Downlink DS-CDMA Transmission with Joint MMSE Equalization and ICI Cancellation

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***Abstract*— The bit error rate (BER) performance of downlink DS-CDMA in a frequency-selective fading channel can be significantly improved by the use of frequency-domain equalization (FDE) based on minimum mean square error (MMSE) criterion. However, the presence of residual inter-chip-inference (ICI) after FDE produces the orthogonality distortion among the spreading codes and the BER performance degrades as the number of multiplex order increases. In this paper, we propose a joint MMSE-FDE and ICI cancellation to improve the BER performance of the DS-CDMA downlink. In the proposed scheme, the residual ICI replica in the frequency-domain is generated and subtracted from each frequency component of the received signal after MMSE-FDE. The MMSE weight at each iteration is derived taking into account the residual ICI. The effect of the proposed ICI cancellation scheme is confirmed by computer simulation.**

Keyword; DS-CDMA, frequency-domain equalization, MMSE, ICI cancellation

I. INTRODUCTION

The wireless channel is composed of many propagation paths with different time delays, producing frequency-selective multipath fading [1]. Direct-sequence code division multiple access (DS-CDMA) can exploit the channel frequency-selectivity by the use of coherent rake combining, that resolves the propagation paths having different time delays and coherently combines them to obtain the path diversity gain [2]. Wideband DS-CDMA [3] has been adopted as a wireless access technique in the 3rd generation (3G) mobile communication systems for data transmissions of up to a few Mbps. Recently, demands for broadband services in mobile communication systems are becoming stronger and stronger and a lot of research attention has been paid to the development of the next generation mobile communication systems, that support much higher data rate services than 3G (e.g., higher than a few tens of Mbps) [4]. However, the bit error rate (BER) performance of DS-CDMA with rake combining may significantly degrade due to strong inter-path interference (IPI). An advanced equalization technique is inevitable to overcome a severe frequency-selective fading channel. Although a time-domain equalizer using maximum likelihood sequence estimation (MLSE) is known to provide a good BER performance, its computational complexity exponentially grows with the number of multipath fading.

Recently, it has been shown [5-8] that frequency-domain equalization (FDE) based on minimum mean square error (MMSE) criterion can replace the conventional rake combining with much improved BER performance for the DS-CDMA signal reception over a severe frequency-selective channel.

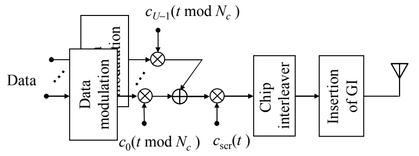
In this paper, we consider the orthogonal multicode DS-CDMA downlink signal transmission, which requires much higher rate transmission than the uplink as in high speed downlink packet access (HSDPA) [9]. Although FDE can significantly improve the downlink BER performance, the presence of a residual inter-chip interference (ICI) after FDE distorts the orthogonality among the spreading codes and the downlink BER performance degrades as the number of multiplex order increases. In this paper, we propose a joint MMSE-FDE and frequency-domain ICI cancellation to improve the BER performance of the DS-CDMA downlink signal transmission. In the proposed scheme, the residual ICI replica in the frequency-domain is generated and subtracted from each frequency component of the received signal after MMSE-FDE. The MMSE weight at each iteration is derived taking into account the residual ICI. The effect of the frequency-domain ICI cancellation is evaluated by computer simulation.

II. TRANSMISSION SYSTEM MODEL

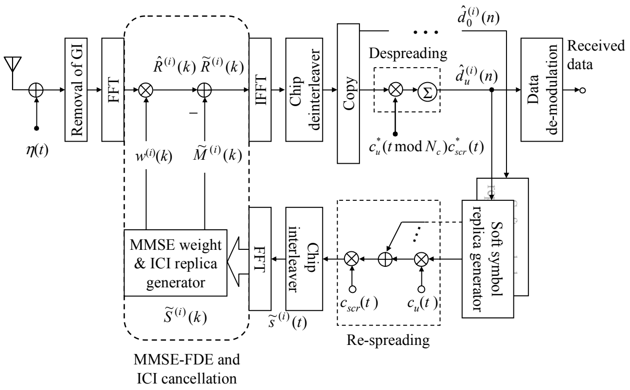
A. Overall transmission system model

The transmission system model for DS-CDMA with joint MMSE-FDE and ICI canceller is illustrated in Fig.1. At the base station transmitter, the uth user's binary data sequence, u=0~(U−1), is transformed into a data modulated symbol sequence $d_u(n)$ and then spread by multiplying an orthogonal spreading sequence $c_u(t)$. The resultant U chip sequences are multiplexed and further multiplied by a common scramble sequence $c_{scr}(t)$ to make the resultant multicode DS-CDMA signal white-noise like. Then, the orthogonal multicode DS-CDMA signal is divided into a sequence of blocks of N_c chips each and then, the last N_g chips of each block are copied as a cyclic prefix and inserted into the guard interval (GI) at the beginning of each block.

The GI-inserted orthogonal multicode DS-CDMA signal is transmitted over a frequency-selective fading channel and is received at a receiver. After the removal of the GI, the received chip sequence is decomposed by N_c-point fast Fourier transform (FFT) into N_c subcarrier components (the terminology "subcarrier" is used for explanation purpose only although subcarrier modulation is not used). MMSE-FDE is carried out, then ICI cancellation is performed in the frequency-domain. Inverse FFT (IFFT) is applied to obtain the time-domain received chip sequence for despreading and soft decision. A series of MMSE-FDE, ICI cancellation, IFFT, despreading and soft decision is repeated a sufficient number of times. Finally, data-demodulation is carried out to obtain the received data.



(a) Transmitter



(b) Receiver

Figure 1. Transmitter/receiver strucuture.

B. Tranmit and receive signals

Throughout this paper, chip-spaced time representation of the transmitted signals is used. Without loss of generality, a transmission of U data symbol sequences $\{d_u(n); n=0\sim N_c/SF-1\}$ is considered, where N_c and SF are chosen so that the value of N_c/SF becomes an integer. The spread signal chip sequence $\{\hat{s}(t); t=-N_g\sim N_c-1\}$, to be transmitted after the GI insertion, for one block of N_c+N_g chips can be expressed, using the equivalent lowpass representation, as

$$\hat{s}(t)=\sqrt{2E_c/T_c}\,s(t \bmod N_c) \quad , (1)$$

where E_c and T_c denote the chip energy and the chip duration, respectively, and $s(t)$ is given by

$$s(t)=\left[\sum_{u=0}^{U-1} d_u\left(\lfloor t/SF \rfloor\right)c_u(t \bmod SF)\right]c_{scr}(t) \quad (2)$$

with $|c_u(t)|=|c_{scr}(t)|=1$ for $t=0\sim(N_c-1)$, where $\lfloor x \rfloor$ represents the largest integer smaller than or equal to x.

The propagation channel is assumed to be a frequency-selective block fading channel having chip-spaced L discrete paths, each subjected to independent fading. The assumption of block fading means that the path gains remain constant over at least one block duration. The impulse response $h(t)$ of a multipath channel can be expressed as [10]

$$h(t)=\sum_{l=0}^{L-1} h_l\delta(t-\tau_l) \quad , (3)$$

where h_l and τ_l are the complex-valued path gain and time delay of the lth path ($l=0\sim L-1$), respectively, with $\sum_{l=0}^{L-1} E[|h_l|^2]=1$ ($E[.]$ denotes the ensemble average operation). The received chip sequence $\{r(t); t=-N_g\sim N_c-1\}$ can be expressed as

$$r(t)=\sum_{l=0}^{L-1} h_l\hat{s}(t-\tau_l)+\eta(t) \quad , (4)$$

where $\eta(t)$ is a zero-mean complex Gaussian process with a variance of $2N_0/T_c$ with N_0 being the single-sided power spectrum density of the additive white Gaussian noise (AWGN) process.

C. MMSE-FDE and ICI cancellation

A joint MMSE-FDE and ICI cancellation is repeated in an iterative fashion. Below, the ith iteration is described.

After the removal of the GI from the received chip sequence $r(t)$, N_c-point FFT is applied to decompose $\{r(t); t=0\sim N_c-1\}$ into N_c subcarrier components $\{R(k); k=0\sim N_c-1\}$. The kth subcarrier component $R(k)$ is written as

$$\begin{aligned}R(k)&=\sum_{t=0}^{N_c-1} r(t)\exp\left(-j2\pi k\frac{t}{N_c}\right)\\&=H(k)S(k)+\Pi(k)\end{aligned} \quad , (5)$$

where $S(k)$, $H(k)$ and $\Pi(k)$ are the kth subcarrier component of the transmitted signal sequence $\{s(t); t=0\sim N_c-1\}$ of N_c chips, the channel gain and the noise component due to the AWGN, respectively. They are given by

$$\begin{cases}S(k)=\sum_{t=0}^{N_c-1} s(t)\exp\left(-j2\pi k\dfrac{t}{N_c}\right)\\ H(k)=\sqrt{\dfrac{2E_c}{T_c}}\sum_{l=0}^{L-1} h_l\exp\left(-j2\pi k\dfrac{\tau_l}{N_c}\right)\\ \Pi(k)=\sum_{t=0}^{N_c-1} \eta(t)\exp\left(-j2\pi k\dfrac{t}{N_c}\right)\end{cases} \quad . (6)$$

MMSE-FDE is carried out as follows:

$$\begin{aligned}\hat{R}(k)&=w^{(i)}(k)R(k)\\&=S(k)\hat{H}^{(i)}(k)+\hat{\Pi}^{(i)}(k)\end{aligned} \quad (7)$$

with

$$\begin{cases}\hat{H}^{(i)}(k)=w^{(i)}(k)H(k)\\ \hat{\Pi}^{(i)}(k)=w^{(i)}(k)\Pi(k)\end{cases} \quad , (8)$$

where $w^{(i)}(k)$ is the equalization weight at the *i*th iteration and $\hat{H}^{(i)}(k)$ and $\hat{\Pi}^{(i)}(k)$ are the equivalent channel gain and the noise component, after performing MMSE-FDE at the *i*th iteration, respectively. The equalization weight is derived in Sect. IV. ICI cancellation is performed as

$$\tilde{R}^{(i)}(k) = \hat{R}^{(i)}(k) - \tilde{M}^{(i)}(k) \quad ,(9)$$

where $\tilde{M}^{(i)}(k)$ is the residual ICI replica, and given by [11]

$$\tilde{M}^{(i)}(k) = \begin{cases} 0 & \text{for } i=0 \\ \left\{\hat{H}^{(i)}(k) - A^{(i)}\right\}\tilde{S}^{(i)}(k) & \text{for } i>0 \end{cases}, (10)$$

where $\tilde{S}^{(i)}(k)$ is the *k*th frequency component of the transmitted chip replica, which is generated by feeding back the (*i*−1)th ICI cancellation result and $A^{(i)}$ is given by

$$A^{(i)} = \frac{1}{N_c}\sum_{k'=0}^{N_c-1}\hat{H}^{(i)}(k') \text{. (11)}$$

N_c-point IFFT is applied to transform the frequency-domain signal $\{\tilde{R}^{(i)}(k); k = 0 \sim N_c - 1\}$ into time-domain chip sequence $\{\tilde{r}^{(i)}(t); t = 0 \sim N_c - 1\}$:

$$\tilde{r}^{(i)}(t) = \frac{1}{N_c}\sum_{k=0}^{N_c-1}\tilde{R}^{(i)}(k)\exp\left(j2\pi t\frac{k}{N_c}\right). \qquad (12)$$

Then, despreading is carried out on $\tilde{r}^{(i)}(t)$ to obtain the *u*th user's decision variable for data demodulation on $d_u(n)$, giving

$$\hat{d}_u^{(i)}(n) = \frac{1}{SF}\sum_{t=nSF}^{(n+1)SF-1}\tilde{r}^{(i)}(t)c_u^*(t \bmod SF)c_{scr}^*(t)\,, \qquad (13)$$

$\hat{d}_u^{(i)}(n)$ is the soft decision variable associated with $d_u(n)$ after the *i*th iteration.

III. ICI REPLICA GENERATION

The soft decision variable is used to generate the replica $\{\tilde{s}^{(i)}(t)\}$ of the transmitted chip sequence. Using the *u*th user's soft decision variable $\hat{d}_u^{(i-1)}(n)$, the log-likelihood ratio (LLR) for the *x*th bit in the *n*th symbol (n=0~N_c/SF−1), where x=0~$\log_2 M$−1 and *M* is the modulation level, can be computed as [12]

$$L_x(n) = \ln\left(\frac{p(b_{n,x}=1)}{p(b_{n,x}=0)}\right) \approx \frac{\left|\hat{d}_u^{(i-1)}(n) - A^{(i-1)}d_{b_{n,x}=0}^{\min}\right|^2}{2\hat{c}^2} - \frac{\left|\hat{d}_u^{(i-1)}(n) - A^{(i-1)}d_{b_{n,x}=1}^{\min}\right|^2}{2\hat{c}^2} \quad ,(14)$$

where $d_{b_{n,x}=0}^{\min}$ (or $d_{b_{n,x}=1}^{\min}$) is the most probable symbol that has the minimum Euclidean distance from $\hat{d}_u^{(i-1)}(n)$ among all candidate symbols $\{d\}$ with $b_{n,x}=0$ (or 1). $2\hat{\sigma}^2$ is the variance of the noise plus residual ICI.

The soft decision symbol $\{\tilde{d}_u(n)$;n=0~N_c/SF−1} can be obtained from

$$\tilde{d}_u(n) = \sum_{d\in D} d\prod_{b_{n,x}\in d} p(b_{n,x}) \quad , (15)$$

where *D* is the set of $\{d\}$ and $p(b_{n,x}=0)$ and $p(b_{n,x}=1)$ are given, from Eq. (14), as

$$\begin{cases} p(b_{n,x}=0) = -\dfrac{1}{2}\tanh\left(\dfrac{L_x(n)}{2}\right) + \dfrac{1}{2} \\ p(b_{n,x}=1) = \dfrac{1}{2}\tanh\left(\dfrac{L_x(n)}{2}\right) + \dfrac{1}{2} \end{cases} \qquad (16)$$

since $p(b_{n,x}=1) + p(b_{n,x}=0) = 1$.

For QPSK data modulation and 16-quadrature amplitude modulation (16QAM), $\tilde{d}_u(n)$ becomes

$$\begin{cases} \tilde{d}_u^{QPSK}(n) = \dfrac{1}{\sqrt{2}}\tanh\left(\dfrac{L_0(n)}{2}\right) + j\dfrac{1}{\sqrt{2}}\tanh\left(\dfrac{L_1(n)}{2}\right) \\ \tilde{d}_u^{16QAM}(n) = \dfrac{1}{\sqrt{10}}\tanh\left(\dfrac{L_0(n)}{2}\right)\left\{2 + \tanh\left(\dfrac{L_1(n)}{2}\right)\right\} \\ \qquad + j\dfrac{1}{\sqrt{10}}\tanh\left(\dfrac{L_2(n)}{2}\right)\left\{2 + \tanh\left(\dfrac{L_3(n)}{2}\right)\right\} \end{cases} .(17)$$

The replica $\{\tilde{s}^{(i)}(t); t = 0 \sim N_c - 1\}$ of the transmitted chip sequence *s*(*t*) is generated as

$$\tilde{s}^{(i)}(t) = \left[\sum_{u=0}^{U-1}\tilde{d}_u\left(\lfloor t/SF \rfloor\right)c_u(t \bmod SF)\right]c_{scr}(t) \quad .(18)$$

Then, N_c-point FFT is applied to decompose the replica $\tilde{s}^{(i)}(t)$ into N_c subcarrier components $\{\tilde{S}^{(i)}(k)$;k=0~(N_c−1)} as

$$\tilde{S}^{(i)}(k) = \sum_{t=0}^{N_c-1}\tilde{s}^{(i)}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) .(19)$$

Substituting Eq. (19) into Eq. (10), we obtain the frequency-domain ICI replica $\tilde{M}^{(i)}(k)$.

IV. MMSE WEIGHT DERIVATION

To derive the MMSE weight, taking into account the residual ICI, we define the equalization error *e*(*k*) between the frequency component $\{\tilde{R}^{(i)}(k); k=0\sim N_c-1\}$ after the ICI cancellation and the transmitted frequency component $\{S(k); k=0\sim N_c-1\}$ as

$$e(k)=\tilde{R}^{(i)}(k)-A^{(i)}S(k)\ ,(20)$$

where $A^{(i)}S(k)$ is used as a reference signal since $E[\tilde{R}^{(i)}(k)]=A^{(i)}S(k)$ (the residual ICI is assumed to be zero-mean). $w^{(i)}(k)$ is the weight that minimizes the mean square error (MSE) $E[|e(k)|^2]$ for the given $H(k)$, i.e., $\partial E[|e(k)|^2]/\partial w^{(i)}(k)=0$.

Since $\Pi(k)$ is a zero-mean complex Gaussian process with variance $2\sigma^2(=2N_0N_c/T_c)$, $E[|e(k)|^2]$ is given by

$$E[|e(k)|^2]=\rho^{(i)}\left|w^{(i)}(k)H(k)-A^{(i)}\right|^2+2\sigma^2\left|w^{(i)}(k)\right|^2\ ,(21)$$

where

$$\rho^{(i)}=\sum_{t=0}^{N_c-1}\left\{\left|s(t)\right|^2-\left|\tilde{s}^{(i)}(t)\right|^2\right\}\ .(22)$$

Since $s(t)$ is unknown, we use the hard decision chip sequence replica $\bar{s}^{(i)}(t)$ instead of $s(t)$ in Eq. (22). The tentative data decision is first performed on $\{\hat{d}_u^{(i-1)}(n)\}$ of Eq. (13) to obtain the hard decision symbol replica $\bar{d}_u(\lfloor t/SF\rfloor)$. Then, $\bar{s}^{(i)}(t)$ is generated as

$$\bar{s}^{(i)}(t)=\left[\sum_{u=0}^{U-1}\bar{d}_u(\lfloor t/SF\rfloor)c_u(t\bmod SF)\right]c_{scr}(t)\ .(23)$$

Replacing $s(t)$ in Eq. (22) by $\bar{s}^{(i)}(t)$, we have

$$\rho^{(i)}\approx\sum_{t=0}^{N_c-1}\left\{\left|\bar{s}^{(i)}(t)\right|^2-\left|\tilde{s}^{(i)}(t)\right|^2\right\}.(24)$$

From $\partial E[|e(k)|^2]/\partial w^{(i)}(k)=0$, the following MMSE weight is obtained:

$$w^{(i)}(k)=\frac{H^*(k)}{\rho^{(i)}\left|H(k)\right|^2+2\sigma^2}\ .(25)$$

V. COMPUTER SIMULATION

We assume QPSK data modulation and 16QAM, FFT block size of N_c=256 chips and a GI of N_g=32 chips. The channel is assumed to be a frequency-selective block Rayleigh fading channel having a chip-spaced 16-path uniform power delay profile (i.e., $E[|h_l|^2]=1/L$ for all l). Perfect chip timing and ideal channel estimation are assumed.

The simulated BER performance of frequency-domain ICI cancellation is plotted in Fig. 2 for $SF=U=1$ as a function of the average received bit energy-to-AWGN noise power spectrum density ratio E_b/N_0, defined as $E_b/N_0=(1/\log_2 M)SF(E_c/N_0)(1+N_g/N_c)$. For comparison, the theoretical lower bound [7] and the BER performance using $\rho^{(i)}$ optimized by computer simulation at each iteration [11] are also plotted. The BER performance with i=0 corresponds to the case of MMSE-FDE without ICI cancellation. It can be seen from Fig. 2 that the proposed ICI cancellation can significantly improve the BER performance and achieve almost the same BER performance as the ICI cancellation using the computer-optimized $\rho^{(i)}$. When i=3, for QPSK, the E_b/N_0 reduction for BER=10^{-4} from the case without ICI cancellation is as much as 4.9 dB and the BER performance gets close to the theoretical lower bound by about 1.9 dB (including a 0.5 dB loss due to the GI insertion). For 16QAM (see Fig. 2(b)), the Euclidean distance between different symbols is shorter and hence, decision error due to the residual ICI is more likely than for QPSK. ICI cancellation is found to be very effective to improve the BER performance even for 16QAM. An E_b/N_0 reduction of as much as 5.4 dB can be achieved for BER=10^{-4}.

The BER performance for the case of SF=16 is plotted in Fig. 3 with U as a parameter. When U=1, the BER performance is better for SF=16 than for SF=1 (compare Figs. 2 and 3) since the residual ICI can be better suppressed by the despreading process. As U increases, the BER performance degrades for the no ICI cancellation case. This is because a severe orthogonality distortion is produced by the residual ICI. The use of ICI cancellation can significantly improve the BER performance. When U=16, the E_b/N_0 reduction from the no ICI cancellation case is as much as 6.9 (8.3) dB for QPSK (16QAM).

VI. CONCLUSION

In this paper, joint frequency-domain ICI cancellation and MMSE-FDE was proposed. The MMSE weight taking into account the residual ICI was derived. The BER performance with the proposed ICI cancellation was evaluated by computer simulation. It was found that, when $SF=U$=16, the E_b/N_0 reduction for achieving BER=10^{-4} from the no cancellation case is as much as about 6.9 (8.3) dB for QPSK (16QAM). In this paper, ideal channel estimation was assumed. The performance evaluation assuming the practical channel estimation is left for a future work.

REFERENCE

[1] F. Adachi, "Wireless past and future - evolving mobile communications systems," IEICE Trans. Fundamentals, Vol. E83-A, pp.55-60, Jan. 2001.

[2] W. C., Jakes Jr., Ed., *Microwave mobile communications*, Wiley, New York, 1974.

[3] F. Adachi, M. Sawahashi, and H. Suda, "Wideband DS-CDMA for next generation mobile communications systems," IEEE Commun. Mag., vol. 36, pp.56-69, Sept. 1998.

[4] Y. Kim et al., "Beyond 3G : vision, requirements., and enabling technologies," IEEE Commun. Mag., Vol. 41, pp.120-124, Mar. 2003.

[5] F. W. Vook, T. A. Thomas, and K. L. Baum, "Cyclic-prefix CDMA with antenna diversity," Proc. IEEE VTC 2002 Spring, pp. 1002-1006, May 2002.

[6] F. Adachi, T. Sao, and T. Itagaki, "Performance of multicode DS-CDMA using frequency domain equalization in a frequency selective fading channel," Electronics Letters, Vol. 39, pp.239-241, Jan. 2003.

[7] F. Adachi and K. Takeda, “Bit error rate analysis of DS-CDMA with joint frequency-domain equalization and antenna diversity combining,” IEICE Trans. Commun., Vol. E87-B, pp. 2991-3002, Oct. 2004.

[8] I. Martoyo, G. M.A. Sessler, J. Luber and F. K. Jondral, “Comparing equalizers and multiuser detections for DS-CDMA downlink systems,” Proc. IEEE VTC 2004-Spring, May 2004.

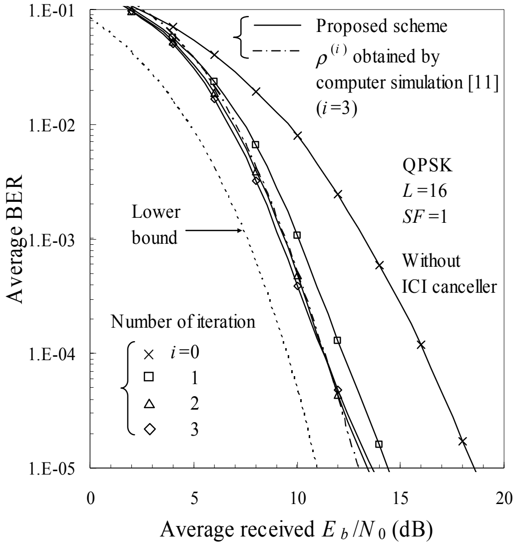
[9] *Physical layer aspects of UTRA High Speed Downlink Packet Access* (Release 4), Mar. 2001. 3GPP, 3G TR25.848 V4.0.0.

[10] T. S. Rappaport, *Wireless communications*, Prentic Hall, 1996.

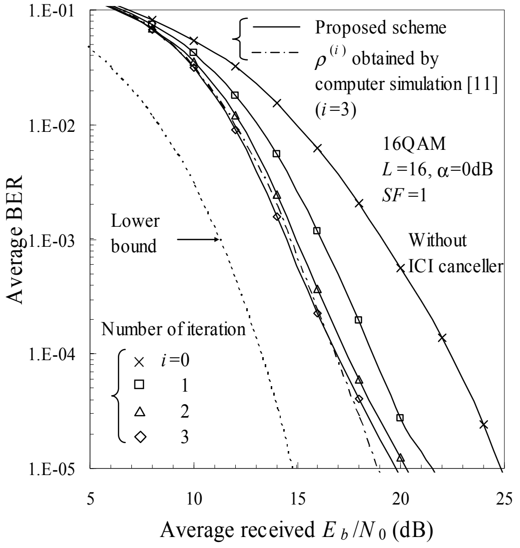
[11] K. Takeda and F. Adachi, “Iterative frequency-domain inter-chip interference cancellation for DS-CDMA,” Proc. 8th WPMC, Aalborg, Denmark, 18-22 Sept. 2005.

[12] A. Stefanov and T. Duman, “Turbo coded modulation for wireless communications with antenna diversity,” Proc. IEEE VTC99-Fall, pp.1565-1569, Netherlands, Sept. 1999.

[13] K. Honjo and T. Ohtuki, “Turbo –BLAST signal detection using QR decomposition in MIMO systems,” (in Japanese) Proc. IEICE General Conference , B-5-100, p.549, Mar.

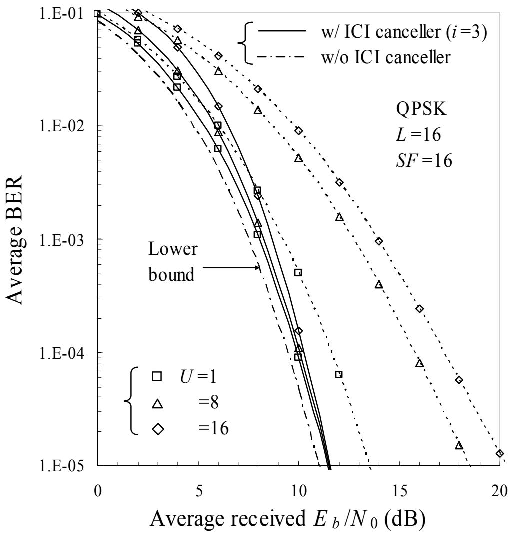


(a) QPSK

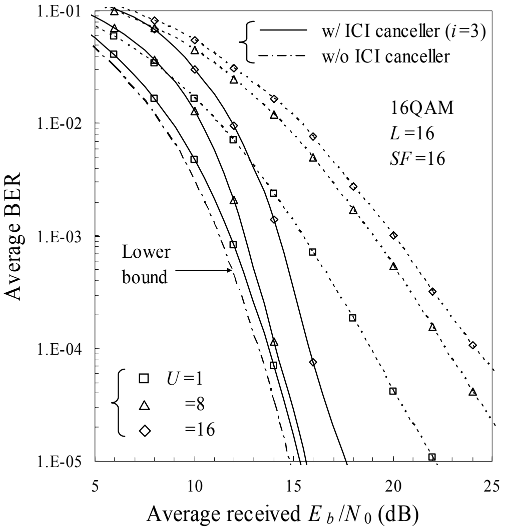


(b) 16QAM

Figure 2. Simulated BER performance with frequency-domain ICI cancellation with MMSE-FDE for *SF*=*U*=1.



(a) QPSK



(b) 16QAM

Figure 3. Simulated BER performance for *SF*=16.