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Joint Tx/iterative Rx FDE for broadband direct-sequence code division multiple access

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Summary

Direct-sequence code division multiple access (DS-CDMA) is a promising uplink access technology for broadband cellular mobile communication systems. In this paper, we propose a joint transmit (Tx)/iterative receive (Rx) frequency-domain equalization (FDE) for DS-CDMA. In the proposed scheme, simple one-tap FDE at the transmitter and iterative one-tap FDE at the receiver are jointly performed using the common knowledge of channel state information (CSI). Before transmitting the signal, the transmitter predicts the degree of residual inter-chip interference (ICI) power after the iterations of Rx FDE to determine the Tx FDE weight while the receiver performs the iterative Rx FDE. In practice, it is almost impossible to share the perfect knowledge of CSI. In this paper, a Gaussian approximation of channel estimation (CE) errors is introduced. We theoretically derive a set of Tx and iterative Rx FDE weights based on the minimum mean square error (MMSE) criterion in the presence of the CE error. We show by computer simulation that the proposed scheme provides better bit error rate (BER) performance than the conventional scheme which uses iterative Rx FDE only (i.e., without Tx FDE) even in the presence of CE error. Copyright © 2010 John Wiley & Sons, Ltd.

KEY WORDS: direct-sequence code division multiple access (DS-CDMA); frequency-domain equalization (FDE); iterative interference cancellation; imperfect channel estimation

1. INTRODUCTION

Direct-sequence code division multiple access (DS-CDMA) is adopted as a multiple access technology for 3G mobile communication systems [1]. As the chip rate increases, the number of resolvable propagation paths tends to increase and the channel frequency-selectivity gets stronger, thereby producing severe inter-chip interference (ICI) and the bit error rate (BER) performance degrades [2,3]. The simple one-tap minimum mean square error (MMSE) frequency-domain equalization (FDE) can be applied to suppress the ICI and hence, improve the BER performance [3–8]. However, the performance improvement is limited due to the presence of residual ICI after Rx FDE.

Recently, iterative Rx FDE scheme [9,10] was proposed for DS-CDMA [8,11] to improve the BER performance by suppressing the residual ICI after Rx FDE. In this scheme, at first, a series of one-tap Rx FDE, despreading, and channel decoding is carried

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out. Then, the residual ICI replica is generated in frequency-domain using the log-likelihood ratio (LLR) of the channel decoder output and is subtracted from the received signal after the FDE. A series of one-tap Rx FDE, ICI cancellation, despreading, and channel decoding is repeated sufficient number of times. In each iteration stage, the Rx FDE weight is updated based on the MMSE criterion by taking into account the residual ICI power after the cancellation. It was shown [11] that the iterative Rx FDE provides much better performance than the non-iterative Rx FDE.

In this paper, in order to further improve the performance of DS-CDMA using iterative Rx FDE, we propose a joint Tx/iterative Rx FDE, in which simple one-tap FDE at the transmitter is jointly applied with the iterative Rx FDE using the common knowledge of channel state information (CSI). Before transmitting the signal, the transmitter predicts the degree of residual ICI power after the final iteration of Rx FDE to determine the Tx FDE weight while the receiver performs the iteartive Rx FDE by computing the Tx FDE weight. By using the proposed joint Tx/iterative Rx FDE, the inter-chip interference (ICI) can be significantly reduced. Assuming the time-division duplex (TDD), where the same carrier frequency is used for the transmission and reception and using the channel reciprocity [12], the transmitter can estimate the CSI of the transmitting channel by using the received signal and compute the Tx FDE weight. However, in practical systems, sharing the perfect knowledge of CSI between the transmitter and receiver is almost impossible. Therefore, in this paper, we introduce a Gaussian approximation of channel estimation (CE) errors [13–15]. We theoretically derive a set of Tx and iterative Rx FDE weights based on the MMSE criterion in the presence of CE error. We show by computer simulation that the proposed scheme provides better BER performance than the conventional scheme which uses iterative Rx FDE only even in the presence of CE error.

The rest of the paper is organized as follows. Overall system model of the propose scheme is depicted in Section 2. The principle operation is given in Section 3. In Section 4, the set of MMSE-FDE weights are derived. Section 5 shows the simulation results. Section 6 concludes this paper.

2. OVERALL SYSTEM MODEL

Figure 1 illustrates the transmitter/receiver structure of DS-CDMA using the joint Tx/iterative Rx FDE. In the proposed scheme, one-tap FDE is introduced

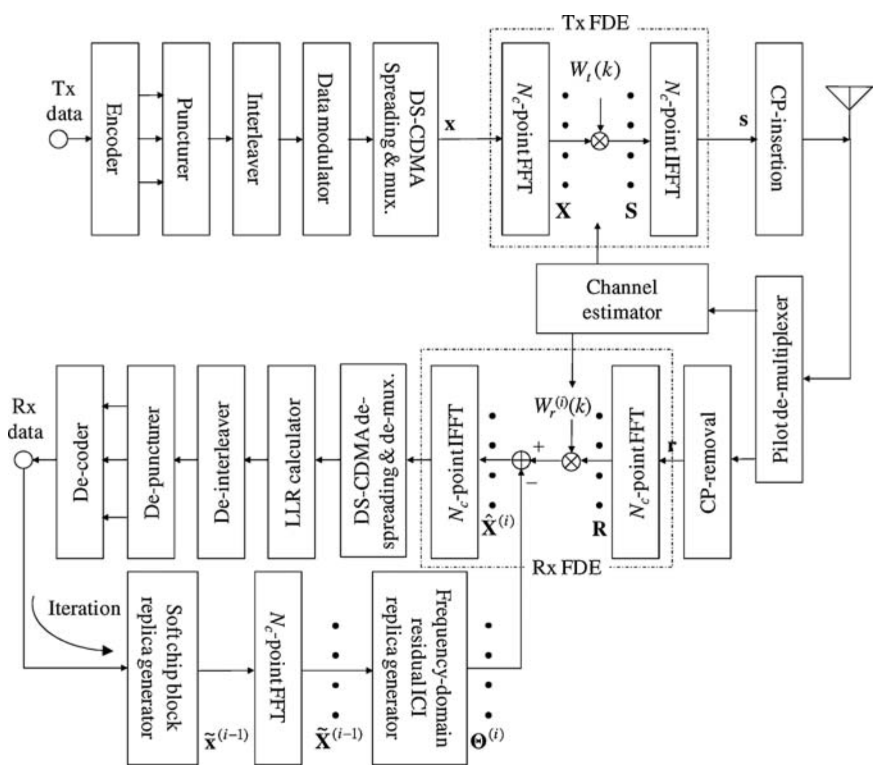

Figure 1. Transmitter/receiver structure.

at the transmitter. At the receiver, a series of one-tap Rx FDE, ICI cancellation, despreading, and decoding is carried out iteratively. In each iteration stage, soft symbol replicas are generated using the decoder output LLR sequence. Then, the Rx FDE weight is updated based on the MMSE criterion and the residual ICI replicas for cancellation are regenerated using the decoder output. As the number of iterations increases, the residual ICI is better reduced and hence, the Rx FDE weight gets closer to the maximum ratio combining (MRC) FDE weight [6].

In the proposed scheme, before transmitting the signal, the transmitter predicts the degree of residual ICI after the final iteration stage of the receiver. The predicted power of residual ICI is used for computing the Tx FDE weight. If the transmitter predicts that the receiver cannot cancel the residual ICI at all, the Tx FDE tries to reduce the frequency-selectivity of the equivalent channel seen by the receiver to reduce the residual ICI. On the other hand, if the transmitter predicts that the receiver can cancel the residual ICI, the Tx FDE tries to allocate more power to the frequencies having higher channel gain in order to improve the received signal-to-noise power ratio (SNR).

3. PRINCIPLE OF OPERATION

In this paper, chip-spaced discrete-time signal representation is used. The fast Fourier transform (FFT) (or inverse FFT (IFFT)) block size and the cyclic prefix (CP) length are denoted by N_c and N_g, respectively. The CP length is assumed to be longer than the maximum time delay difference among the propagation paths. Since the joint Tx/iterative Rx FDE is performed block-by-block, without loss of generality, we consider the transmission of one block in this section.

3.1. Transmit signal

The data-modulated symbol sequence is converted to Q parallel streams. Each stream is spread by using an orthogonal spreading code with the spreading factor SF. After multiplexing Q chip sequences, the chip block is multiplied by a scramble code. The resultant DS-CDMA chip block is represented using a vector form as $\mathbf{x}=[x(0), \ldots, x(t), \ldots, x(N_c-1)]^T$, where $\{x(t);\ t=0\sim N_c-1\}$ is given by

$$x(t) = \sum_{q=0}^{Q-1} d_q\left(\left\lfloor \frac{t}{SF} \right\rfloor\right) c_q(t \bmod \mathrm{SF}) c_{\mathrm{scr}}(t) \quad (1)$$

where $d_q(n)$ is the nth data symbol of the qth stream ($n=0\sim N_c/\mathrm{SF}-1$, $q=0\sim Q-1$), $c_q(t)$ ($q=0\sim Q-1$, $t=0\sim \mathrm{SF}-1$) is the tth chip of the qth orthogonal spreading code, and $c_{\mathrm{scr}}(t)$ is the scramble code.

N_c-point FFT is carried out on $\mathbf{x}$ to obtain the frequency-domain transmit signal $\mathbf{X}=[X(0), \ldots, X(k), \ldots, X(N_c-1)]^T$ as

$$\mathbf{X} = \mathbf{F}\mathbf{x} \quad (2)$$

with $\mathbf{F}$ being an $N_c \times N_c$ FFT matrix (or called unitary discrete Fourier transform (DFT) matrix) given as

$$\mathbf{F} = \frac{1}{\sqrt{N_c}} \begin{bmatrix} 1 & 1 & \cdots & 1 \\ 1 & e^{-j2\pi\frac{(1\times 1)}{N_c}} & \cdots & e^{-j2\pi\frac{(1\times(N_c-1))}{N_c}} \\ \vdots & \vdots & \ddots & \vdots \\ 1 & e^{-j2\pi\frac{((N_c-1)\times 1)}{N_c}} & \cdots & e^{-j2\pi\frac{((N_c-1)\times(N_c-1))}{N_c}} \end{bmatrix} \quad (3)$$

The Tx FDE weight $\mathbf{W}_t=\mathrm{diag}\{W_t(0), \ldots, W_t(k), \ldots, W_t(N_c-1)\}$ with $\mathrm{tr}[\mathbf{W}_t\mathbf{W}_t^H]=N_c$ is multiplied to $\mathbf{X}$ as

$$\mathbf{S} = [\, S(0),\ldots,S(k),\ldots,S(N_c-1)\,]^T = \mathbf{W}_t\mathbf{X} \quad (4)$$

An N_c-point IFFT is applied to $\mathbf{S}$ to obtain a time-domain transmit signal block $\mathbf{s}=[s(0), \ldots, s(t), \ldots, s(N_c-1)]^T=\mathbf{F}^H\mathbf{S}$. After the insertion of CP, the signal block is transmitted over a frequency-selective channel.

3.2. Received signal

The channel is assumed to be a chip-spaced L-path frequency-selective block-fading channel. The complex-valued path gain and the delay time of the lth path are, respectively, denoted by h_l and τ_l, $l=0\sim L-1$. The channel impulse response is given by

$$h(\tau) = \sum_{l=0}^{L-1} h_l \delta(\tau-\tau_l) \quad (5)$$

with $\sum_{l=0}^{L-1} E[|h_l|^2]=1$ and $\tau_l=l$.

The received signal block $\mathbf{r}=[r(0), \ldots, r(t), \ldots, r(N_c-1)]^T$ after the CP-removal can be expressed as

$$\mathbf{r} = \sqrt{\frac{2E_c}{T_c}}\mathbf{h}\mathbf{s} + \mathbf{n} \quad (6)$$

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where E_c and T_c are the transmit chip energy and chip duration, respectively, $\mathbf{h}$ is an $N_c \times N_c$ circulant channel matrix given by

$$\mathbf{h} = \begin{bmatrix} h_0 & & & h_{L-1} & \cdots & h_1 \\ h_1 & \ddots & & & \ddots & \vdots \\ \vdots & & h_0 & 0 & & h_{L-1} \\ h_{L-1} & & h_1 & \ddots & & \\ & \ddots & \vdots & & \ddots & \\ 0 & & h_{L-1} & \cdots & \cdots & h_0 \end{bmatrix} \quad (7)$$

and $\mathbf{n} = [n(0), \ldots, n(t), \ldots, n(N_c-1)]^T$ is the noise vector with $n(t)$ being a zero-mean additive white Gaussian noise (AWGN) having variance $2N_0/T_c$ (N_0 is the one-sided noise power spectrum density).

An N_c-point FFT is carried out on $\mathbf{r}$ to obtain the frequency-domain received signal block $\mathbf{R} = [R(0), \ldots, R(k), \ldots, R(N_c-1)]^T$ as

$$\mathbf{R} = \mathbf{Fr} = \sqrt{\frac{2E_c}{T_c}}\mathbf{HS} + \mathbf{N} = \sqrt{\frac{2E_c}{T_c}}\mathbf{HW}_t\mathbf{X} + \mathbf{N} \quad (8)$$

where $\mathbf{N} = \mathbf{Fn}$ and $\mathbf{H} = \mathbf{FhF}^H$. Due to the circulant property of $\mathbf{h}$, the channel gain matrix $\mathbf{H}$ of size $N_c \times N_c$ is diagonal. The kth diagonal element of $\mathbf{H}$ is given by

$$H(k) = \sum_{l=0}^{L-1} h_l \exp\left(-j2\pi k \frac{\tau_l}{N_c}\right) \quad (9)$$

The Rx FDE weight is multiplied to the frequency-domain received signal $\mathbf{R}$. Then, the residual ICI replica is subtracted. Here, the ith iteration stage is considered ($0 < i \leq I$). The frequency-domain received chip block $\hat{\mathbf{X}}^{(i)} = [\hat{X}^{(i)}(0), \ldots, \hat{X}^{(i)}(k), \ldots, \hat{X}^{(i)}(N_c-1)]^T$ obtained after carrying out the Rx FDE and ICI cancellation can be written as

$$\begin{aligned} \hat{\mathbf{X}}^{(i)} &= \mathbf{W}_r^{(i)}\mathbf{R} - \mathbf{\Theta}^{(i)} \\ &= \sqrt{\frac{2E_c}{T_c}}\mathbf{W}_r^{(i)}\mathbf{HW}_t\mathbf{X} - \mathbf{\Theta}^{(i)} + \mathbf{W}_r^{(i)}\mathbf{N} \end{aligned} \quad (10)$$

where $\mathbf{W}_r^{(i)} = \mathrm{diag}\{W_r^{(i)}(0), \ldots, W_r^{(i)}(k), \ldots, W_r^{(i)}(N_c-1)\}$ is an $N_c \times N_c$ diagonal Rx FDE weight matrix for the ith iteration stage. $\mathbf{\Theta}^{(i)}$ is an $N_c \times 1$ vector representing the residual ICI replica block, given as

$$\mathbf{\Theta}^{(i)} = \sqrt{\frac{2E_c}{T_c}}\left\{\mathbf{W}_r^{(i)}\mathbf{HW}_t - \mathbf{I}\right\}\tilde{\mathbf{X}}^{(i-1)} \quad (11)$$

with $\tilde{\mathbf{X}}^{(i-1)} = [\tilde{X}^{(i-1)}(0), \ldots, \tilde{X}^{(i-1)}(k), \ldots, \tilde{X}^{(i-1)}(N_c-1)]^T$ being the frequency-domain DS-CDMA chip block replica. The replica generation will be presented in the next subsection.

$\hat{\mathbf{X}}^{(i)}$ is transformed by an N_c-point IFFT into the time-domain chip block $\hat{\mathbf{x}}^{(i)}$ associated with the transmitted DS-CDMA chip block $\mathbf{x}$ as

$$\begin{aligned} \hat{\mathbf{x}}^{(i)} &= [\hat{x}^{(i)}(0), \ldots, \hat{x}^{(i)}(t), \ldots, \hat{x}^{(i)}(N_c-1)]^T = \mathbf{F}^H\hat{\mathbf{X}}^{(i)} \\ &= \sqrt{\frac{2E_c}{T_c}}\mathbf{x} \\ &\quad + \sqrt{\frac{2E_c}{T_c}}\left\{\mathbf{F}^H\mathbf{W}_r^{(i)}\mathbf{HW}_t\mathbf{F} - \mathbf{I}\right\}\left\{\mathbf{x} - \tilde{\mathbf{x}}^{(i-1)}\right\} \\ &\quad + \mathbf{F}^H\mathbf{W}_r^{(i)}\mathbf{Fn} \end{aligned} \quad (12)$$

where $\tilde{\mathbf{x}}^{(i-1)} = [\tilde{x}^{(i-1)}(0), \ldots, \tilde{x}^{(i-1)}(t), \ldots, \tilde{x}^{(i-1)}(N_c-1)]^T = \mathbf{F}^H\tilde{\mathbf{X}}^{(i-1)}$ and the first, second, and third terms denote the desired signal, residual ICI after the ith Rx FDE and ICI cancellation, and noise, respectively. Finally, despreading is carried out to obtain a decision variable for $d_q(n)$ as

$$\hat{d}_q(n) = \frac{1}{SF}\sum_{t=nSF}^{(n+1)SF-1} \hat{x}(t)c_q^*(t)c_{scr}^*(t) \quad (13)$$

The decision variable block is input to the decoder. The turbo decoder considered in this paper comprises two maximum *a posteriori* probability (MAP) decoders that are connected *via* interleaver/de-interleaver each other [16,17]. The bit LLRs are computed by the decoders. The first MAP decoder uses the output LLRs from the second MAP decoder in the previous iteration as a *priori* information to compute the *a posteriori* LLRs. Then, the second MAP decoder computes the improved *a posteriori* LLRs using the output LLRs from the first MAP decoder as *a priori* information. The resultant bit LLRs of the turbo decoder are used for generating the transmitted symbol block replica.

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3.3. Residual ICI replica generation

The bit LLR $\lambda_{n,q}^{(i)}(m)$ associated with the mth bit of the nth data symbol of the qth parallel stream in a block after the decoding is expressed as

$$\lambda_{n,q}^{(i)}(m) = \ln \frac{p^{(i)}\left(b_{n,q,m}=1\right)}{p^{(i)}\left(b_{n,q,m}=0\right)} \tag{14}$$

where $m = 0 \sim \log_2 M - 1$ (M is the modulation level), $p^{(i)}(b_{n,q,m}=1 \text{ or } 0)$ represents the *a posteriori* probability of $b_{n,q,m}=1$ or 0 after the decoding.

The frequency-domain DS-CDMA chip block replica $\tilde{\mathbf{X}}^{(i)}$ is obtained similar to that given in Reference [11]. The nth soft symbol replica of the qth parallel stream, $\tilde{d}_q^{(i)}(n)$, is given as

$$\tilde{d}_q^{(i)}(n) = \sum_{d \in Y} d \prod_{b_{n,q,m} \in d} p^{(i)}\left(b_{n,q,m} | \lambda_{n,q}^{(i)}(m)\right) \tag{15}$$

where d represents the candidate symbol having $b_{n,q,m}=0$ or 1 in the symbol set Y and $p^{(i)}(b_{n,q,m})$ are given by

$$\begin{cases} p^{(i)}\left(b_{n,q,m}=0|\lambda_{n,q}^{(i)}(m)\right) = -\frac{1}{2}\tanh\left(\frac{\lambda_{n,q}^{(i)}(m)}{2}\right) + \frac{1}{2} \\ p^{(i)}\left(b_{n,q,m}=1|\lambda_{n,q}^{(i)}(m)\right) = \frac{1}{2}\tanh\left(\frac{\lambda_{n,q}^{(i)}(m)}{2}\right) + \frac{1}{2} \end{cases} \tag{16}$$

since

$$p^{(i)}\left(b_{n,q,m}=0|\lambda_{n,q}^{(i)}(m)\right) + p^{(i)}\left(b_{n,q,m}=1|\lambda_{n,q}^{(i)}(m)\right) = 1 \tag{17}$$

According to Equations (16) and (17), $\tilde{d}_q^{(i)}(n)$ is obtained as [9–11]

$$\tilde{d}_q^{(i)}(n) = \begin{cases} \frac{1}{\sqrt{2}}\tanh\left(\frac{\lambda_{n,q}^{(i)}(0)}{2}\right) + \frac{j}{\sqrt{2}}\tanh\left(\frac{\lambda_{n,q}^{(i)}(1)}{2}\right) \text{for QPSK}, \\ \frac{1}{\sqrt{10}}\tanh\left(\frac{\lambda_{n,q}^{(i)}(0)}{2}\right)\left(2+\tanh\left(\frac{\lambda_{n,q}^{(i)}(1)}{2}\right)\right) + \frac{j}{\sqrt{10}}\tanh\left(\frac{\lambda_{n,q}^{(i)}(2)}{2}\right)\left(2+\tanh\left(\frac{\lambda_{n,q}^{(i)}(3)}{2}\right)\right) \text{for 16QAM}. \end{cases} \tag{18}$$

Using the soft symbol replica, the DS-CDMA chip replica can be generated as

$$\tilde{x}^{(i)}(t) = \sum_{q=0}^{Q-1} \tilde{d}_q^{(i)}\left(\left\lfloor \frac{t}{SF} \right\rfloor\right) c_q(t \bmod SF) c_{scr}(t) \tag{19}$$

which is the tth chip of the time-domain chip block replica $\tilde{\mathbf{x}}^{(i)}$. The frequency-domain chip block replica is generated as $\tilde{\mathbf{X}}^{(i)} = \mathbf{F}\tilde{\mathbf{x}}^{(i)}$.

4. DERIVATION OF TX AND ITERATIVE RX MMSE-FDE WEIGHTS

4.1. Channel estimation error model

The CE error is modeled as a zero-mean complex Gaussian random variable. The estimated channel gain matrices at the transmitter and the receiver can be, respectively, expressed as [13–15]

$$\begin{cases} \breve{\mathbf{H}}_t = \text{diag}\left\{ \breve{H}_t(0), \ldots, \breve{H}_t(k) \ldots, \breve{H}_t(N_c-1) \right\} = \mathbf{H} + \boldsymbol{\nu}_t \\ \breve{\mathbf{H}}_r = \text{diag}\left\{ \breve{H}_r(0), \ldots, \breve{H}_r(k) \ldots, \breve{H}_r(N_c-1) \right\} = \mathbf{H} + \boldsymbol{\nu}_r \end{cases} \tag{20}$$

where $\boldsymbol{\nu}_{t(\text{or } r)} = \text{diag}\{\nu_{t(\text{or } r)}(0), \ldots, \nu_{t(\text{or } r)}(k), \ldots, \nu_{t(\text{or } r)}(N_c-1)\}$ is a diagonal CE error matrix with each element $\nu_{t(\text{or } r)}(k)$ being a zero-mean complex Gaussian variable having the variance $2\sigma_{t(\text{or } r)}^2$. It should be noted that the CE errors at the transmitter and the receiver are assumed to be independent (i.e., $E[\nu_t(k)\,\nu_r^*(k')] = 0$ for $\forall k' = 0 \sim N_c - 1$).

4.2. Rx MMSE-FDE weight

At first, the Rx MMSE-FDE weight $\mathbf{W}_r^{(i)}$ at the ith iteration stage is derived. A concatenation of the transmit FDE and the propagation channel is viewed as an equivalent channel. Using the channel estimate matrix given by Equation (20), Equation (9) can be rewritten as

$$\mathbf{R} = \sqrt{\frac{2E_c}{T_c}}\breve{\mathbf{H}}_r\mathbf{W}_t\mathbf{X} + \sqrt{\frac{2E_c}{T_c}}\nu_r\mathbf{W}_t\mathbf{X} + \mathbf{N} \tag{21}$$

Using Equations (10), (11), (20), and (21), $\hat{\mathbf{x}}^{(i)}$ of Equation (12) can be rewritten as

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$$\hat{\mathbf{x}}^{(i)} = \sqrt{\frac{2E_c}{T_c}}\mathbf{x} + \sqrt{\frac{2E_c}{T_c}}\left\{\mathbf{F}^H\mathbf{W}_r^{(i)}\breve{\mathbf{H}}_r\mathbf{W}_t\mathbf{F}-\mathbf{I}\right\}\left\{\mathbf{x}-\tilde{\mathbf{x}}^{(i-1)}\right\} + \sqrt{\frac{2E_c}{T_c}}\mathbf{F}^H\mathbf{W}_r^{(i)}\boldsymbol{\nu}_r\mathbf{W}_t\mathbf{F}\mathbf{x} + \mathbf{F}^H\mathbf{W}_r^{(i)}\mathbf{F}\mathbf{n}. \quad (22)$$

The error vector $\mathbf{e}^{(i)}$ between the received DS-CDMA signal $\hat{\mathbf{x}}^{(i)}$ at the ith iteration stage and the transmitted DS-CDMA signal $\mathbf{x}$ is defined as

$$\mathbf{e}^{(i)} = \hat{\mathbf{x}}^{(i)} - \frac{\mathbf{x}}{\sqrt{2E_c/T_c}} = \left\{\mathbf{F}^H\mathbf{W}_r^{(i)}\breve{\mathbf{H}}_r\mathbf{W}_t\mathbf{F}-\mathbf{I}\right\}\left\{\mathbf{x}-\hat{\mathbf{x}}^{(i-1)}\right\} + \mathbf{F}^H\mathbf{W}_r^{(i)}\boldsymbol{\nu}_r\mathbf{W}_t\mathbf{F}\mathbf{x} + \frac{\mathbf{F}^H\mathbf{W}_r^{(i)}\mathbf{F}\mathbf{n}}{\sqrt{2E_c/T_c}} \quad (23)$$

The mean square error (MSE) $e^{(i)}$ at the ith iteration stage is given as

$$e^{(i)} = \mathrm{tr}\left[E(\mathbf{e}^{(i)}\mathbf{e}^{(i)H})\right] = \mathrm{tr}\left[\left\{\mathbf{W}_r^{(i)}\breve{\mathbf{H}}_r\mathbf{W}_t-\mathbf{I}\right\}\rho^{(i-1)}\left\{\mathbf{W}_r^{(i)}\breve{\mathbf{H}}_r\mathbf{W}_t-\mathbf{I}\right\}^H\right] + 2\sigma_r^2\cdot\mathrm{tr}\left[\mathbf{W}_r^{(i)}\mathbf{W}_t\mathbf{W}_t^H\mathbf{W}_r^{(i)H}\right] + \gamma^{-1}\cdot\mathrm{tr}\left[\mathbf{W}_r^{(i)}\mathbf{W}_r^{(i)H}\right] \quad (24)$$

where $\gamma = (Q/\mathrm{SF})(E_s/N_0)$ with E_s denoting the energy per transmit symbol and $\rho^{(i-1)}$ represents the residual ICI power estimated by the receiver given by [11]

$$\rho^{(i-1)} = E\left[\left|x(t)-\tilde{x}^{(i-1)}(t)\right|^2\right] = E\left[\left|d(n)-\tilde{d}^{(i-1)}(n)\right|^2\right] \approx \frac{\mathrm{SF}}{QN_c}\sum_{q=0}^{Q-1}\sum_{n=0}^{N_c/SF-1}\left|d_q(n)-\tilde{d}_q^{(i-1)}(n)\right|^2$$
$$\approx \begin{cases} \dfrac{\mathrm{SF}}{QN_c}\displaystyle\sum_{q=0}^{Q-1}\sum_{n=0}^{N_c/SF-1}\left\{1-\left|\tilde{d}_q^{(i-1)}(n)\right|^2\right\}\text{for QPSK}, \\ \dfrac{\mathrm{SF}}{QN_c}\displaystyle\sum_{q=0}^{Q-1}\sum_{n=0}^{N_c/SF-1}\left\{\frac{4}{10}\tanh\left(\frac{\lambda_{n,q}^{(i-1)}(1)}{2}\right)+\frac{4}{10}\tanh\left(\frac{\lambda_{n,q}^{(i-1)}(3)}{2}\right)+1-|\tilde{d}_q^{(i-1)}(n)|^2\right\} \\ \text{for 16QAM} \end{cases} \quad (25)$$

By solving $\partial e^{(i)}/\partial\mathbf{W}_r^{(i)} = \mathbf{0}$, the Rx MMSE-FDE weight can be derived as

$$\mathbf{W}_r^{(i)} = \rho^{(i-1)}\mathbf{W}_t^H\breve{\mathbf{H}}_r^H\left\{\mathbf{W}_t^H\breve{\mathbf{H}}_r^H\rho^{(i-1)}\breve{\mathbf{H}}_r\mathbf{W}_t + 2\sigma_r^2\cdot\mathbf{W}_t\mathbf{W}_t^H + \gamma^{-1}\cdot\mathbf{I}\right\}^{-1} \quad (26)$$

At each iteration stage, the Rx FDE weight needs to be updated. As i increases, the reliability of the symbol replica improves and, therefore, $\rho^{(i-1)}$ approaches zero.

The Rx MMSE-FDE weight requires the knowledge of the variance of CE error at the receiver, $2\sigma_r^2$, and the Tx MMSE-FDE weight matrix $\mathbf{W}_t$. $2\sigma_r^2 = E[|H_r(k)-\breve{H}_r(k)|^2]$ can be estimated [18] following Reference [19] by using a pilot signal having constant amplitude in the frequency-domain (e.g., Chu sequence). However, in this paper, ideal estimation of $2\sigma_r^2$ is assumed. The receiver must predict $\mathbf{W}_t$ (the kth entity $\sqrt{G_0(k)}$ of which is given by Equation (35)). This is done by using the receiver's channel estimate (TDD system is considered in this paper).

4.3. Tx MMSE-FDE weight

Next, the Tx MMSE-FDE weight is derived. In the proposed scheme, the transmitter predicts the degree of the residual ICI after the final iteration stage of the receiver before transmitting the signal to estimate $\mathbf{W}_r^{(I-1)}$. The predicted value of $\rho^{(I-1)}$ is denoted by ρ_{tx}. For deriving the Tx FDE weight $\mathbf{W}_t$, it is assumed that the receiver uses the Rx FDE weight $\mathbf{W}_r^{tx}$ given by

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$$\mathbf{W}_r^{tx} = \rho_{tx}\mathbf{W}_t^H\breve{\mathbf{H}}_t^H \left\{\mathbf{W}_t^H\breve{\mathbf{H}}_t^H\rho_{tx}\breve{\mathbf{H}}_t\mathbf{W}_t + 2\sigma_t^2\cdot\mathbf{W}_t\mathbf{W}_t^H + \gamma^{-1}\cdot\mathbf{I}\right\}^{-1} \tag{27}$$

which is the Rx FDE weight at the final iteration stage $(i = I-1)$ given by Equation (26), but by replacing $\rho^{(I-1)}$, $\breve{\mathbf{H}}_r$, and σ_r^2 by ρ_{tx}, $\breve{\mathbf{H}}_t$, and σ_t^2, respectively. The MSE e_{tx} between the transmit and received DS-CDMA signals for the given $\mathbf{W}_r^{tx}$ of Equation (27) is given by replacing $\rho^{(I-1)}$, $\breve{\mathbf{H}}_r$, and σ_r^2 of Equation (24) with ρ_{tx}, $\breve{\mathbf{H}}_t$, and σ_t^2, respectively, as

$$\begin{aligned} e_{tx} &= \mathrm{tr}\left[\left\{\mathbf{W}_r^{tx}\breve{\mathbf{H}}_t\mathbf{W}_t - \mathbf{I}\right\}\rho_{tx}\left\{\mathbf{W}_r^{tx}\breve{\mathbf{H}}_t\mathbf{W}_t - \mathbf{I}\right\}^H\right] \\ &\quad + 2\sigma_t^2\cdot\mathrm{tr}\left[\mathbf{W}_r^{tx}\mathbf{W}_t\mathbf{W}_t^H\mathbf{W}_r^{txH}\right] \\ &\quad + \gamma^{-1}\cdot\mathrm{tr}\left[\mathbf{W}_r^{tx}\mathbf{W}_r^{txH}\right] \end{aligned} \tag{28}$$

Substituting Equation (27) into Equation (28) gives

$$\begin{aligned} e_{tx} &= \rho_{tx}\cdot\mathrm{tr}\left[\left\{2\sigma_t^2\mathbf{W}_t\mathbf{W}_t^H + \gamma^{-1}\cdot\mathbf{I}\right\}\left\{\mathbf{W}_t^H\breve{\mathbf{H}}_t^H\rho_{tx}\breve{\mathbf{H}}_t\mathbf{W}_t + 2\sigma_t^2\cdot\mathbf{W}_t\mathbf{W}_t^H + \gamma^{-1}\cdot\mathbf{I}\right\}^{-1}\right] \\ &= \sum_{k=0}^{N_c-1}\frac{\rho_{tx}\left(2\sigma_t^2|W_t(k)|^2 + \gamma^{-1}\right)}{\rho_{tx}|W_t(k)|^2|H(k)|^2 + 2\sigma_t^2|W_t(k)|^2 + \gamma^{-1}} \end{aligned} \tag{29}$$

The Tx MMSE-FDE weight is the one that minimizes e_{tx} of Equation (29) under the power constraint of Equation (6). This problem can be rewritten, by replacing $|W_t(k)|^2$ by $G(k)$ for simplicity, as

$$\begin{aligned} &\text{minimize } e_{tx} = \sum_{k=0}^{N_c-1}\frac{\rho_{tx}\left(2\sigma_t^2G(k) + \gamma^{-1}\right)}{\rho_{tx}G(k)\,|H(k)|^2 + 2\sigma_t^2G(k) + \gamma^{-1}} \\ &\text{subject to } \sum_{k=0}^{N_c-1}G(k) - N_c = 0 \\ &\text{with } \quad G(k)\geq 0 \quad \textit{for} \quad \forall k = 0\sim N_c-1 \end{aligned} \tag{30}$$

We define a Lagrange function $J(\{G(k); k = 0\sim N_c-1\}, \mu, \{\psi_k; k = 0\sim N_c-1\})$ as

$$\begin{aligned} J(\{G(k)\}, \mu, \{\psi_k\}) &= \sum_{k=0}^{N_c-1}\frac{\rho_{tx}\left(2\sigma_t^2G(k) + \gamma^{-1}\right)}{\rho_{tx}G(k)\,|H(k)|^2 + 2\sigma_t^2G(k) + \gamma^{-1}} \\ &\quad + \mu\cdot\left\{\sum_{k=0}^{N_c-1}G(k) - N_c\right\} + \sum_{k=0}^{N_c-1}\psi_k\cdot\{-G(k) - 0\} \end{aligned} \tag{31}$$

The solution of Equation (30), $\{G_0(k); k = 0\sim N_c-1\}$, satisfies the following KKT condition [20–22]:

$$\begin{aligned} \left.\frac{\partial J(\{G(k)\}, \mu, \{\psi_k\})}{\partial G(k)}\right|_{G(k)=G_0(k)} &= \frac{-\rho_{tx}^2\,|H(k)|^2\,\gamma^{-1}}{\left\{\rho_{tx}G(k)\,|H(k)|^2 + 2\sigma_t^2G(k) + \gamma^{-1}\right\}^2} + \mu - \psi_k \\ &= 0 \end{aligned} \tag{32}$$

$$\sum_{k=0}^{N_c-1}G_0(k) - N_c = 0 \tag{33}$$

$$\psi_k \geq 0,\ G_0(k)\geq 0,\ \psi_kG_0(k) = 0 \quad \text{for} \quad \forall k \tag{34}$$

The solution that satisfies the above KKT condition is given by

$$G_0(k) = \max\left[\left\{\frac{1}{\sqrt{\mu}}\frac{\rho_{tx}\gamma^{-1/2}\left|\breve{H}_t(k)\right| - \gamma^{-1}}{\rho_{tx}\left|\breve{H}_t(k)\right|^2 + 2\sigma_t^2}\right\}, 0\right] \tag{35}$$

The Tx MMSE-FDE weight $W_t(k)$ is given by $\sqrt{G_0(k)}$. μ is chosen so as to satisfy $\mathrm{tr}[\mathbf{W}_t\mathbf{W}_t^H] = N_c$.

Tx MMSE-FDE requires the knowledge of $2\sigma_t^2$ and ρ_{tx}. We consider TDD and thus, similar to $2\sigma_r^2$ in Equation (26), $2\sigma_t^2$ can also be estimated by using the received pilot signals. In this paper, we assume ideal estimation of $2\sigma_t^2$ at the transmitter. The ρ_{tx} that minimizes the average BER is found by preliminary computer simulation at each average transmit E_b/N_0 for the given channel power delay profile, spreading factor, and coding rate.

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5. SIMULATION RESULTS

The achievable BER performance of DS-CDMA using the proposed scheme is evaluated by computer simulation. The propagation channel is assumed to be an $L = 16$-path frequency-selective block Rayleigh fading channel having uniform power delay profile. For the coded performance comparison, a turbo encoder using two (13, 15) recursive systematic convolutional (RSC) encoders is used. A 2048 bit length codeword with the coding rate $R = 1/2$ is generated by puncturing the parity bit sequences. The decoder consists of two log-MAP decoders. In both coded and uncoded cases, the number of iterations in the receiver is set to $I = 6$. $N_c = 256$ and $N_g = 32$ are considered. QPSK data modulation is used. Before the signal transmission, the Tx FDE weight is computed by Equation (35) after obtaining $\breve{\mathbf{H}}_t$, σ_t^2, and γ. At the receiver, before the signal reception, $\breve{\mathbf{H}}_r$, σ_r^2, and γ are obtained and the Tx FDE weight is computed. Then, the Rx FDE weight $\mathbf{W}_t$ is computed using Equation (26). These can be done in TDD systems.

In multicode DS-CDMA, the BER performance for the given transmit bit energy-to-noise power spectrum density ratio E_b/N_0 $(=0.5/R(\mathrm{SF}E_c/N_0)(1 + N_g/N_c))$ does not depend on the relation between SF and C; it only depends on the equivalent spreading factor $\mathrm{SF}_{\mathrm{eq}} = \mathrm{SF}/C$. Therefore, for the sake of simplicity, we consider single-code (i.e., $C = 1$) only in the following computer simulation.

5.1. Average BER performance With ideal CE

Achievable average BER performance using the proposed scheme assuming ideal CE is plotted in Figure 2 as a function of average transmit E_b/N_0 for both uncoded and coded cases. For comparison, average BER performance using the conventional iterative Rx FDE only [11] is also plotted. Spreading factor SF is set to 1–16.

It can be seen that the proposed scheme provides better BER performance than the conventional scheme. As SF increases, the performance improvement by the proposed scheme gets larger. When the value of SF is small, both the proposed and conventional schemes suffer from the residual ICI after the iterative FDE. As SF gets larger, the residual ICI can be further suppressed through the despreading process. However, in the conventional scheme, even when SF is sufficiently large, the BER performance improvement is limited since the same power is given to all the frequencies for the transmission. On the other hand, in the proposed scheme, when SF is sufficiently large, Tx MMSE-FDE allocates more power to frequencies having good condition. Therefore, the transmit signal spectrum is distorted and this produces large ICI; however, the resultant ICI can be sufficiently suppressed through the despreading process.

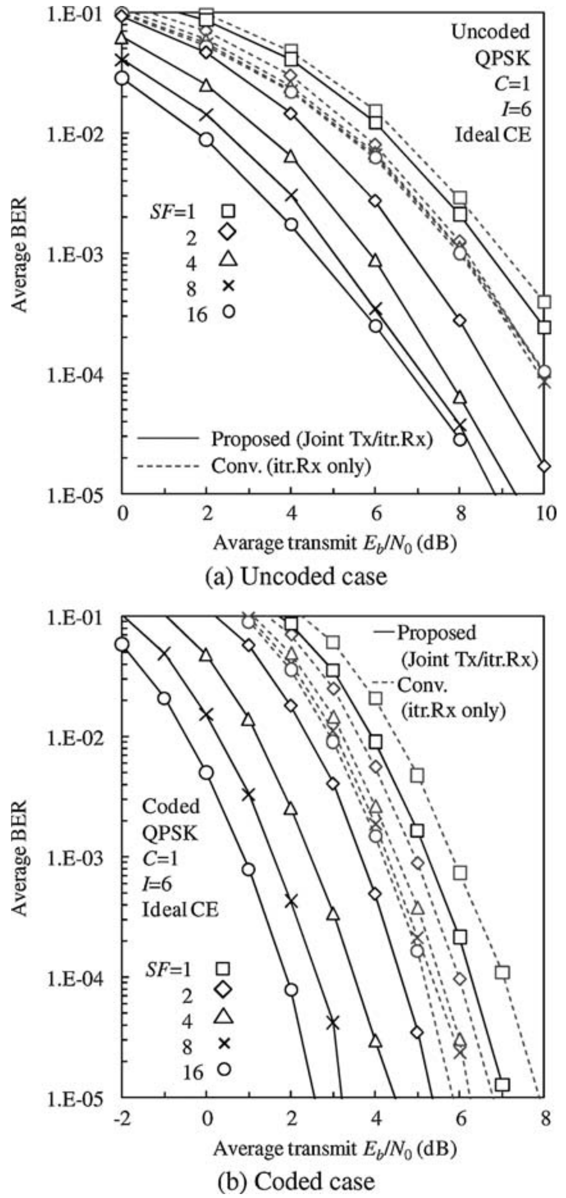

Figure 2. Average BER performance (a) uncoded case (b) coded case.

In the coded case, the performance improvement by the proposed scheme is larger than that in the uncoded case. When SF = 1, for example, the required E_b/N_0

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for achieving $\mathrm{BER} = 10^{-3}$ can be reduced by about 0.8 dB by using the proposed scheme in the coded case; however, it is only about 0.3 dB in the uncoded case. This is because the decoding process is involved in the iterative Rx FDE and hence, more accurate residual ICI replicas can be generated.

5.2. Impact of spreading factor, SF

Figure 3 shows the required E_b/N_0 for achieving $\mathrm{BER} = 10^{-3}$ as a function of SF. Also shown are the required E_b/N_0 of the iterative Rx FDE only [11] and that of the non-iterative Rx FDE only [5]. It should be noted that in the case of non-iterative Rx FDE only, turbo decoding is iterated *I* times for fair comparison in terms of coding gain. Ideal CE is assumed.

It can be seen from Figure 3 that as SF increases, the required E_b/N_0 reduces. However, if the iterative (or non-iterative) Rx FDE only is used, the required E_b/N_0 reduction is saturated. For iterative Rx FDE only, the required E_b/N_0 becomes almost the same with $\mathrm{SF} \geq 2$ (4) for uncoded (coded) case. On the other hand, for the proposed scheme, the required E_b/N_0 reduces as SF increases. Therefore, the required E_b/N_0 differences between the proposed and conventional schemes are significant for a large SF. For example, when $\mathrm{SF} = 16$, the required E_b/N_0 difference between the proposed scheme and the iterative Rx FDE only is about 3.5 dB, while it is about 0.8 dB when $\mathrm{SF} = 1$.

5.3. Impact of imperfect CE

Next, we show how the CE error affects the BER performance of the proposed scheme. Figure 4 shows the average BER of the proposed scheme as a function of CE error variance $2\sigma_r^2$ at the receiver. We assume that CE error variance at the transmitter of the proposed scheme is two times larger than that at the receiver, i.e., $2\sigma_t^2 = 4\sigma_r^2$. For comparison, the average BER of the iterative Rx FDE only assuming the same CE variance $2\sigma_r^2$ is also plotted. Spreading factor SF is set to 1–16. The transmit E_b/N_0 is set as 6 dB (3 dB) for the uncoded (coded) case.

As $2\sigma_r^2$ gets larger, the BER performance is degraded. It can be seen that the proposed scheme provides better BER than the conventional scheme irrespective of the degree of SF. When SF is small, the BER difference between the proposed and conventional schemes is small for a negligible amount of CE error (e.g., $2\sigma_r^2 = 10^{-4}$). However, even for a large $2\sigma_r^2$, the proposed scheme provides almost the same BER as the

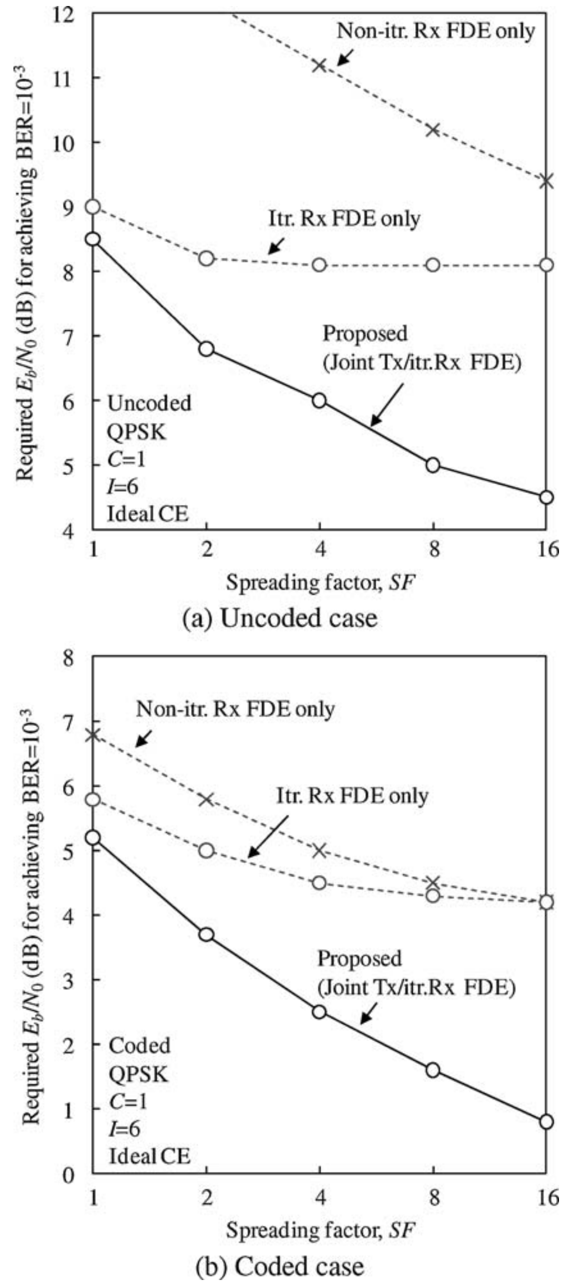

Figure 3. Required E_b/N_0 for achieving $\mathrm{BER} = 10^{-3}$ (a) uncoded case (b) coded case.

conventional scheme. This indicates that the proposed scheme is robust against the CE error at the transmitter (note that the CE error is present at the transmitter in the proposed scheme). As SF increases, the BER difference gets larger. The big advantage of the proposed scheme over the conventional scheme can be seen for a large SF even in the presence of CE error.

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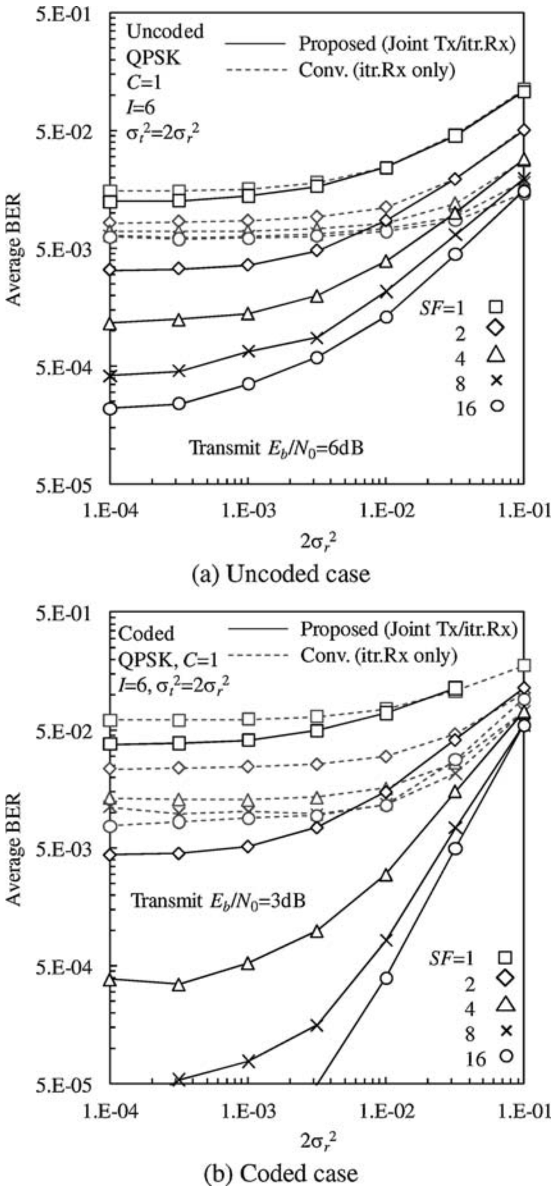

Figure 4. Average BER *versus* CE error variance (a) uncoded case (b) coded case.

6. CONCLUSION

In this paper, we proposed joint Tx/iterative Rx FDE for DS-CDMA. In the proposed scheme, the transmitter predicts the degree of the residual ICI after the final iteration stage of the receiver for computing the Tx FDE weight. The set of Tx and Rx MMSE-FDE weights were derived by taking into account the variance of CE error. We showed by computer simulation that the proposed scheme can significantly improve the BER performance of DS-CDMA in the presence of CE error.

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