Pilot-assisted Channel Estimation for Overlap FDE of DS-CDMA Signals

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Abstract— **Overlap frequency-domain equalization (FDE) requires no guard interval (GI) insertion while achieving the frequency diversity gain and hence can be applied to the present cellular systems using direct sequence-code division multiple access (DS-CDMA). It has been confirmed that assuming the ideal channel estimation, overlap FDE can provide almost the same bit error rate (BER) performance as the conventional FDE using GI insertion. In this paper, we propose a pilot-assisted channel estimation suitable for overlap FDE, in which a short pilot chip sequence is repeated twice in the chip block so that the received pilot sequence becomes a circular convolution between the transmitted pilot sequence and channel impulse response. We evaluate by computer simulation the BER performance achievable by the proposed pilot-assisted channel estimation in a frequency-selective Rayleigh fading channel.**

Keywords— **Overlap FDE, pilot-assisted channel estimation, DS-CDMA**

I. INTRODUCTION

For the next generation mobile communication systems, high speed packet data transmission services are demanded. In the third generation mobile communication systems, direct sequence-code division multiple access (DS-CDMA) with coherent rake combining is used to obtain the path diversity gain (or frequency diversity gain) [1]. However, the broadband wireless channel is composed of many distinct paths having different time delays [2] and therefore, the transmission performance of DS-CDMA with coherent rake combining significantly degrades due to the strong inter-chip interference (ICI) arising from severe frequency-selective fading channel.

Multi-carrier (MC) transmissions [3-5] (e.g., orthogonal frequency division multiplexing (OFDM) and MC-CDMA) have been attracting attention since the use of frequency-domain equalization (FDE) can take advantage of channel frequency-selectivity to improve the bit error rate (BER) performance. Recently, single-carrier (SC) transmission including DS-CDMA is also attracting much attention [6, 7]. Replacing the coherent rake combining by FDE based on minimum mean square error (MMSE) criterion can significantly improve the DS-CDMA transmission performance. As a consequence, both MC- and DS-CDMA techniques are considered to be potential candidates for the next generation wireless access.

However, the use of FDE requires the insertion of cyclic prefix into the guard interval (GI) so that the received signal becomes a circular convolution between the transmitted signal and the channel impulse response. However, the GI insertion reduces the transmission rate for the given bandwidth and also prevents the application of FDE to the present cellular systems using DS-CDMA.

Recently, we have proposed overlap FDE that requires no GI insertion while achieving the frequency diversity gain for DS-CDMA [8, 9]. In our previous work, it has been confirmed that overlap FDE can provide almost the same BER performance as the conventional FDE using GI insertion. However, we have assumed the perfect channel estimation.

In this paper, we propose a pilot-assisted channel estimation scheme suitable for overlap FDE and evaluate the achievable BER performance of DS-CDMA by computer simulation. The reminder of this paper is organized as follows. Sect. II introduces the overlap FDE. In Sect. III, the proposed channel estimation scheme is described. Sect. IV presents the simulation results. The paper is concluded in Sec. V.

II. OVERLAP FDE [8]

The overlap FDE requires no modification to the transmitter structure of conventional DS-CDMA systems. Multi-code DS-CDMA using code-multiplexing order U is considered. In this paper, sample-spaced discrete-time, equivalent lowpass representation is used.

A. Transmit signal

At the transmitter, the data modulated symbol sequence is serial-to-parallel (S/P) converted into U parallel streams $\{d_u(i); i=\ldots,-1,0,1,\ldots\}$, $u=0\sim U-1$. Then, each stream is spread by using an orthogonal spreading code with spreading factor SF $\{c_u(t); t=0\sim SF-1\}$, $u=0\sim U-1$. After code-multiplexing the U chip sequences, the multi-code chip sequence is multiplied by a scramble sequence $\{c_{scr}(t); t=\ldots,-1,0,1,\ldots\}$ to obtain

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

where $\lfloor x \rfloor$ represents the largest integer smaller than or equal to x.

B. Received signal

The transmitted multi-code DS-CDMA signal is received via a frequency-selective fading channel. We assume an L-path frequency-selective block Rayleigh fading channel having the impulse response as

$$h(\tau)=\sum_{l=0}^{L-1} h_l \delta(\tau-\tau_l), \quad (2)$$

where h_l and τ_l respectively denote the complex valued path gain with $\sum_{l=0}^{L-1} E[|h_l|^2]=1$ and delay time of the lth path.

The receiver structure is illustrated in Fig. 1. Without loss of generality, the received signal $\{r(t); t=0\sim N_c-1\}$ is considered, where N_c is the fast Fourier transform (FFT) block size. $r(t)$ can be expressed as

$$\begin{aligned} r(t) &= \sqrt{\frac{2E_c}{T_c}}\sum_{l=0}^{L-1} h_l s(t-\tau_l) + \eta(t) \\ &= \sqrt{\frac{2E_c}{T_c}}\sum_{l=0}^{L-1} h_l s((t-\tau_l) \bmod N_c) + \nu(t) + \eta(t) \end{aligned}, \quad (3)$$

where E_c and T_c represent the chip energy per multi-code stream and the chip duration, respectively, $\eta(t)$ represents the noise due to additive white Gaussian noise (AWGN) having the one-sided power spectrum density N_0, and $\nu(t)$ represents the inter-block interference (IBI). In conventional FDE, a cyclic prefix is inserted into the GI placed at the beginning of the FFT block to avoid the IBI, i.e., $\nu(t)=0$. However, in overlap FDE, since the GI is not inserted and hence, the IBI is present at the beginning of the received N_c-chip block. $\nu(t)$ is expressed as

$$\begin{aligned} \nu(t) = \sqrt{\frac{2E_c}{T_c}}\sum_{l=0}^{L-1} h_l \{s(t-\tau_l) - s((t-\tau_l) \bmod N_c)\} \\ \times \{u_0(t) - u_0(t-\tau_l)\} \end{aligned}, \quad (4)$$

with $u_0(t)=0$ (1) for $t<0$ (otherwise).

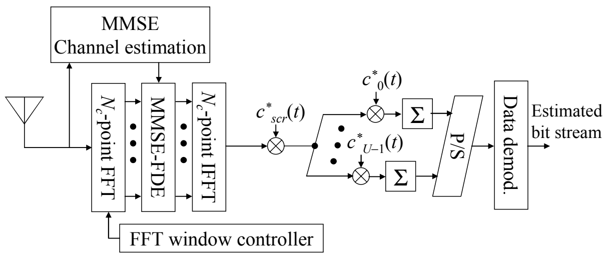

Fig. 1. Receiver structure.

C. Conventional FDE

The received signal block $\{r(t);\ t=0\sim N_c-1\}$ is transformed into the frequency-domain signal $\{R(k);\ k=0\sim N_c-1\}$ by the use of N_c-point FFT as

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

where $H(k)$, $S(k)$, $N(k)$, and $\Pi(k)$ are respectively the channel gain, frequency-domain signal, IBI component, and the noise at the kth frequency, given by

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

$\{R(k);\ k=0\sim N_c-1\}$ is multiplied by the MMSE-FDE weight and then, transformed back to the time-domain signal by the use of N_c-point inverse FFT (IFFT). By approximating the IBI component as a Gaussian variable, the MMSE-FDE weight $w(k)$ can be given by [8]

$$w(k) = \frac{\tilde{H}^*(k)}{U\left|\tilde{H}(k)\right|^2 + 2\tilde{\sigma}^2}, \quad (7)$$

where $\tilde{H}(k)$ is the estimated channel gain and $2\tilde{\sigma}^2$ is the estimate of the IBI plus noise.

After MMSE-FDE, the time-domain received signal $\{\hat{r}(t);\ t=0\sim N_c-1\}$ is obtained by N_c-point IFFT. $\hat{r}(t)$ is given by

$$\begin{aligned} \hat{r}(t) &= \frac{1}{N_c}\sum_{k=0}^{N_c-1} w(k)R(k)\exp\left(j2\pi k\frac{t}{N_c}\right) \\ &= \left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}(k)\right)s(t) + \mu(t) + \hat{\nu}(t) + \hat{\eta}(t) \end{aligned}, \quad (8)$$

where $\hat{H}(k) = w(k)H(k)$ is the equivalent channel gain after MMSE-FDE. The first, second, third, and fourth terms of Eq. (8) represent the desired signal, residual ICI, residual IBI, and noise, respectively. They are given as

$$\begin{cases} \mu(t) = \frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}(k)\left[\sum_{\substack{\tau=0 \\ \neq t}}^{N_c-1} s(\tau)\exp\left(j2\pi k\frac{t-\tau}{N_c}\right)\right] \\ \hat{\nu}(t) = \frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{N}(k)\exp\left(j2\pi k\frac{t}{N_c}\right) \\ \hat{\eta}(t) = \frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{\Pi}(k)\exp\left(j2\pi k\frac{t}{N_c}\right) \end{cases}. \quad (9)$$

D. Overlap FDE

The residual IBI after MMSE-FDE is a circular convolution of the IBI and the circular impulse response of MMSE-FDE filter. Since the MMSE-FDE filter impulse response concentrates at a vicinity of $\tau=0$ [8], the residual IBI is localized only near the both ends of N_c-chip FFT block. In overlap FDE, the received sequence is divided into a sequence of M-chip blocks ($M<N_c$). Then, N_c-point FFT is applied to an N_c-chip interval centering an M-chip block of interest. After MMSE-FDE, M-chip block $\{\hat{r}(t);\ t=N_c/2-M/2\sim N_c/2+M/2-1\}$ is picked up from the equalized N_c-chip block $\{\hat{r}(t);\ t=0\sim N_c-1\}$ to reduce the residual IBI. Therefore, the FFT intervals for consecutive M-chip blocks are overlapped as shown in Fig. 2. After obtaining a sequence of M-chip blocks, de-spreading is done to obtain a decision variable for $d_u(i)$ as

$$\hat{d}_u(i) = \frac{1}{SF}\sum_{t=iSF}^{(i+1)SF-1}\hat{r}(t)c^*_{scr}(t)c^*_u(t \bmod SF). \quad (10)$$

As M reduces, the residual IBI can be better suppressed, however, the number of FFT/IFFT operations increases N_c/M times. Therefore, M should be as large as possible in order not to increase the computational complexity excessively while sufficiently suppressing the residual IBI.

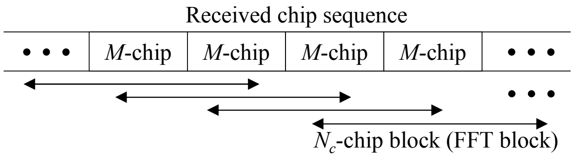

Fig. 2. Overlap FDE.

III. Channel Estimation

We use the time-multiplexed pilot assisted channel estimation (PACE). In Ref. [10], a frequency-domain PACE using the GI-inserted pilot chip block $p(t \bmod N_c)$, $t=-N_g\sim N_c-1$, is presented. The received pilot block is transformed into the frequency-domain pilot and then, the pilot modulation is removed by multiplying the frequency-domain MMSE weight to obtain the channel gain estimates. It was shown [10] that when the GI is inserted, this frequency-domain MMSE channel estimation gives accurate channel estimates. However, without the GI insertion, the accuracy of channel estimation significantly degrades due to the severe IBI. To overcome this problem, in this paper, a short pilot chip sequence $p(t)$, $t=0\sim N_c-1$, having a period of $N_c/2$ with $|p(t)|=1$ is repeated twice in the pilot block. The pilot block can be expressed as $p(t \bmod N_c/2)$, $t=0\sim N_c-1$. The first half of the pilot chip block, $p(t)$, $t=0\sim N_c/2-1$, plays a role as the cyclic prefix for the latter half of the pilot chip block $p(t)$, $t=N_c/2\sim N_c-1$. The latter half of the received pilot block is free from the IBI if the maximum delay time τ_{max} of the channel is shorter than $N_c/2$ chips.

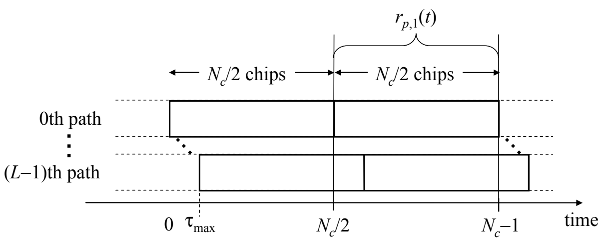

Fig. 3. FFT block for channel estimation using the latter half of the received pilot block.

A. Instantaneous channel gain estimate

The received pilot block can be expressed as

$$r_p(t)=\sqrt{2S_p}\sum_{l=0}^{L-1}h_l\,p(t \bmod N_c/2)+\eta(t)\,, \tag{11}$$

where S_p $(=UE_c/T_c)$ denotes the received pilot power and $\eta(t)$ is the zero-mean AWGN. The second half $\{r_p(t);\ t=N_c/2\sim N_c-1\}$ of the received pilot block is used for the channel estimation. $N_c/2$-point FFT is applied to $\{r_p(t);\ t=N_c/2\sim N_c-1\}$ to obtain the frequency-domain pilot as

$$\begin{aligned}R_p(q)&=\sum_{t=0}^{N_c/2-1}r_p(t-N_c/2)\exp\left(-j2\pi q\frac{t}{(N_c/2)}\right)\\&=\sqrt{U}H(2q)P(q)+\Pi(2q)\end{aligned}\ , \tag{12}$$

where $H(2q)$ and $\Pi(2q)$ are the channel gain and noise of Eq. (6) at even frequency $k=2q$, respectively, and $P(q)$ is given by

$$P(q)=\sum_{t=0}^{N_c/2-1}p(t)\exp\left(-j2\pi q\frac{t}{(N_c/2)}\right). \tag{13}$$

The channel gain $H(2q)$ at even frequency can be estimated using the latter half of the received pilot block and its estimate is denoted by $\tilde{H}(2q)$, $q=0\sim N_c/2-1$. The zero-forcing (ZF) technique can be used (i.e., $R_p(q)$ is divided by $P(q)$). However, if a pseudo-noise (PN) sequence is used as the pilot $p(t)$, the ZF channel estimation produces the noise enhancement since the amplitude spectrum $P(q)$ is not constant. To suppress the noise enhancement, we apply the MMSE channel estimation [10]. The MMSE channel estimation minimizes the mean square error between $\tilde{H}(2q)$ and $H(2q)$, which is given by

$$\tilde{H}(2q)=R_p(q)X(q)\,, \tag{14}$$

with

$$X(q)=\frac{P^*(q)}{|P(q)|^2+(S_p/\hat{\sigma}^2)^{-1}}\,. \tag{15}$$

where $\hat{\sigma}^2$ is the noise power estimate and can be obtained according to [10].

B. Interpolation

The instantaneous channel gain edstimates are noisy. The delay time-domain windowing technique can be applied to reduce the noise power [10, 11]. Since only the channel gains at even frequencies can be estimated using the pilot, the interpolation technique is necessary to estimate the channel gain at the odd frequencies. The delay time-domain windowing technique can perform interpolation while reducing the noise power.

At first, $N_c/2$-point IFFT is applied to $\{\tilde{H}(2q);\ q=0\sim N_c/2-1\}$ to obtain the estimated channel impulse response $\{\tilde{h}(\tau);\ \tau=0\sim N_c/2-1\}$ as

$$\tilde{h}(\tau)=\frac{1}{N_c}\sum_{q=0}^{N_c/2-1}\tilde{H}(2q)\exp\left(j2\pi q\frac{\tau}{(N_c/2)}\right) \tag{16}$$

Assuming that the real channel impulse response $h(\tau)$ is confined within a delay time interval of $\tau=0\sim\tau_{max}-1$, $\tilde{h}(\tau)$ is replaced by zero for $\tau=\tau_{max}\sim N_c/2-1$. Then, applying N_c-point FFT to the estimated channel impulse response, the noise-reduced and interpolated channel gain estimates $\{\tilde{H}(k);\ k=0\sim N_c-1\}$ is obtained as

$$\tilde{H}(k) = \sum_{\tau=0}^{\tau_{max}-1} \tilde{h}(\tau)\exp\left(-j2\pi k\frac{\tau}{N_c}\right) = \sum_{q=0}^{N_c/2-1} A(k-2q)\tilde{H}(2q) \quad , \tag{17}$$

where $A(n)$ is the interpolation co-efficient given by

$$A(n) = \frac{1}{N_c}\exp\left(j\pi(\tau_{max}-1)\frac{2q-k}{N_c}\right)\frac{\sin\left(\pi\tau_{max}\frac{2q-k}{N_c}\right)}{\sin\left(\pi\frac{2q-k}{N_c}\right)} \quad . \tag{18}$$

C. Improved channel estimation

If the receiver knows the value of τ_{max}, the channel estimation can be carried out without causing IBI using the fist half $\{r_p(t);\ t=\tau_{max}\sim N_c/2-1+\tau_{max}\}$ of the received pilot block. The frequency-domain pilot signal obtained by $N_c/2$-point FFT is denoted by $R_0(q)$ and is given by

$$R_{p,0}(q) = \sum_{t=0}^{N_c/2-1} r_{p,0}(t)\exp\left(-j2\pi q\frac{t}{(N_c/2)}\right) = \sqrt{U}H(2q)P(q)\exp\left(j2\pi q\frac{\tau_{max}}{(N_c/2)}\right) + \Pi_0(2q) \quad , \tag{19}$$

where $\Pi_0(q)$ is the noise. The MMSE channel estimation is done using $\{R_{p,0}(q);\ q=0\sim N_c/2-1\}$, similar to Eqs. (12)~(15), to obtain the instantaneous channel gain estimates $\{\tilde{H}_0(2q);\ q=0\sim N_c/2-1\}$.

Denoting the instantaneous channel gain estimate using the received pilot block $\{r_p(t);\ t=N_c/2\sim N_c-1\}$ by $\{\tilde{H}_1(2q);\ q=0\sim N_c/2-1\}$, the improved instantaneous channel gain estimate is obtained by averaging two instantaneous channel gain estimates as

$$\tilde{H}(2q) = \frac{1}{2}\left\{\tilde{H}_0(2q)\exp\left(-j2\pi q\frac{\tau_{max}}{(N_c/2)}\right) + \tilde{H}_1(2q)\right\} . \tag{20}$$

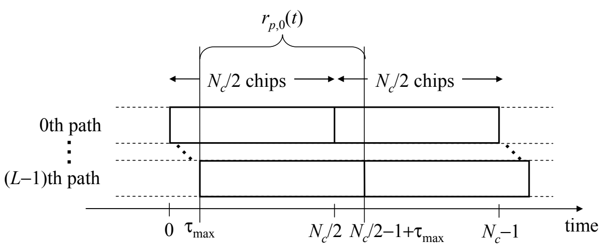

Fig. 4. FFT block for channel estimation using the first half of the received pilot block.

IV. Simulation Results

The achievable average BER performance with overlap FDE using the proposed PACE is evaluated by computer simulation. The spreading factor SF=16 and the full code-multiplexing U=SF is considered. Pilot chip sequence of $N_c/2$=128 chips is taken from a PN sequence of 4095-chip length. Pilot modulation is BPSK while data modulation is QPSK. A chip-spaced L=16-path frequency-selective block Rayleigh fading channel having a uniform power delay profile is assumed. Ideal estimation of the channel maximum time delay is also assumed. For overlap FDE, the number M of chips to be pick up from N_c=256-chip block is set as M=128 when $\tau_l=l$ and as M=64 when $\tau_l=2l$, $l=0\sim L-1$.

Table 1. Simulation condition

Transmitter	Pilot modulation		BPSK
	Pilot sequence		PN sequence
	Data modulation		QPSK
	Spreading factor		SF=16
	No. of mux. codes		U=SF (Full code-mux.)
Channel	Frequency-selective block Rayleigh		
	No. of paths		L=16
	Power delay profile		Uniform
	Path time delay		$\tau_l=l\Delta$, Δ=1, 2
	Normalized Doppler frequency		$f_D N_c T_c$=0
Receiver	Overlap FDE	FFT/IFFT block size	N_c=256
		No. of chips to be picked up	M=64, 128
	Channel estimation	Maximum time-delay	known

The BER performance with overlap FDE using the proposed PACE is plotted in Fig. 5 as a function of the average received E_b/N_0 (including a pilot insertion loss of 0.28dB). The two channel estimation cases are considered: the first and second half of the received pilots are used for channel estimation (w/ averaging) and only the second half of the received pilot is used (w/o averaging). For comparison, the BER performance is also plotted for conventional PACE using N_g=16-chip GI-inserted pilot (PACE w/ GI) [10] and that using no GI-inserted pilot (PACE w/o GI). It can be seen from Fig. 5 that the proposed CE w/averaging can provide almost the same BER performance as the PACE w/ GI; the E_b/N_0 degradation from the ideal CE for BER=10^{-2} is as small as 1 dB (out of which 0.28 dB is due to the pilot insertion) from the ideal channel estimation case. On the other hand, the BER performance of PACE w/o GI significantly degrades due to the residual IBI.

Figure 6 plots the BER performance when the channel delay exceeds the GI-length in the conventional PACE w/GI. The conventional PACE suffers from the IBI, resulting in the degraded BER performance. On the other hand, our proposed PACE does not suffer from the IBI as far as $\tau_{max}<N_c/2$. As seen from Fig. 6, our proposed PACE is robust against the channel frequency-selectivity and provides much better BER performance than PACE w/ GI even in a strong frequency-selective fading channel (i.e., $\tau_{max}>N_g$).

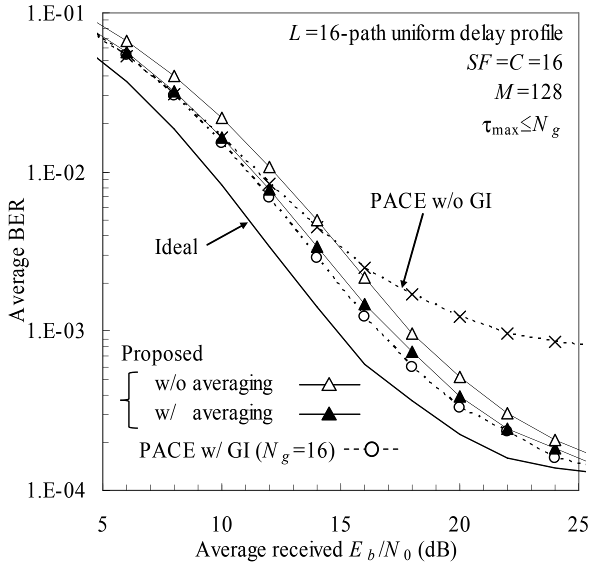

Fig. 5. BER performance when $\tau_{max} \leq N_g$.

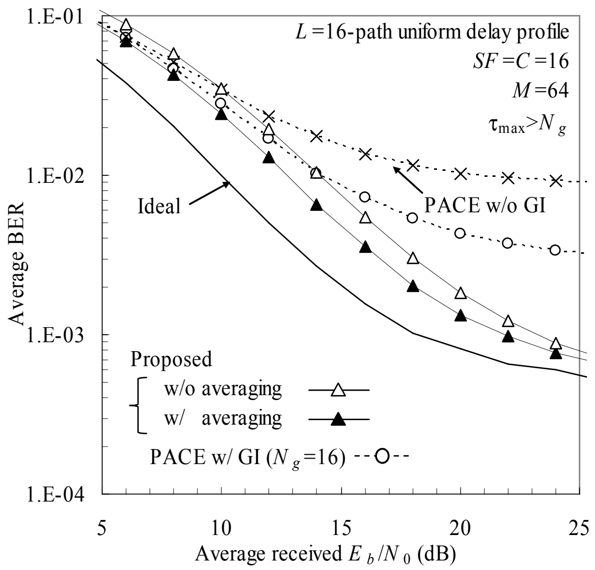

Fig. 6. BER performance when $\tau_{max} > N_g$.

V. Conclusion

In this paper, we proposed a pilot assisted channel estimation (PACE) suitable for DS-CDMA with overlap FDE. A short pilot chip sequence is transmitted twice in the chip block. The computer simulation comfirmed that if $\tau_{max} \leq N_g$ (i.e., the channel maximum time delay is shorter than the GI length), the proposed PACE provides almost the same BER performance as the conventional PACE with GI-inserted pilot. If $\tau_{max} > N_g$, our proposed PACE provides much better BER performance than the conventional PACE with GI-inserted pilot.

References

[1] F. Adachi, M. Sawahashi, and H. Suda, "Wideband DS-CDMA for next-generation mobile communications systems," IEEE Commun. Mag., Vol. 36, No. 9, pp. 56-59, Sep. 1998.

[2] J.G. Proakis, Digital communications, 2nd ed., McGraw-Hill, 1995.

[3] S.Hara and R.Prasad, "Overview of multicarrier CDMA," IEEE Commun. Mag., Vol. 35, No.12, pp.126-144, Dec. 1997.

[4] S. Hara and R. Prasad, "Design and performance of multicarrier CDMA system in a frequency-selective Rayleigh fading channels," IEEE Trans. Veh. Technol., Vol. 48, pp. 1584-1595, Sep. 1999.

[5] L. Hanzo, W. Webb, and T. Keller, Single- and multi-carrier quadrature amplitude modulation, John Wiley & Sons, 2000.

[6] D. Falconer, S. L. Ariyavisitakul, A. Benyamin-Seeyar, and B. Eidson, "Frequency domain equalization for single-carrier broadband wireless systems," IEEE Commun. Mag., Vol. 40, No. 4, pp. 58-66, Apr. 2002.

[7] F. Adachi, D. Garg, S. Takaoka, and K. Takeda, "Broadband CDMA techniques," IEEE Wireless Commun., Vol. 12, No. 2, pp. 8-18, Apr. 2005.

[8] K. Takeda, H. Tomeba, and F. Adachi, "Iterative overlap FDE for DS-CDMA without GI," IEEE VTC '06 fall, Montreal, Quebec, Canada, Sept. 2006.

[9] Kazuki Takeda, H. Tomeba, Kazuaki Takeda and F. Adachi, "Throughput of DS-CDMA HARQ with overlap frequency-domain equalization," 10th IEEE International Conference on Communication Systems (ICCS 2006), Singapore, 30 Oct. -1 Nov. 2006

[10] K. Takeda and F. Adachi, "SNR estimation for pilot-assisted frequency-domain MMSE channel estimation," Proc. IEEE VTS 2nd APWCS, Hokkaido University, Japan, 4-5 Aug. 2005.

[11] J. J. de Beek, O. Edfors, M. Sandell, S. K. Wilson, and P. O. Borjesson, "On channel estimation in OFDM systems," IEEE VTC'95, Vol. 2, pp. 815-819, July 1995.