Iterative Overlap FDE for DS-CDMA without GI

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***Abstract*—Frequency-domain equalization (FDE) based on minimum mean square error (MMSE) criterion can replace the conventional rake combining with significantly improved bit error rate (BER) performance for the downlink DS-CDMA in a frequency-selective fading channel. Recently, a new MMSE-FDE technique that requires no GI insertion has been proposed to avoid inter-block interference (IBI) (this technique is called overlap FDE). However, the residual IBI is produced and degrades the BER performance. Furthermore, the presence of an inter-chip interference (ICI) after MMSE-FDE limits the BER performance improvement. In this paper, we propose an iterative overlap FDE for DS-CDMA downlink to suppress both IBI and ICI. In the iterative overlap FDE, joint MMSE-FDE and ICI cancellation is repeated a sufficient number of times and the middle part of *M* chips is picked up. The BER performance with the iterative overlap FDE is evaluated by computer simulation.**

Keyword; DS-CDMA, MMSE-FDE, ICI cancellation, iterative overlap FDE

I. INTRODUCTION

There have been tremendous demands for high-speed data transmissions in mobile communications [1]. A mobile communication channel is composed of many distinct propagation paths having different time delays, resulting in a frequency-selective fading channel [2]. Direct sequence code division multiple access (DS-CDMA) using rake combining is used in the present cellular mobile communication systems for data transmissions of up to around a few Mbps [3,4]. Recently, a lot of research attention is paid to the next generation mobile communication systems that will support transmission data rates higher than several tens of Mbps. The wireless channel for high speed data transmission is severely frequency-selective and the BER performance with the rake combining degrades due to a strong inter-path interference. Hence, an advanced equalization technique is indispensable.

Recently, it has been shown [5-7] that FDE based on minimum mean square error (MMSE) criterion can replace the rake combining and improve the BER performance for the DS-CDMA signal reception over a severe frequency-selective channel. In DS-CDMA with MMSE-FDE, the insertion of the guard interval (GI) at the transmitter is necessary to avoid the presence of inter-block interference (IBI). However, the transmission efficiency is reduced by the insertion of the GI. For the case without the GI, the circularity of the channel matrix is distorted only at the edge of the matrix. The impulse response of MMSE-FDE does not spread over the entire fast Fourier transform (FFT) block. This fact can be exploited to avoid the GI insertion. The presence of IBI can be suppressed by overlapping the FFT block, and picking up the middle *M* chips from FFT block for despreading and data demodulation (this technique is called overlap FDE) [8,9].

However, the residual IBI still remains even if overlap FDE is applied. Furthermore, the presence of a residual inter-chip interference (ICI) after FDE distorts the orthogonality among the spreading codes. The residual IBI and ICI degrade the downlink BER performance as the code multiplexing order increases. The frequency-domain interference cancellation for DS-CDMA uplink has been proposed in [10]. Recently, we have proposed a joint MMSE-FDE and frequency-domain ICI cancellation to improve the BER performance of the DS-CDMA downlink signal transmission [11]. However, in [10], [11], DS-CDMA with the GI is considered. To the best of author's knowledge, the joint effect of the ICI cancellation and overlap FDE has not been examined.

In this paper, we propose an iterative overlap FDE for downlink DS-CDMA without the GI. In the iterative overlap FDE, joint MMSE-FDE and ICI cancellation is repeated a sufficient number of times and the middle part of *M* chips is picked up. If the ICI is sufficiently suppressed, the MMSE weight approaches the maximum ratio combining (MRC) weight. The impulse response of MRC-FDE is concentrated in close vicinity to the edge of FFT block, hence the residual IBI can be sufficiently suppressed by the iterative overlap FDE using ICI cancellation.

II. OVERLAP FDE AND ICI CANCELLATION

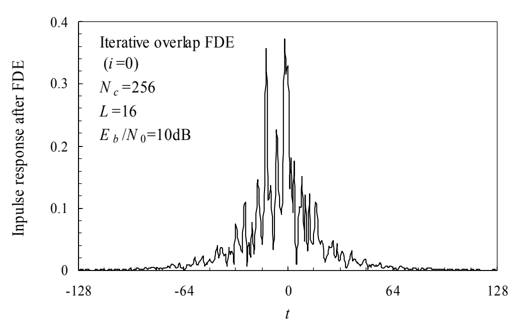
In this section, we investigate the impulse response of the MMSE-FDE filter to explain why overlap FDE can work successfully without requiring the GI insertion. The simulated achievable BER performance with overlap FDE is shown to introduce an iterative overlap FDE.

We assume QPSK data modulation, an FFT block size of N_c=256 chips. The channel is assumed to be a frequency-selective block Rayleigh fading channel having a chip-spaced L=16-path uniform power delay profile. Perfect chip timing and ideal channel estimation are assumed.

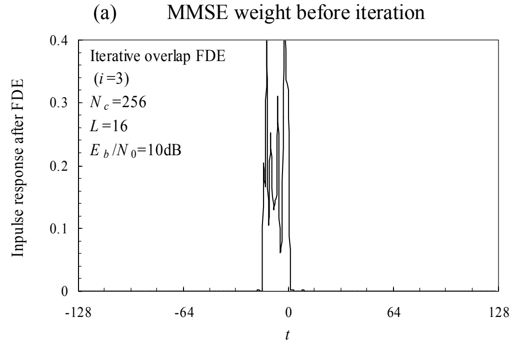
One shot observation of the MMSE-FDE impulse response is shown in Fig. 1 (a). The impulse response does not spread over the entire FFT block. Therefore, the residual IBI after MMSE-FDE does not spread over the entire FFT block if the maximum delay of the multipath τ_{L-1} is sufficiently short compared with the FFT block size N_c. This is exploited by overlap FDE. Overlap FDE does not require the GI [9]. The FFT blocks of N_c chips are overlapped, and the middle part of M ($M \le N_c$) chips after MMSE-FDE is picked up to avoid the residual IBI (Fig. 2).

The simulated BER performance with overlap FDE is plotted for various values of M in Fig. 3 as a function of the average received bit energy-to-AWGN noise power spectrum density ratio E_b/N_0, defined as $E_b/N_0=0.5SF(E_c/N_0)$, where SF is the spreading factor and E_c/N_0 is the average chip energy-to-AWGN power spectrum density ratio. SF and the number U of users are assumed to be $SF=U=16$. For comparison, the theoretical lower bound [6] and the BER performance of DS-CDMA with a GI of $N_g=32$ are also plotted (the GI insertion loss in E_b/N_0 is taken into account). The BER performance of overlap FDE is close to that with the GI, but a BER floor still exists due to the residual IBI. There is a big performance gap between the theoretical lower bound and the BER performance of overlap FDE. This is due to the residual ICI. Even with the GI (i.e., no IBI is present), the performance gap is as much as 7.7 dB for BER=10^{-4}. This indicates that the use of ICI cancellation is effective to improve the BER performance irrespective of the insertion of the GI.

In this paper, we propose an iterative overlap FDE to sufficiently suppress both the residual ICI and IBI. In the iterative overlap FDE, after joint MMSE-FDE and ICI cancellation is repeated a sufficient number of times, the middle part of M chips is picked up. As the number of iteration increases, the FDE weight approaches the maximum ratio combining (MRC) weight since the residual ICI can be better suppressed. Fig. 1 (b) shows one shot observation of the impulse response for the MMSE weight after 3 iterations, which is seen to be much close to the MRC weight. The impulse response concentrates in close vicinity to $t=0$ compared to the original MMSE weight before iteration. (see Fig. 1 (a) and (b)). This suggests that the IBI can also significantly be reduced by the iterative overlap FDE compared to the original overlap FDE, thereby reducing the error floor caused by the residual IBI as shown in Fig. 3.



(a) MMSE weight before iteration



(b) MMSE weight after 3 iterations

Figure 1. One shot obsevation of the impulse response of the FDE weight.

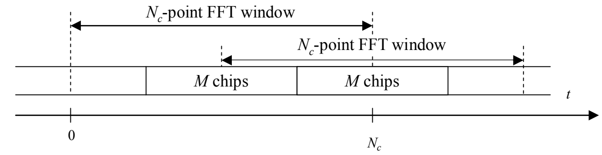


Figure 2. Received signal sequence and FFT window.

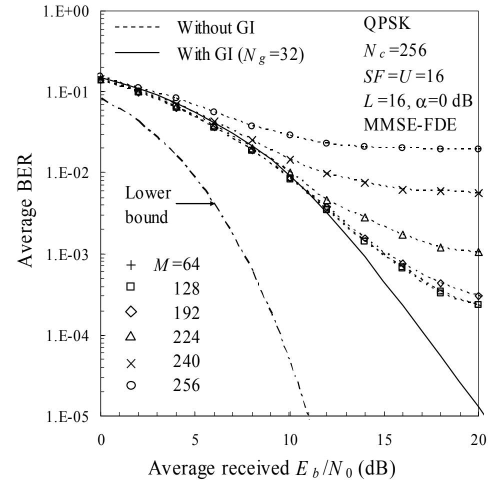
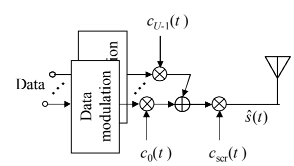
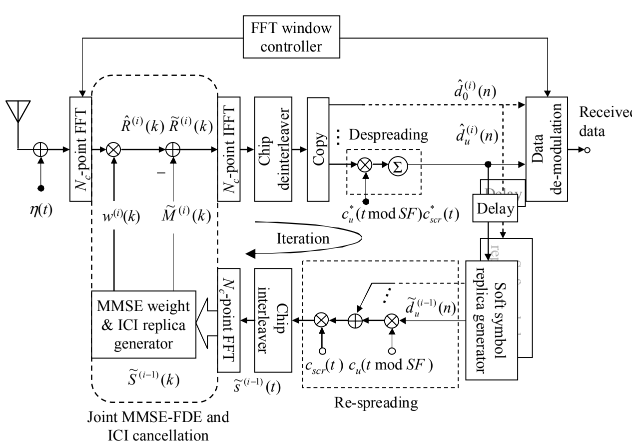


Figure 3. BER performance with overlap FDE.



(a) Transmitter



(b) Receiver

Figure 4. Transmitter/receiver strucutre.

III. TRANSMISSION SYSTEM MODEL

A. Overall transmission system model

The transmission system model for DS-CDMA with the iterative overlap FDE is illustrated in Fig. 4. 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 $c_{scr}(t)$.

The orthogonal multicode DS-CDMA signal is transmitted over a frequency-selective fading channel and is received at a receiver. The received chip sequence is decomposed by N_c-point 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. Joint MMSE-FDE and ICI cancellation is repeated a sufficient number of times. Finally, the middle M chips of N_c chips are despread to recover the received data.

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 users' data symbol sequences $\{d_u(n); n=..,-1,0,1,..\}$ is considered. The spread signal chip sequence $\{\hat{s}(t); t=..,-1,0,1,..\}$ to be transmitted can be expressed, using the equivalent lowpass representation, as

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

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 $|c_u(t)|=|c_{scr}(t)|=1$, 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 an independent fading. The assumption of block fading means that the path gains remain constant over at least one FFT block duration. The impulse response $h(t)$ of a multipath channel can be expressed as [12]

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

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

$$r(t)=\sqrt{2E_c/T_c}\sum_{l=0}^{L-1} h_l\,s(t-\tau_l)+\eta(t) \quad ,(4)$$

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.

Without loss of generality, we assume that N_c-point FFT block is applied to the received signal block of t=0~N_c–1. Eq. (4) can be rewritten as

$$r(t)=\sqrt{2E_c/T_c}\sum_{l=0}^{L-1} h_l\,s((t-\tau_l) \bmod N_c)+\nu(t)+\eta(t) \,,(5)$$

where $\nu(t)$ is the IBI component and $\nu(t)=0$ for $t>\tau_{L-1}$.

C. Joint MMSE-FDE and ICI cancellation

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

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 $R(k)$ can be written as

$$\begin{aligned}R(k)&=\sum_{t=0}^{N_c-1} r(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ &=H(k)S(k)+\mathrm{N}(k)+\Pi(k)\end{aligned} \quad ,(6)$$

where $S(k)$, $H(k)$, $\mathrm{N}(k)$ 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, the IBI component and the noise component due to the AWGN, respectively.

MMSE-FDE is carried out as follows:

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

where $w^{(i)}(k)$ is the equalization weight at the ith iteration and $\hat{H}^{(i)}(k)$, $\hat{\mathrm{N}}^{(i)}(k)$ and $\hat{\Pi}^{(i)}(k)$ are the equivalent channel gain, the IBI component and the noise component, after performing MMSE-FDE at the ith iteration, respectively. The MMSE weight is given by

$$w^{(i)}(k)=\frac{H^*(k)}{|H(k)|^2\rho^{(i-1)}+2\sigma^2} \quad ,(8)$$

where $\rho^{(i-1)}$ is an interference factor and $\rho^{(-1)}=1$, which is described 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]

$$\tilde{M}^{(i)}(k)=\begin{cases}0 & \text{for } i=0\\ \left\{\hat{H}^{(i)}(k)-A^{(i)}\right\}\tilde{S}^{(i-1)}(k) & \text{for } i\geq 1\end{cases} \,,(10)$$

where $\tilde{S}^{(i-1)}(k)$ is the kth frequency component of the transmitted chip replica generated by feeding back a decision variable of the (i–1)th iteration stage and $A^{(i)}$ is given by

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

D. Despreading

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

$$\tilde{r}^{(i)}(t) = \frac{1}{N_c}\sum_{k=0}^{N_c-1}\tilde{R}^{(i)}(k)\exp\left(j2\pi t\frac{k}{N_c}\right) = A^{(i)}s(t) + \mu^{(i)}(t) + \hat{v}^{(i)}(t) + \hat{\eta}^{(i)}(t), \quad (12)$$

where $s(t)$ in the first term represents the transmitted chip, $\mu^{(i)}(t)$ is the residual ICI, $\hat{v}^{(i)}(t)$ is the residual IBI and $\hat{\eta}^{(i)}(t)$ is the noise component. As understood from Fig. 3, the residual ICI is the predominant cause of the performance degradation.

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

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

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

IV. ICI Replica Generation and MMSE Weight

In this section, we consider the $i(\geq 1)$th iteration. ICI replica generation for the ith iteration is presented. Then, MMSE weight taking into account the residual ICI is derived and the residual IBI after ICI cancellation is described.

A. ICI replica generation

The soft decision variable is used to generate the replica $\{\tilde{s}^{(i-1)}(t); t = 0 \sim N_c - 1\}$ of the transmitted chip sequence so as to avoid the error propagation due to the erroneous decision variable.

Using the uth user's soft decision variable $\hat{d}_u^{(i-1)}(n)$, the log-likelihood ratio (LLR) for the xth bit in the nth symbol $d_u(n)$ (n=0~N_c/SF−1), where x=0~$\log_2$M−1 and M is the modulation level, can be computed as

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

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

For QPSK data modulation, the soft symbol replica $\{\tilde{d}_u^{(i-1)}(n); n$=0~$N_c/SF$−1$\}$ can be obtained from [11]

$$\tilde{d}_u^{(i-1)}(n) = \frac{1}{\sqrt{2}}\tanh\left(\frac{L_0(n)}{2}\right) + j\frac{1}{\sqrt{2}}\tanh\left(\frac{L_1(n)}{2}\right). \quad (15)$$

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

$$\tilde{s}^{(i-1)}(t) = \left[\sum_{u=0}^{U-1}\tilde{d}_u^{(i-1)}(\lfloor t/SF \rfloor)c_u(t \bmod SF)\right]c_{scr}(t). \quad (16)$$

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

$$\tilde{S}^{(i-1)}(k) = \sum_{t=0}^{N_c-1}\tilde{s}^{(i-1)}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right). \quad (17)$$

Substituting Eq. (17) into Eq. (10), we obtain the frequency-domain ICI replica $\tilde{M}^{(i)}(k)$.

B. MMSE weight and interference factor

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

$$e(k) = \tilde{R}^{(i)}(k) - A^{(i)}S(k), \quad (18)$$

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

$$w^{(i)}(k) = \frac{H^*(k)}{|H(k)|^2\rho^{(i-1)} + 2\sigma^2}, \quad (19)$$

where $\rho^{(i-1)}$ is an interference factor given by

$$\rho^{(i-1)} = \sum_{t=0}^{N_c-1}\left\{|s(t)|^2 - |\tilde{s}^{(i-1)}(t)|^2\right\}. \quad (20)$$

Since $s(t)$ is unknown, we use the hard decision chip sequence replica $\bar{s}^{(i-1)}(t)$ instead of $s(t)$ in Eq. (20).

C. Residual IBI after ICI cancellation

Assuming that the residual ICI is perfectly cancelled (i.e., $\tilde{S}^{(i)}(k) = S(k)$ and $\tilde{s}^{(i)}(t) = s(t)$), the MMSE weight of Eq. (19) approaches the MRC weight since $\rho^{(i)} = 0$ from Eq. (20). The MRC weight is given by

$$w^{(i)}(k) = H^*(k). \quad (21)$$

The frequency-domain signal $\{\tilde{R}^{(i)}(k); k = 0 \sim N_c - 1\}$ after MRC-FDE and perfect ICI cancellation is transformed into time-domain chip sequence $\{\tilde{r}^{(i)}(t); t = 0 \sim N_c - 1\}$ by applying N_c-point IFFT:

$$\tilde{r}^{(i)}(t) = A^{(i)}s(t) + \hat{v}^{(i)}(t) + \tilde{\eta}(t) \quad ,(22)$$

where $\{\hat{v}^{(i)}(t); t = 0 \sim (N_c - 1)\}$ is the residual IBI component given by

$$\hat{v}^{(i)}(t) = \begin{cases} 0 & \tau_{L-1} < t < N_c - \tau_{L-1} \\ \sum_{l=0}^{L-1} h_l^* v(t + \tau_l \bmod N_c) & \text{otherwise} \end{cases} \quad .(23)$$

From Eq. (23), if the residual ICI is perfectly cancelled, the residual IBI is not present in the region of $\tau_{L-1} < t < N_c - \tau_{L-1}$. Hence, the residual IBI can be avoided when the received chip sequence is picked up from $\tau_{L-1} < t < N_c - \tau_{L-1}$.

V. COMPUTER SIMULATION

The simulation parameters are the same as in Sect. II. The simulated BER performance of the iterative overlap FDE is plotted in Fig. 5 with the number *i* of iterations as a parameter for *SF*=*U*=1 and 16, and *M*=160. For comparison, the theoretical lower bound is also plotted. The BER performance of *i*=0 corresponds to the case with the original overlap FDE. The iterative overlap FDE using ICI cancellation is very effective to improve the BER performance even without the GI. On the other hand, the BER floor is seen for overlap FDE (*i*=0) due to the residual IBI. A good BER performance is achieved for *i*=3, and the required E_b/N_0 reduction is as much as 5.1 dB for BER=10^{-3}. The E_b/N_0 degradation from the theoretical lower bound is only 1.2 dB. When *SF*=*U*=16, the BER performance is greatly improved by iterative overlap FDE, similar to the case of *SF*=1. An E_b/N_0 degradation of as small as 0.9 dB from the theoretical lower bound is confirmed. Note that the computational complexity of the iterative overlap FDE increases because of N_c/M times increase in the number of FFT/IFFT operations over the conventional FDE and because of the iterative ICI cancellation operation.

VI. CONCLUSION

In this paper, iterative overlap FDE was proposed. Although overlap FDE can improve the BER performance of DS-CDMA without the GI, the BER floor exists due to the residual IBI. The residual ICI is also present after MMSE-FDE, which is the predominant cause of the performance degradation from the theoretical lower bound. In the iterative overlap FDE, after joint MMSE-FDE and ICI cancellation is repeated a sufficient number of times, the middle part of *M* chips is picked up to reduce the effect of IBI and ICI. The BER performance with iterative overlap FDE was evaluated by computer simulation. It was found that, when *SF*=*U*=1 (16), the E_b/N_0 reduction from the performance with *i*=0 is as much as about 5.1 (6.4) dB for achieving BER=10^{-3}. The E_b/N_0 degradation from the theoretical lower bound is only 1.2 (0.9) dB for *SF*=*U*=1 (16).

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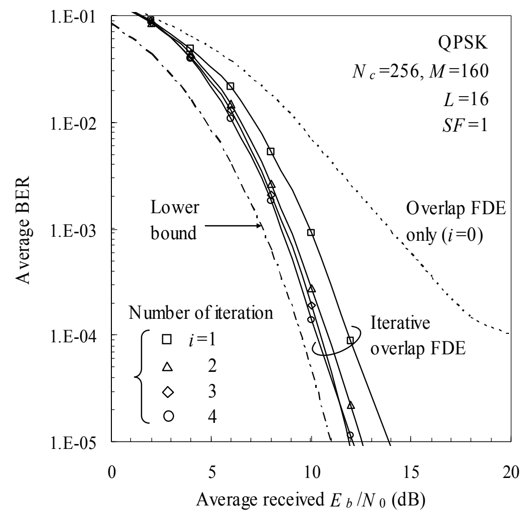
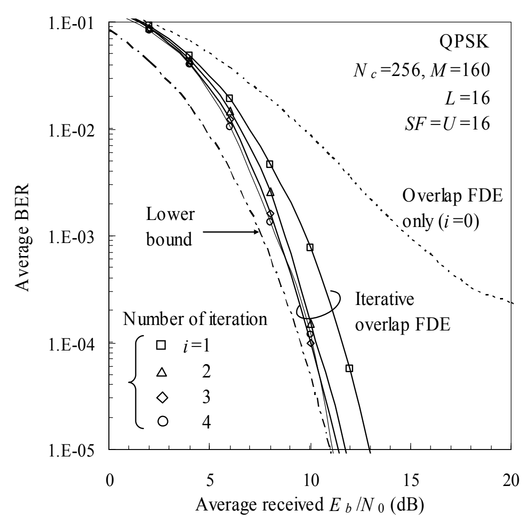


(a) *SF*=1



(b) *SF*=*U*=16

Figure 5. BER performance of iterative overlap FDE.