Moreover, the noncausal effects of the discrete-time channel model can be removed by introducing proper delays at the receiver.

III. Multiuser Channel Estimation

In this section, we focus on the estimation of the channel coefficients at pilot positions, which will be interpolated to obtain the time-varying channel coefficients over one entire slot. In order to exploit the channel intertap correlation information, the proposed algorithm is based on the MMSE criterion, which is capable of utilizing and preserving the intertap correlation information of the fading channel.

We assume that the physical fading channels of different users are uncorrelated to each other, then the multiuser channel intertap correlation matrix $\mathbf{R}_h = E[\mathbf{h}(j)\mathbf{h}^{\mathrm{H}}(j)]$ can be written as

$$\mathbf{R}_h = \begin{bmatrix} \mathbf{R}_{h_1} & 0 & \cdots & 0 \\ 0 & \mathbf{R}_{h_2} & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & \mathbf{R}_{h_m} \end{bmatrix} \tag{15}$$

where $\mathbf{R}_{h_m} = E[\mathbf{h}_m(j)\mathbf{h}_m^{\mathrm{H}}(j)] \in \mathbb{C}^{\lambda_m \times \lambda_m}$ is the intertap correlation matrix of the mth user. With the definition of $\mathbf{h}_m(j)$ and (9), $\mathbf{R}_{h_m}$ can be written as

$$\mathbf{R}_{h_m} = \begin{bmatrix} \rho_m(-L_{m1}, -L_{m1}) & \cdots & \rho_m(-L_{m1}, L_{m2}) \\ \vdots & \ddots & \vdots \\ \rho_m(L_{m2}, -L_{m1}) & \cdots & \rho_m(L_{m2}, L_{m2}) \end{bmatrix}. \tag{16}$$

It can be seen from (10) and (16) that the multiuser channel intertap correlation matrix $\mathbf{R}_h$ is a function of the power delay profile $G_m(\mu)$, the relative transmission delay Δ_m, and the tap delay index vector $\mathbf{l}_m$. These parameters are usually unavailable at the receiver. Therefore, $\mathbf{R}_h$ is generally not known to the receiver. In this section, we focus on the formulation of the MMSE-based channel estimation algorithm. The estimation of $\mathbf{R}_h$ will be discussed in the next section.

For a QS-CDMA system, it is assumed that all the users are slot synchronized, and the received signals at pilot positions are the superposition of the faded pilot symbols of all the users. For convenience of representation, the symbol "1" is used as pilot symbols. According to (11), the received samples contributed exclusively by pilot symbols can be written as

$$\mathbf{y}(j_p) = \mathbf{C}\mathbf{h}(j_p) + \mathbf{z}(j_p) \tag{17}$$

where j_p is the index of the pth pilot symbol for a slot with P pilot symbols, $\mathbf{y}(j_p) = [y_{j_p}(0), y_{j_p}(1), \cdots, y_{j_p}(N-1)]^{\mathrm{T}}$, $\mathbf{z}(j_p) = [z_{j_p}(0), z_{j_p}(1), \cdots, z_{j_p}(N-1)]^{\mathrm{T}}$, and the multiuser code matrix $\mathbf{C}$ is defined by

$$\mathbf{C} = \left[\mathbf{C}_1 \;\vdots\; \mathbf{C}_2 \;\vdots\; \dots \;\vdots\; \mathbf{C}_M \right] \tag{18}$$

with

$$\mathbf{C}_m = [\mathbf{c}_m(-L_{m1}), \cdots, \mathbf{c}_m(L_{m2})] \tag{19}$$

where $\mathbf{c}_m(i) = [c_m(N-i), \cdots, c_m(N-1), c_m(0), \cdots, c_m(N-i-1)]^{\mathrm{T}}$ is obtained from circularly shifting i symbols of the original code vector $\mathbf{c}_m$. It is important to note that the multiuser code matrix $\mathbf{C}$ is determined by both the spreading codes and the tap delay index vector $\mathbf{l}_m$. For a multiuser detector, the spreading code of each user is known to the base station, while $\mathbf{l}_m$ needs to be estimated. We will show in the next section that $\mathbf{l}_m$ can be jointly estimated with $\mathbf{R}_h$ based on a novel iterative method.

Based on (17), we can immediately obtain the multiuser channel estimation at pilot symbol locations by utilizing least-squares (LS) method [33] and MMSE method [34], and the results are stated as follows.

The LS-based estimation of the multiuser channel tap coefficients $\mathbf{h}(j_p)$ at pilot symbol positions is given by

$$\hat{\mathbf{h}}_{\mathrm{LS}}(j_p) = \mathbf{C}^{\dagger}\mathbf{y}(j_p) \quad \text{for } p = 1, 2, \cdots, P \tag{20}$$

where $\mathbf{C}^{\dagger}$ is the pseudoinverse of the multiuser code matrix $\mathbf{C}$ and j_p is the position index of the pth pilot symbol in a slot with P pilot symbols.

The MMSE-based estimation of the multiuser channel tap coefficients $\mathbf{h}(j_p)$ at pilot symbol positions is given by

$$\hat{\mathbf{h}}_{\mathrm{MMSE}}(j_p) = \mathbf{R}_h\mathbf{C}^{\mathrm{H}} \cdot [\mathbf{C}\mathbf{R}_h\mathbf{C}^{\mathrm{H}} + N_0\mathbf{I}_N]^{-1} \cdot \mathbf{y}(j_p) \quad \text{for } p = 1, 2, \cdots, P \tag{21}$$

where N_0 is the additive noise variance and $\mathbf{I}_N$ is an $N \times N$ identity matrix.

To fulfill the MMSE-based channel estimation, we need to know the multiuser channel intertap correlation matrix $\mathbf{R}_h$, the delay index vector $\mathbf{l}_m$, and the additive noise variance N_0. The estimation of these parameters are discussed in the next section.

For comparison purpose, a pilot-assisted subspace-based estimation of $\mathbf{h}(j_p)$ for quasi-static multiuser fading channels is derived in the Appendix.

IV. MMSE Parameter Estimation

In this section, we consider the estimation of the multiuser channel intertap correlation matrix $\mathbf{R}_h$, the tap delay index vector $\mathbf{l}_m$, and the additive noise variance N_0. The noise variance N_0 is estimated by exploiting the eigen structure of the received signals. Likewise, the matrix $\mathbf{R}_h$ and the vector $\mathbf{l}_m$ are jointly estimated by a novel iterative method.

A. Estimation of the Additive Noise Variance N_0

The variance N_0 of the additive noise component $\mathbf{z}(j)$ can be extracted from the received data symbols $\mathbf{y}(j)$ with the method presented in [13].

According to the discrete-time model of the multiuser CDMA system as described in (13), the correlation matrix of the received symbol vector $\mathbf{y}(j)$ can be written as

$$\mathbf{R}_y = E\left[\mathbf{D}(j)\mathbf{R}_h\mathbf{D}^{\mathrm{H}}(j)\right] + N_0\mathbf{I}_N. \tag{22}$$

As shown in [13], the noise variance N_0 is equal to the smallest $N-M$ eigenvalues of the correlation matrix $\mathbf{R}_y$. Therefore, an estimation of N_0 can be obtained as the average of the smallest $N-M$ eigenvalues of $\mathbf{R}_y$. The correlation matrix $\mathbf{R}_y$ can be estimated from the received data samples as

$$\widehat{\mathbf{R}}_y = \frac{1}{J}\sum_{j=1}^{J}\mathbf{y}(j)\mathbf{y}^{\mathrm{H}}(j) \tag{23}$$

where J is the number of symbols in one slot. It is apparent that the larger the value of J, the more accurate the estimation of $\mathbf{R}_y$ is. To increase the estimation accuracy, several consecutive slots can be used to form a hyperslot in the estimation process. After $\widehat{\mathbf{R}}_y$ is obtained, we can perform the eigenvalue decomposition of it, and the average of the smallest $N-M$ eigenvalues is the estimated value of N_0.

B. Joint Estimation of $\mathbf{R}_h$ and $\mathbf{l}_m$

In this subsection, an iterative method is proposed for the joint estimation of the multiuser channel intertap correlation matrix $\mathbf{R}_h$ and tap delay index vector $\mathbf{l}_m$, for $m = 1, 2, \cdots, M$. The elements of the intertap correlation matrix $\mathbf{R}_h$ are mainly determined by the relative transmission delay Δ_m and the fading channel power delay profile $G_m(\mu)$ of each user. In the tapped delay line representation of the fading channel, the effects of Δ_m and the delay spread of $G_m(\mu)$ are incorporated into the T_c-spaced tap delay index vector $\mathbf{l}_m = [-L_{m1}, \cdots, L_{m2}]^{\mathrm{T}}$. Both $\mathbf{R}_h$ and $\mathbf{l}_m$ are interacted to each other, and this interaction can be utilized for the joint estimation of $\mathbf{R}_h$ and $\mathbf{l}_m$. It is pointed out here that our algorithm does not need to estimate either $G_m(\mu)$ or Δ_m.

The interactions between $\mathbf{R}_h$ and $\mathbf{l}_m$ can be explored via the help of the statistical properties of the received pilot samples. According to (17), the correlation matrix $\mathbf{R}_{y_p} = E[\mathbf{y}(j_p)\mathbf{y}(j_p)^{\mathrm{H}}]$ of the received pilot symbols can be written as

$$\mathbf{R}_{y_p} = \mathbf{C}\mathbf{R}_h\mathbf{C}^{\mathrm{H}} + N_0\mathbf{I}_N. \tag{24}$$

In the equation above, the noise variance N_0 can be obtained from the method described in Section IV-A, $\mathbf{R}_{y_p}$

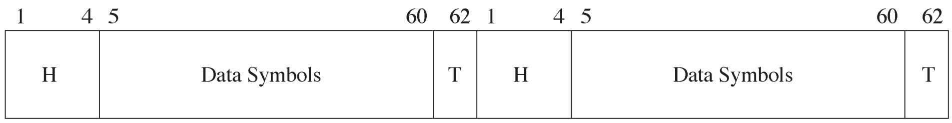

Fig. 2. The slot structure to be utilized for simulations.

can be estimated from the received pilot symbols as $\widehat{\mathbf{R}}_{y_p} = (1/P)\sum_{p=1}^{P}\mathbf{y}(j_p)\mathbf{y}^{\mathrm{H}}(j_p)$, while $\mathbf{R}_h$and the multiuser code matrix $\mathbf{C}$ are two unknown matrices to be determined. From the definition of $\mathbf{C}$ in (19), we can see that $\mathbf{C}$ is determined by both the spreading codes and the tap delay index vector $\mathbf{l}_m = [-L_{m1}, \cdots, L_{m2}]^{\mathrm{T}}$. As discussed in Section II, the vector $\mathbf{l}_m$ is obtained by discarding the channel taps with power smaller than a certain threshold, and the channel tap power $E[h_m(jN,l) \cdot h_m^*(jN,l)] = \rho_m(l,l)$ can be found from the diagonal of $\mathbf{R}_h$. If we know the power of all the possible channel taps, then we can obtain $\mathbf{l}_m$ by discarding the negligible taps. Furthermore, when $\mathbf{l}_m$ is known, we can form the code matrix $\mathbf{C}$, with which $\mathbf{R}_h$ can be computed from (24).

Based on the reciprocal relationship between $\mathbf{R}_h$ and $\mathbf{l}_m$, an iterative method is proposed for the joint estimation of these two parameters.

Algorithm: Joint estimation of the multiuser channel intertap correlation matrix $\mathbf{R}_h$ and the tap delay index vector $\mathbf{l}_m$.

Step 1) Set the initial value of the tap delay index vector as $\mathbf{l}_m = [-D-1, L_{m0}+D]$,[4] where D is the maximum relative transmission delay factor and $L_{m0} \approx \tau_{\max}^{(m)}/T_{\mathrm{c}}$ with $\tau_{\max}^{(m)}$ being the maximum possible delay spread of the mth user's physical channel.

Step 2) Based on the current value of $\mathbf{l}_m$, construct the multiuser code matrix $\mathbf{C}$ according to (19). With the estimated value of $\mathbf{R}_{y_p}$, N_0, and the current value of $\mathbf{C}$, compute $\mathbf{R}_h$ as follows:

$$\mathbf{R}_h = \mathbf{C}^{\dagger}\left(\mathbf{R}_{y_p} - N_0\mathbf{I}_N\right)\left(\mathbf{C}^{\mathrm{H}}\right)^{\dagger}. \tag{25}$$

Step 3) With the diagonal elements of $\mathbf{R}_h$ obtained from Step 2, find the maximum power taps for each user and represent the maximum tap power of the mth user as $\mathcal{P}_m$. For all the taps of the mth user, discard the taps that are smaller than $\eta \cdot \mathcal{P}_m$, with $0 < \eta < 1$ being a predefined threshold value.

Step 4) After discarding the negligible taps for each user, a new tap delay index vector $\mathbf{l}_m$ for each user can be formed, then go back to Step 2. If there are no more taps to discard, or the maximum number of iterations is reached, then the current values of $\mathbf{R}_h$ and $\mathbf{l}_m$ are the desired values.

With the proposed iterative method, the value of $\mathbf{R}_h$ and $\mathbf{l}_m$ can be jointly estimated from the received pilot samples for each slot. When we set $\eta = 1\%$, simulations show that the iterative method usually converges within two iterations, and it leads to accurate estimations of $\mathbf{R}_h$ and $\mathbf{l}_m$. The estimated values of $\mathbf{R}_h$, $\mathbf{l}_m$, and N_0 are utilized to form the MMSE solution of the multiuser channel tap coefficients $\mathbf{h}_m(j_p)$ at pilot positions as stated in (21).

The time-varying channel coefficients of one entire slot can be obtained by interpolating the MMSE-estimated CIR at pilot positions. The topic of channel interpolation for systems with PSAM has been researched extensively in the literature [28]–[30]. Among these methods, Weiner filter interpolation [28] is an optimum solution in the sense of mean square error (MSE), provided that the temporal correlation of each time-varying channel tap is accurately known to the receiver, which is unlikely in practice. Therefore, we adopt a suboptimum constant matrix interpolation method [30], which was proposed for time division multiple access-based systems. The extension of this method to CDMA system is straightforward, and details are omitted here for brevity.

V. Simulation Results

Simulations are carried out in this section to evaluate the performance of the proposed multiuser channel estimation algorithm for QS-CDMA systems that undergo time-varying and frequency-selective channel fading.

A. System Configurations

The slot structure used in our simulations is shown in Fig. 2. Each slot has four head pilot symbols and two tail pilot symbols with 56 data symbols in the middle. The time duration of each slot is 2 ms, and every two slots are combined as a hyperslot in the estimation process. Gold sequences with processing gain of $N = 127$ are used as the spreading codes. The chip rate of the system is chosen to be 3.84 MHz. The transmitted data are quaternary phase-shift keying modulated. RRC filter with roll-off factor of 0.22 is used as both the transmit filter and the chip matched filter. Vehicular A propagation profile [22], [32] shown in Fig. 3 is chosen to be the power delay profile $G_m(\tau)$ for simulations with the normalized Doppler frequency set to $f_{\mathrm{d}}T_{\mathrm{slot}} = 0.1$. The time-varying channel fading is generated with the model described in [31]. The relative transmission delay Δ_m of each user is uniformly distributed in $[-DT_{\mathrm{c}}, DT_{\mathrm{c}}]$. Unless otherwise stated, we set $D = 3$ in the simulations. In the iterative estimation of $\mathbf{R}_h$ and $\mathbf{l}_m$, the maximum number of iterations is set to 2, and the discarding threshold η is set to 1%.

A successive interference canceller [2] is employed for coherent multiuser detection. The users are sorted according to their received power, which can be obtained from the diagonal elements of $\mathbf{R}_h$. The users are then detected and cancelled from the strongest to the weakest. In the detection of each of the

[4]The reason we choose $-D-1$ as the smallest possible tap delay index lies in the fact that the absolute amplitude of the raised cosine filter is very small for samples one chip period away from the peak value and the physical fading channel is always causal.

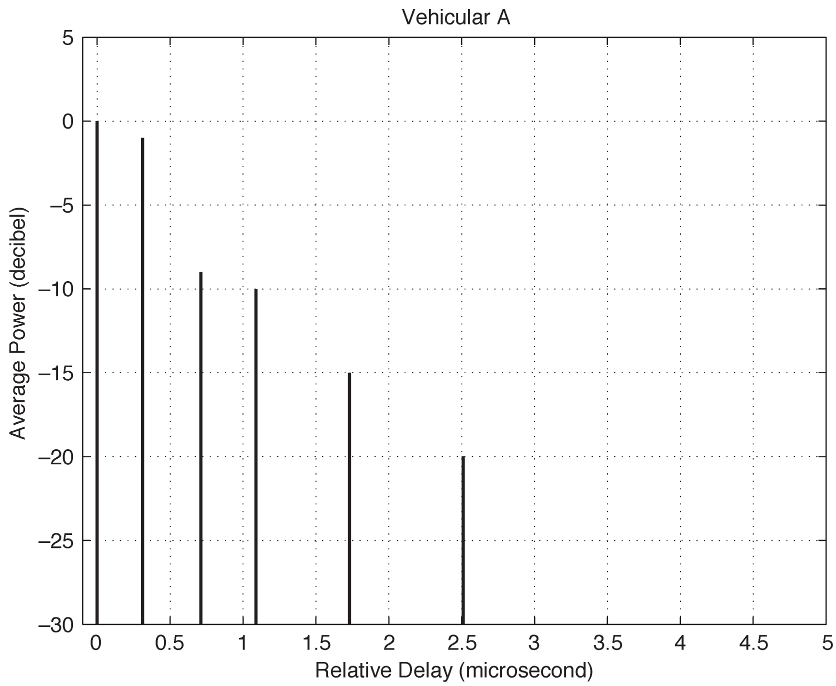

Fig. 3. Vehicular A propagation profile. The differential delays between multiple paths are noninteger of the chip period T_c.

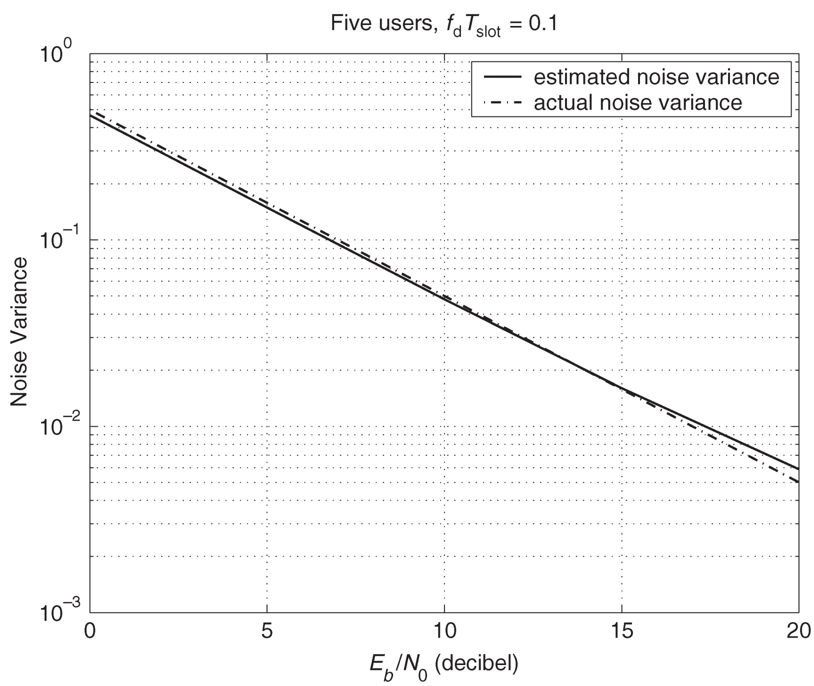

Fig. 4. Comparison of the estimated noise variance and its corresponding actual noise variance at different levels of E_b/N_0.

users, coherent RAKE combining is used, with the number of RAKE fingers equal to the number of channel taps of the corresponding user being detected, which is λ_m for the mth user.

B. Performance Evaluation

The proposed multiuser channel estimation algorithm requires knowledge of the additive noise variance N_0, the multiuser channel intertap correlation matrix $\mathbf{R}_h$, and the tap delay index vector $\mathbf{l}_m$. These parameters can be obtained from the received data samples with the methods described in Section IV. The validity of these methods is evaluated in the sequel.

Consider a QS-CDMA system with five users. The additive noise variance N_0 is estimated by utilizing the method described in Section IV-A. For comparison purpose, Fig. 4 shows the estimated noise variances along with the actual noise variances for various values of E_b/N_0. The estimation results are based on averaging 1000 estimated values. As we can see in Fig. 4, the estimation of the noise variances is very accurate.

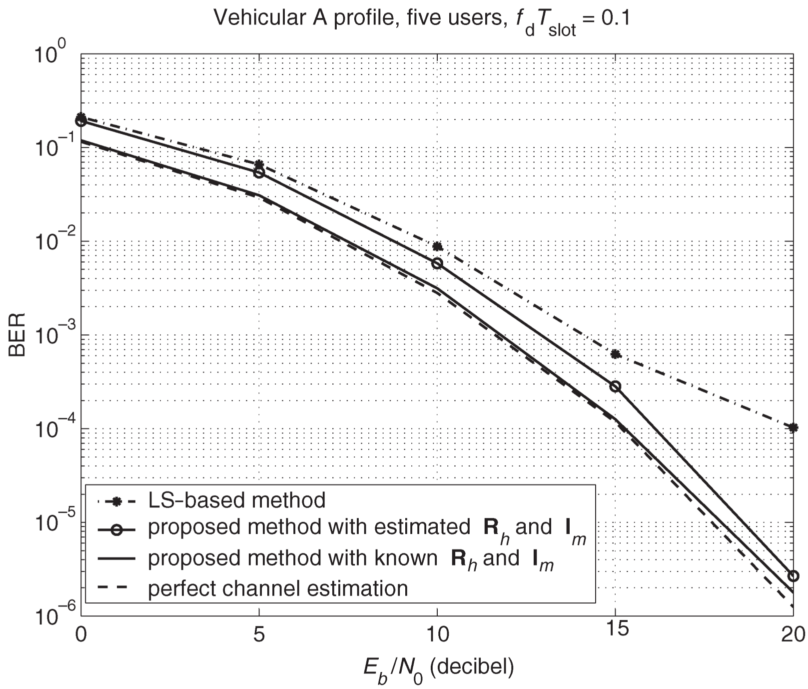

Fig. 5. BER performance comparison of the system that employs our multiuser channel estimation algorithm and the ideally perfect channel estimation.

The effectiveness of the estimation for $\mathbf{R}_h$ and $\mathbf{l}_m$ can be demonstrated by the BER performance of a five-user CDMA system that employs the proposed channel estimation algorithm. In Fig. 5, four cases are depicted for BER comparison. The first case is the BER performance of the system that has perfect knowledge of the multiuser fading channels. This BER shall serve as a benchmark for our channel estimation performance. The second case is the BER of the system with knowledge of the multiuser channel intertap correlation matrix $\mathbf{R}_h$ and the tap index vector $\mathbf{l}_m$, which are utilized to estimate (and interpolate) the multiuser fading channels. The third case is the BER of the system that has estimated both $\mathbf{R}_h$ and $\mathbf{l}_m$, which are then utilized to estimate (and interpolate) the multiuser fading channels. The fourth case is the BER of the system that employs the LS-based method to estimate the channel coefficients at pilot locations, where the LS-based algorithm only utilizes the first-order statistics of the fading channel, and the results are labeled as "LS-based method." As can be seen in Fig. 5, when the receiver knows the channel intertap correlation matrix $\mathbf{R}_h$ and the tap index vector $\mathbf{l}_m$, our multiuser channel estimation algorithm has nearly the same BER performance as the ideally perfect channel estimation case. However, as expected, when the receiver has to estimate both the channel correlation matrix and the tap index vector, our multiuser channel estimation algorithm will have a little degradation on the BER performance from the perfect channel estimation case, but the degradation is within an acceptable range. For example, it is about 0.8 dB when the BER is at the level of 10^{-4}.

Comparing the four curves discussed above, we can see that the multiuser channel intertap correlation matrix $\mathbf{R}_h$ plays a very important role in the performance of the estimation algorithm. If we do not take advantage of this correlation information for multiuser channel estimation, we will get a BER performance penalty, which can be significant compared to our proposed MMSE-based algorithm.

In Fig. 6, we compare the BER performance of the system with our proposed MMSE-based channel estimation algorithm to that of the pilot-assisted subspace-based algorithm shown in the Appendix. The subspace-based estimation algorithm is derived for quasi-static fading channels, where the channels can be viewed as deterministic during the entire estimation process; therefore, the correlation information does not play an important role in the estimation. For quasi-static fading channels $(f_\mathrm{d}T_\mathrm{slot} = 0)$, we can see in the figure that the subspace-based algorithm can achieve nearly the same performance as the proposed MMSE algorithm. However, when we increase $f_\mathrm{d}T_\mathrm{slot}$ to 0.1, performance degradation can be clearly observed for the subspace-based method for $E_b/N_0 \geq 10$ dB, while the performance of the proposed algorithm is not apparently affected.

To further show our proposed algorithm's ability of estimating the tap index vector $\mathbf{l}_m$, we consider two cases that the maximum transmission delay factor D is set to 3 and 6. This means that the relative transmission delay Δ_m of each user is uniformly distributed in $[-3T_\mathrm{c}, 3T_\mathrm{c}]$ and $[-6T_\mathrm{c}, 6T_\mathrm{c}]$, respectively. In the simulation, we consider two scenarios for both $D = 3$ and $D = 6$ cases. First, we assume that the tap index vector $\mathbf{l}_m$ is known to the receiver, and the obtained BER curves are labeled "known $\mathbf{l}_m$" as shown in Fig. 7. Second, when the vector $\mathbf{l}_m$ is estimated with our proposed iterative method, the obtained BER curves are labeled "estimated $\mathbf{l}_m$." It is noted that all the four curves in Fig. 7 are based on the estimated $\mathbf{R}_h$ by using our iterative algorithm. We have three observations in Fig. 7. First, when the receiver has knowledge of $\mathbf{l}_m$, changing the maximum transmission delay range has no apparent effect on the system performance. Second, for a system with $\mathbf{l}_m$ being estimated, a slight BER degradation will

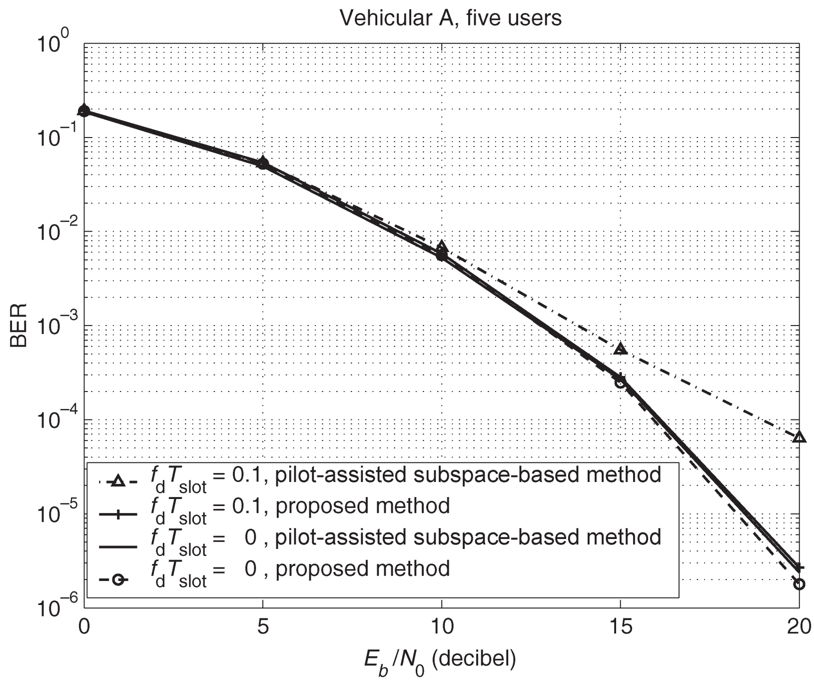

Fig. 6. BER performance comparison of the system that employs our multiuser channel estimation algorithm and the pilot-assisted subspace-based channel estimation.

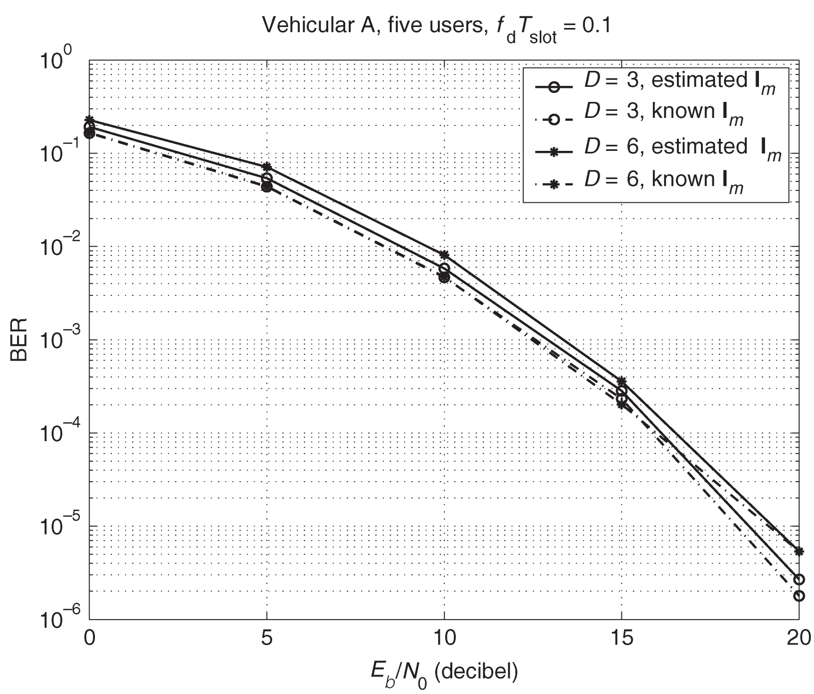

Fig. 7. BER comparison for the effect of the estimated tap delay index vector $\mathbf{l}_m$ and ideally known tap delay index vector on system performance. All the curves are based on estimated $\mathbf{R}_h$ using the proposed algorithm.

occur if the maximum transmission delay range is increased. Third, the BER performance of a system with estimated $\mathbf{l}_m$ are close to that of a system with perfect information of $\mathbf{l}_m$. These results indicate that the proposed algorithm provides accurate estimation of $\mathbf{l}_m$ in a wide range of E_b/N_0.

We are now in a position to take a look at the normalized MSE. Let $\hat{\mathbf{h}}(j)$ be the estimation of the multiuser channel tap coefficient vector $\mathbf{h}(j)$. We define the normalized MSE as $E\{[\mathbf{h}(j)-\hat{\mathbf{h}}(j)]^{\mathrm{H}}[\mathbf{h}(j)-\hat{\mathbf{h}}(j)]\}/E[\mathbf{h}(j)^{\mathrm{H}}\mathbf{h}(j)]$ and present the channel estimation errors of the LS-based method, the pilot-assisted subspace-based method, and the proposed MMSE algorithm in Fig. 8. It is observed that given a value of E_b/N_0, the normalized MSE of the proposed MMSE method with known $\mathbf{R}_h$ and $\mathbf{l}_m$ is always smaller than that of the

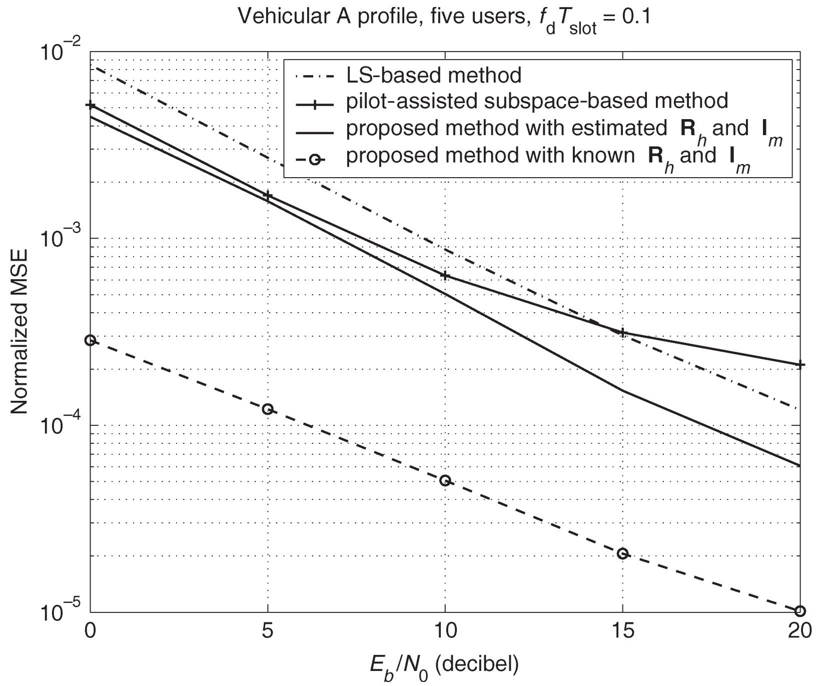

Fig. 8. The normalized MSE performance of the system that uses the proposed algorithm, the LS-based method, and the pilot-assisted subspace-based method.

proposed MMSE method with estimated $\mathbf{R}_h$ and $\mathbf{l}_m$, and the MSE of the LS-based and subspace-based method is always larger than that of the proposed MMSE method with estimated $\mathbf{R}_h$ and $\mathbf{l}_m$. The MSEs of these four cases are well reflected in the BER performances listed in Figs. 5 and 6.

So far, all the simulation results are focused on the vehicular A propagation profile, which has a discrete-time power delay profile (or discrete-time delayed multiple paths). We would like to point out that our algorithm can be directly applied to fading channels that have continuous-time power delay profile. For example, we replace vehicular A power delay profile by an exponentially decaying profile, whose power delay profile is defined as $G_m^c(\tau) = A \cdot \exp[-(\tau/1\ \mu\text{s})]$ with $0 \leq \tau \leq 1.5\ \mu$s. If we keep the rest of the simulation configurations of Fig. 5 unchanged, then we get the corresponding BER comparison for $G_m^c(\tau)$ as shown in Fig. 9, which indicates that our multiuser channel estimation algorithm is still very effective under continuous-time power delay profile fading environment. However, it should be pointed out that most existing channel estimation algorithms will fail under this fading condition.

VI. Conclusion

In this paper, a pilot-assisted MMSE multiuser channel estimation algorithm was proposed for QS-CDMA systems undergoing time-varying and frequency-selective channel fading. The algorithm was developed based on the only assumption that the base station receiver knows the spreading codes and pilot symbols of all the mobile users, which is very reasonable in practice. The combined effects of the frequency-selective physical fading channel, the transmit filter, and receive filter were represented as a symbol-wise time-varying chip-spaced tapped delay line filter with correlated filter taps. A novel iterative method was then proposed for the joint estimation of the multiuser channel intertap correlations and tap delays, which were further utilized to form the MMSE-based multiuser channel tap coefficient estimation. The multiuser channel estimation algorithm can be used to estimate fading channels that have either discrete-time or continuous-time power delay profiles. Simulation results showed that the information of the channel intertap correlations is critical to the performance of the multiuser channel estimation, and the discrete-time composite channel taps at different delays may not be assumed uncorrelated for CDMA systems that experience physical WSSUS fading. Furthermore, when the channel intertap correlation is known to the receiver, the BER performance of the proposed MMSE algorithm is nearly the same as that of the perfect channel estimation case; when the channel intertap correlation is estimated from the received signals, the proposed algorithm's BER performance is close to that of the perfect channel estimation case.

Appendix

Subspace-Based Channel Estimation With Pilot Symbols

In this Appendix, a subspace-based channel estimation algorithm with pilot symbols is derived for systems with quasi-static fading channels. The assumption of quasi-static fading is required for the purpose of proper identification of the signal subspace and noise subspace [3]–[5], and the channel is assumed to be constant during the estimation process, which is one slot in this paper. From (17), we will have

$$\begin{aligned}\mathbf{R}_{y_p} &= E\left[\mathbf{y}(j_p)\mathbf{y}^{\mathrm{H}}(j_p)\right],\\ &= \mathbf{Ch}\cdot\mathbf{h}^{\mathrm{H}}\mathbf{C}^{\mathrm{H}} + N_0\mathbf{I}_N \end{aligned} \tag{26}$$

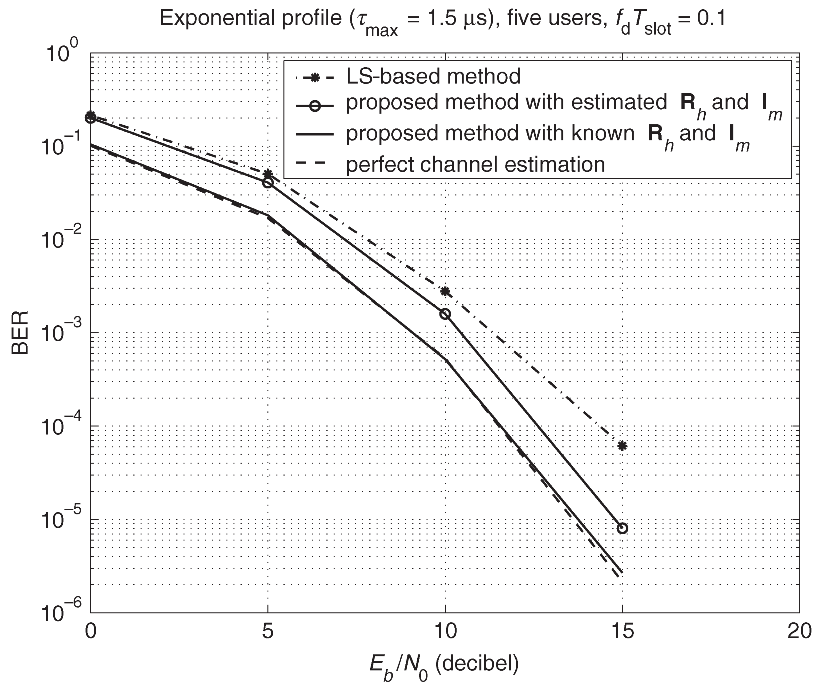

Fig. 9. BER performance comparison of the system that employs our multiuser channel estimation algorithm and the ideally perfect channel estimation, where the power delay profile is a continuous-time exponentially decaying function.

where $\mathbf{h} = \mathbf{h}(j)$, $j = 1, 2, \cdots, J$ for quasi-static fading channels. Since $\mathbf{C} \cdot \mathbf{h}$ is a column vector, the rank of the matrix $\tilde{\mathbf{R}}_{y_p} = \mathbf{R}_{y_p} - N_0 \mathbf{I}_N$ is 1. Therefore, an eigenvalue decomposition of $\tilde{\mathbf{R}}_{y_p}$ defines a signal subspace of dimension 1 and a noise subspace of dimension $N - 1$

$$\tilde{\mathbf{R}}_{y_p} = [\, \mathbf{u}_\mathrm{s} \quad \mathcal{U}_n \,] \begin{bmatrix} \lambda_\mathrm{s} & \mathbf{0} \\ \mathbf{0} & \Lambda_n \end{bmatrix} \begin{bmatrix} \mathbf{u}_\mathrm{s}^\mathrm{H} \\ \mathcal{U}_n^\mathrm{H} \end{bmatrix} \tag{27}$$

where the scalar λ_s contains the largest eigenvalue of $\tilde{\mathbf{R}}_{y_p}$, $\mathbf{u}_\mathrm{s} \in \mathbb{C}^{N\times 1}$ is the corresponding eigenvector defining the signal subspace, and $\mathcal{U}_n = [\mathbf{u}_2, \mathbf{u}_3, \cdots, \mathbf{u}_N] \in \mathbb{C}^{N\times(N-1)}$ are the $N - 1$ orthonormal eigenvectors spanning the $(N - 1)$-dimensional noise subspace.

It follows from the analysis above that the vector $\mathbf{C} \cdot \mathbf{h}$ should be orthogonal to the noise subspace spanned by $\mathcal{U}_n$, i.e., $(\mathbf{C} \cdot \mathbf{h})^\mathrm{H} \cdot \mathbf{u}_k = 0$ for $k = 2, \cdots, N$. Therefore, an estimation of the CIR vector $\mathbf{h}$ can be obtained by

$$\tilde{\mathbf{h}} = \underset{\mathbf{h}}{\operatorname{argmin}} \left[\mathbf{h}^\mathrm{H} \mathbf{C}^\mathrm{H} \left(\sum_{k=2}^{N} \mathbf{u}_k \mathbf{u}_k^\mathrm{H} \right) \mathbf{C}\mathbf{h} \right] \tag{28}$$

and the solution under the constraint $\mathbf{h}\mathbf{h}^\mathrm{H} = 1$ is the eigenvector corresponding to the smallest eigenvalue of the matrix $\mathbf{C}^\mathrm{H}(\sum_{k=2}^{N} \mathbf{u}_k \mathbf{u}_k^\mathrm{H})\mathbf{C}$ up to a multiplicative factor [13], i.e., the estimated CIR $\mathbf{h}$ can be expressed as the product of $\tilde{\mathbf{h}}$ obtained from (28) and a complex-valued scalar ζ

$$\hat{\mathbf{h}}_{SSp} = \tilde{\mathbf{h}} \cdot \zeta. \tag{29}$$

For blind channel estimation, the value of ζ is not attainable, and it is argued in [13] that this problem can be alleviated via differential encoding and decoding, which may result in performance degradation compared to coherent systems. To solve the ambiguity of ζ, we apply the subspace-based method in systems with pilot-assisted modulation, and the value of ζ can be estimated with the help of the transmitted pilot symbols.

From (17) and (29), we have

$$\frac{1}{P} \sum_{p=1}^{P} \mathbf{y}(j_p) = \mathbf{C}\tilde{\mathbf{h}} \cdot \zeta + \bar{\mathbf{z}} \tag{30}$$

where P is the number of pilot symbols within one slot and $\bar{\mathbf{z}} = (1/P) \sum_{p=1}^{P} \mathbf{z}(j_p)$. This is a linear system with N equations and 1 unknown variable, and the value of ζ can be estimated as

$$\zeta = \frac{1}{P} \cdot (\mathbf{C}\tilde{\mathbf{h}})^\dagger \cdot \left[\sum_{p=1}^{P} \mathbf{y}(j_p) \right]. \tag{31}$$

Combining (28), (29), and (31), we will have the subspace estimation of the CIR $\mathbf{h}$.

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Jingxian Wu received the B.S.(EE) degree from the Beijing University of Aeronautics and Astronautics, Beijing, China, in 1998, the M.S.(EE) degree from Tsinghua University, Beijing, China, in 2001, and the Ph.D. degree from the University of Missouri, Columbia, in 2005.

He is an Assistant Professor at the Department of Engineering Science, Sonoma State University, Rohnert Park, CA. His research interests mainly focus on the physical layer of wireless communication systems, including error performance analysis, space–time coding, channel estimation and equalization, and spread spectrum communications.

Chengshan Xiao (M'99–SM'02) received the B.S. degree from the University of Electronic Science and Technology of China, Chengdu, China, in 1987, the M.S. degree from Tsinghua University, Beijing, China, in 1989, and the Ph.D. degree from the University of Sydney, Sydney, Australia, in 1997, all in electrical engineering.

From 1989 to 1993, he was on the Research Staff and then became a Lecturer at the Department of Electronic Engineering at Tsinghua University, Beijing, China. From 1997 to 1999, he was a Senior Member of Scientific Staff at Nortel Networks, Ottawa, ON, Canada. From 1999 to 2000, he was a faculty member of the Department of Electrical and Computer Engineering at the University of Alberta, Edmonton, AB, Canada. Currently, he is an Assistant Professor at the Department of Electrical and Computer Engineering, University of Missouri, Columbia. His research interests include wireless communication networks, signal processing, and multidimensional and multirate systems. He has published extensively in these areas. He holds three U.S. patents in wireless communication area. His algorithms have been implemented into Nortel's base station radios with successful technical field trials and network integration.

Dr. Xiao is an Associate Editor for the IEEE TRANSACTIONS ON WIRELESS COMMUNICATIONS. Previously, he was an Associate Editor for the IEEE TRANSACTIONS ON VEHICULAR TECHNOLOGY, the IEEE TRANSACTIONS ON CIRCUITS AND SYSTEMS—I and the international journal *Multidimensional Systems and Signal Processing*. He is also a Co-Chair of the 2005 IEEE Globecom Wireless Communications Symposium. He is currently serving as the Vice-Chair of the IEEE Communications Society Technical Committee on Personal Communications.

Khaled Ben Letaief (S'85–M'86–SM'97–F'03) received the B.S. degree (with distinction) and M.S. and Ph.D. degrees from Purdue University, West Lafayette, IN, in 1984, 1986, and 1990, respectively.

Since January 1985 and as a Graduate Instructor in the School of Electrical Engineering at Purdue University, he has taught courses in communications and electronics. From 1990 to 1993, he was a faculty member with the University of Melbourne, Melbourne, Australia. Since 1993, he has been with the Hong Kong University of Science and Technology, Kowloon, where he is currently a Professor and Head of the Electrical and Electronic Engineering Department. He is also the Director of the Hong Kong Telecom Institute of Information Technology as well as the Director of the Center for Wireless Information Technology. His current research interests include wireless and mobile networks, broadband wireless access, space–time processing for wireless systems, wideband OFDM, and CDMA systems.

Dr. Letaief served as consultant for different organizations and is currently the founding Editor-in-Chief of the IEEE TRANSACTIONS ON WIRELESS COMMUNICATIONS. He has served on the Editorial Board of other journals, including the IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS Wireless Series (as Editor-in-Chief). He served as the Technical Program Chair of the 1998 IEEE Globecom Mini-Conference on Communications Theory, held in Sydney, Australia. He was also the Co-Chair of the 2001 IEEE ICC Communications Theory Symposium, held in Helsinki, Finland. He is the Co-Chair of the 2004 IEEE Wireless Communications, Networks, and Systems Symposium, held in Dallas, TX, and is the Vice-Chair of the International Conference on Wireless and Mobile Computing, Networking, and Communications, WiMob'05, to be held in Montreal, QC, Canada. He is currently serving as the Chair of the IEEE Communications Society Technical Committee on Personal Communications. In addition to his active research activities, he has also been a dedicated teacher committed to excellence in teaching and scholarship. He was the recipient of the Mangoon Teaching Award from Purdue University in 1990, the Teaching Excellence Appreciation Award by the School of Engineering at HKUST (four times), and the Michael G. Gale Medal for Distinguished Teaching (the highest university-wide teaching award and only one recipient/year is honored for his/her contributions). He is a Distinguished Lecturer of the IEEE Communications Society.