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HARQ THROUGHPUT OF DS-CDMA OVERLAP FDE USING REPETITION PILOT-ASSISTED CHANNEL ESTIMATION

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ABSTRACT

Direct sequence-code division multiple access (DS-CDMA) with overlap frequency-domain equalization (FDE) achieves a good transmission performance while it requires no guard interval (GI) insertion in a severe frequency-selective fading channel. Overlap FDE requires an accurate channel estimation. Recently, we proposed a repetition pilot-assisted channel estimation based on the minimum mean square error criterion (MMSE-RPACE) for overlap FDE. In this paper, we apply MMSE-RPACE to DS-CDMA hybrid ARQ (HARQ) and evaluate by computer simulation the achievable throughput performance. The simulation results confirm that MMSE-RPACE provides only a slight throughput degradation from the ideal channel estimation and achieves a better throughput performance than the conventional FDE.

I. INTRODUCTION

High-speed wireless packet access is the core technology for the next generation mobile communication systems. Hybrid ARQ (HARQ) is known as one of the promising error control techniques for high-speed packet access [1]-[2]. The broadband wireless channel is composed of many propagation paths having different time delays [3]. The throughput performance of single-carrier packet access transmission significantly degrades due to severe inter-symbol interference (ISI). In the third generation mobile communication systems, direct sequence-code division multiple access (DS-CDMA) using well-known coherent rake combining is adopted to provide packet data services of around a few Mbps [4]. However, for packet services of data higher than 100Mbps, the channel frequency-selectivity gets severer and the throughput performance of DS-CDMA HARQ using rake combining severely degrades.

Frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion can take advantage of the channel frequency-selectivity and offers a much improved throughput performance [5]-[6]. The conventional FDE requires the guard interval (GI) insertion to avoid inter-block interference (IBI). This GI insertion reduces the throughput. Recently, overlap FDE that needs no GI insertion was proposed [7]-[9]. We evaluated the throughput performance of DS-CDMA HARQ using overlap FDE [10]-[11]. It was shown that the throughput performance of DS-CDMA HARQ with overlap FDE is better than that with conventional FDE using GI insertion. So far, ideal channel estimation was assumed.

In the previous paper [12], we proposed a repetition pilot-assisted channel estimation based on the MMSE criterion suitable for overlap FDE (called MMSE-RPACE). In this paper, we apply MMSE-RPACE to DS-CDMA HARQ using overlap FDE and evaluate by computer simulation the throughput performance.

The rest of this paper is organized as follows. Sect. II introduces DS-CDMA HARQ with overlap FDE. Sect. III presents MMSE-RPACE. In Sect. IV, computer simulation results are presented. Sect. V concludes this paper.

II. DS-CDMA HARQ WITH OVERLAP FDE [11]

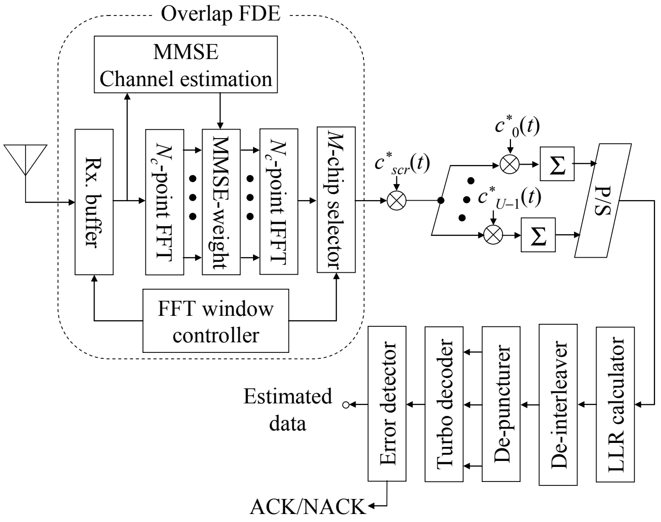

Fig. 1 Receiver structure.

Figure 1 illustrates the receiver structure. In this paper, we use a chip-spaced discrete-time representation. We use the turbo coding and packet combining based on Chase combining [13] for HARQ.

A. Received Signal

At the transmitter, an information bit sequence is turbo encoded and bit-interleaved before transforming into the data-modulated symbol sequence. 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~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~U−1. After multiplexing all U chip sequences, the multicode chip sequence is multiplied by a scramble code $\{c_{scr}(t);\ t=\ldots,-1,0,1,\ldots\}$ to obtain the multicode DS-CDMA signal as

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

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

The transmitted signal is received via a frequency-selective fading channel. In this paper, the channel is assumed to be an L-path frequency-selective block fading channel. It is assumed that the same packet has been retransmitted Q times (including the original packet). The trth (tr=0~Q–1) received packet in a time interval of t=0~N_c–1 can be expressed as

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

where E_c and T_c respectively represent the chip energy and chip duration. $h_l^{(tr)}$ and τ_l are the complex valued path gain and time delay of the lth path, respectively, and $\eta^{(tr)}(t)$ is the additive white Gaussian noise (AWGN) having one-sided power spectrum density N_0.

B. Overlap FDE

In overlap FDE, the received signal is divided into a sequence of M ($<N_c$)-chip blocks, where N_c is the block size of fast Fourier transform (FFT). An N_c-point FFT is applied to transform an N_c-chip interval of the received signal, which has an M-chip block of interest in its center, into the frequency-domain signal, as shown in Fig. 2. The residual IBI after MMSE-FDE is a circular convolution of the IBI and the MMSE-FDE impulse response. Since the MMSE-FDE impulse response concentrates at a vicinity of τ=0 [11], the residual IBI is localized only near the both ends of N_c-chip FFT block. This can be exploited by overlap FDE.

After carrying out FDE and packet combining, a time-domain N_c-chip block is obtained by N_c-point inverse FFT (IFFT). Since the residual IBI is stronger at both edges of the equalized N_c-chip block, only the central M-chip sequence is picked up from the N_c-chip block to suppress the residual IBI. Consecutive FFT interval overlaps. By reducing M, the residual IBI can be more suppressed, but the number of FFT/IFFT operations per N_c-chip block increases by a factor of N_c/M. In this paper, M is set to $N_c/2$.

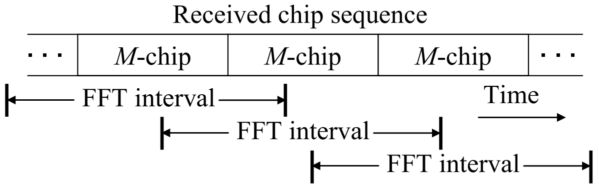

Fig. 2 Overlap FDE.

C. MMSE-FDE and Packet Combining

Eq. (2) can be rewritten as

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

where the first and second terms respectively represent the desired signal and the IBI component, which can be expressed as

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

with $u(t)$=0 (1) for t<0 (0≤t).

N_c-point FFT is applied to transform the received chip block $\{r^{(tr)}(t);\ t=0\sim N_c-1\}$ into the frequency-domain received signal $\{R^{(tr)}(k);\ k=0\sim N_c-1\}$ as

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

where $S(k)$ is the kth frequency-component of $s(t)$, and $H^{(tr)}(k)$, $N^{(tr)}(k)$, and $\Pi^{(tr)}(k)$ are respectively the channel gain, the IBI component, and the noise component at the kth frequency of the trth retransmitted chip block and are given as

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

Joint FDE and packet combining is carried out based on the MMSE criterion as

$$\begin{aligned}\hat{R}(k)&=\sum_{q=0}^{Q-1}w^{(q)}(k)R^{(q)}(k)\\ &=\hat{H}(k)S(k)+\hat{N}(k)+\hat{\Pi}(k)\end{aligned}, \tag{7}$$

where

$$\begin{cases}\hat{H}(k)=\sum_{q=0}^{Q-1}w^{(q)}(k)H^{(q)}(k)\\ \hat{N}(k)=\sum_{q=0}^{Q-1}w^{(q)}(k)N^{(q)}(k)\\ \hat{\Pi}(k)=\sum_{q=0}^{Q-1}w^{(q)}(k)\Pi^{(q)}(k)\end{cases} \tag{8}$$

and $\{w^{(q)}(k);\ k=0\sim N_c-1\}$ is the MMSE packet combining weight given by [11]

$$w^{(q)}(k)=\frac{\{H^{(q)}(k)\}^*}{N_cU\sum_{q'=0}^{Q-1}\frac{\sigma_q^2}{\sigma_{q'}^2}\left|H^{(q')}(k)\right|^2+2\sigma_q^2}, \tag{9}$$

where σ_q^2 represents the IBI plus noise power of the qth retransmitted packet.

After FDE and packet combining, N_c-point IFFT is applied to obtain the time-domain chip block. Only the central M-chip sequence is picked up from the N_c-chip block to suppress the residual IBI. After getting the chip sequence of a whole packet, despreading is applied to obtain a sequence of soft decision symbol associated with $\{d_u(i)\}$.

D. Error Detection

Using the soft decision symbol sequence, a sequence of log-likelihood ratio (LLR) is computed. After de-interleaving and depuncturing, turbo decoding is carried out using LLR. If errors are detected in the received packet, the NACK signal is sent to the transmitter to request the retransmission. In this paper, ideal error detection and ACK/NACK transmission are assumed.

III. MMSE-RPACE

In Ref. [14], a pilot-assisted channel estimation method based on the MMSE criterion (MMSE-PACE) was proposed. A set of channel gains $\{H^{(tr)}(k);\ k=0\sim N_c-1\}$ are estimated by using the GI-inserted pilot chip sequence $p(t \bmod N_c)$, $t=-N_g\sim N_c-1$. However, if the GI insertion is not used, the channel estimation accuracy degrades due to the severe residual IBI. To overcome this problem, we proposed MMSE-RPACE [12]. In this section, MMSE-RPACE is introduced.

A. Instantaneous Channel Gain Estimate

We consider the pilot chip block reception in the trth ($tr=0\sim Q-1$) retransmission. An $N_c/2$-chip pilot sequence $p(t)$, $t=0\sim N_c/2-1$, with $|p(t)|=1$ is repeated twice to generate the pilot chip block of N_c chips. 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 chip block is free from the IBI if the maximum delay time $\tau_{\max}$ of the channel is shorter than $N_c/2$ chips.

The received pilot chip block $\{r_p^{(tr)}(t);\ t=N_c/2\sim N_c-1\}$ in an interval of $t=N_c/2\sim N_c-1$ can be expressed as

$$r_p^{(tr)}(t)=\sqrt{U\frac{2E_c}{T_c}}\sum_{l=0}^{L-1}h_l^{(tr)}p(t \bmod N_c/2)+\eta^{(tr)}(t)\,. \tag{10}$$

$N_c/2$-point FFT is carried out on the received pilot chip block $\{r_p^{(tr)}(t);\ t=N_c/2\sim N_c-1\}$ over an interval of $t=N_c/2\sim N_c-1$ to obtain the frequency-domain received pilot $\{R_{p,1}^{(tr)}(y);\ y=0\sim N_c/2-1\}$ as

$$R_{p,1}^{(tr)}(y)=\sqrt{U}H^{(tr)}(2y)P(y)+\Pi_1^{(tr)}(2y)\,, \tag{11}$$

where $H^{(tr)}(2y)$ and $\Pi_1^{(tr)}(2y)$ represent the channel gain and the noise at the frequency of $k=2y$, $y=0\sim N_c/2-1$, respectively. $P(y)$ is given by

$$P(y)=\sum_{y=0}^{N_c/2-1}p(t)\exp\left(-j2\pi y\frac{t}{(N_c/2)}\right). \tag{12}$$

The channel gain estimate obtained by the latter half of the received pilot chip block is denoted by $\tilde{H}_1^{(tr)}(2y)$. If a pseudo-noise (PN) sequence is used as the pilot $p(t)$, $\{P(y);\ y=0\sim N_c/2-1\}$ is not constant in frequency-domain. To get the accurate channel gain estimate while reducing the noise, we apply MMSE channel estimation [14]. The MMSE channel estimation minimizes the mean square error between $\tilde{H}_1^{(tr)}(2y)$ and $H^{(tr)}(2y)$ and is done as

$$\tilde{H}_1^{(tr)}(2y)=R_{p,1}^{(tr)}(y)X(y) \tag{13}$$

with

$$X(y)=\frac{P^*(y)}{\dfrac{N_c}{2}|P(y)|^2+\left(U\dfrac{E_c}{T_c}\dfrac{1}{\sigma^2}\right)^{-1}}\,, \tag{14}$$

where σ^2 is the noise power. We need to estimate the received signal power and the noise power to obtain $X(y)$. This can be easily done according to Ref. [14]. In this paper, we assume ideal power estimation of the received signal and the noise.

B. Noise Reduction and Interpolation

The delay time-domain windowing technique can be applied to reduce the noise [15]. Since only the channel gains at even frequencies (i.e., $k=2y$) can be estimated using the pilot, the interpolation technique is necessary to estimate the channel gain at odd frequencies (i.e., $k=2y+1$). The delay time-domain windowing technique can perform interpolation while reducing the noise.

$N_c/2$-point IFFT is applied to $\{\tilde{H}_1^{(tr)}(2y);\ y=0\sim N_c/2-1\}$ to obtain the estimated channel impulse response $\{\tilde{h}_1^{(tr)}(\tau);\ \tau=0\sim N_c/2-1\}$ as

$$\tilde{h}_1^{(tr)}(\tau)=\frac{2}{N_c}\sum_{y=0}^{N_c/2-1}\tilde{H}_1^{(tr)}(2y)\exp\left(j2\pi y\frac{\tau}{(N_c/2)}\right). \tag{15}$$

Assuming that the channel impulse response $\{h^{(tr)}(\tau);\ \tau=0\sim\tau_{\max}-1\}$ exists within a delay time interval of $\tau=0\sim\tau_{\max}-1$, $\tilde{h}_1^{(tr)}(\tau)$ is replaced by zeros for $\tau=\tau_{\max}\sim N_c/2-1$. Then, applying N_c-point FFT, the noise-reduced and interpolated channel gain estimate $\{\tilde{H}_1^{(tr)}(k);\ k=0\sim N_c-1\}$ is obtained as

$$\begin{aligned}\tilde{H}_1^{(tr)}(k)&=\sum_{\tau=0}^{\tau_{\max}-1}\tilde{h}_1^{(tr)}(\tau)\exp\left(-j2\pi k\frac{\tau}{N_c}\right)\\&=\sum_{y=0}^{N_c/2-1}A(k-2y)\tilde{H}_1^{(tr)}(2y)\end{aligned}\,, \tag{16}$$

where $A(n)$ is given by

$$A(n)=\frac{1}{N_c}\exp\left(j\pi(\tau_{\max}-1)\frac{n}{N_c}\right)\frac{\sin\left(\pi\tau_{\max}\frac{n}{N_c}\right)}{\sin\left(\pi\frac{n}{N_c}\right)}. \tag{17}$$

C. *Improved Channel Estimation*

The first half of the pilot chip block can also be used to improve the channel estimation accuracy, if $\tau_{\max}$ is known. The same MMSE channel estimation is applied using the first half of the pilot chip block $\{r_p^{(tr)}(t);\ t=\tau_{\max}\sim N_c/2+\tau_{\max}-1\}$ over an interval of $t=\tau_{\max}\sim\tau_{\max}+N_c/2-1$. The frequency-domain pilot signal obtained by $N_c/2$-point FFT is given by

$$\begin{aligned}R_{p,0}^{(tr)}(y)&=\sum_{t=0}^{N_c/2-1}r_p^{(tr)}(t+\tau_{\max})\exp\left(-j2\pi y\frac{t}{(N_c/2)}\right)\\&=\sqrt{U}H^{(tr)}(2y)P(y)\exp\left(j2\pi y\frac{\tau_{\max}}{(N_c/2)}\right)+\Pi_0^{(tr)}(2y)\end{aligned}, \tag{18}$$

where $\Pi_0^{(tr)}(2y)$ is the noise. The MMSE channel estimation is applied using $\{R_{p,0}^{(tr)}(y);\ y=0\sim N_c/2-1\}$, similar to Eqs. (13)~(16), to obtain the instantaneous channel gain estimates $\{\tilde{H}_0^{(tr)}(2y);\ y=0\sim N_c/2-1\}$.

A simple averaging of channel gain estimates obtained from the latter and first half of the pilot chip block is applied to improve the channel estimation accuracy as

$$\tilde{H}^{(tr)}(2y)=\frac{1}{2}\left\{\tilde{H}_0^{(tr)}(2y)\exp\left(-j2\pi y\frac{\tau_{\max}}{(N_c/2)}\right)+\tilde{H}_1^{(tr)}(2y)\right\}. \tag{19}$$

The first half and latter half pilot chip blocks are overlapped. Therefore, noises $\Pi_0^{(tr)}(2y)$ and $\Pi_1^{(tr)}(2y)$ are partially correlated. However, if $\tau_{\max}/N_c<<1$, the degradation in channel estimation accuracy may not be negligible.

Replacing the channel gain in Eq. (9) by the channel gain estimate obtained in Eq. (19), the MMSE packet combining weight is computed.

IV. COMPUTER SIMULATION

The simulation condition is summarized in Table 1. The packet structure is illustrated in Fig. 4. One packet is composed of N_c chips pilot and $16N_c$ coded data chips with N_c=256. We assume an L=16-path frequency-selective block Rayleigh fading channel having uniform power delay profile. The channel gains stay constant over one packet. Full code-multiplexing with $U=SF$=16 is considered.

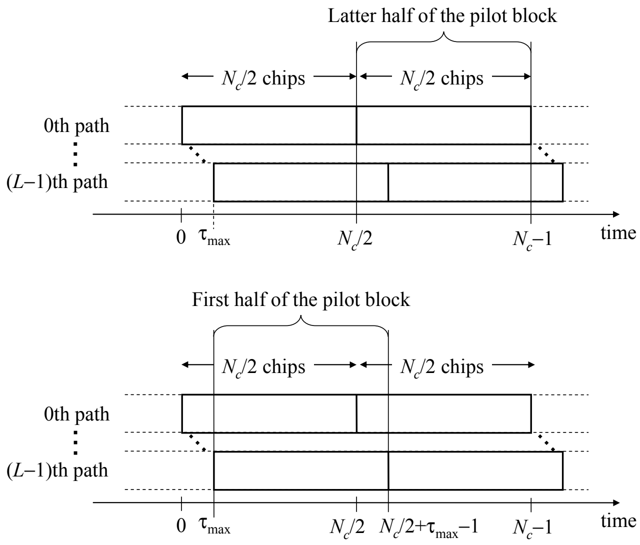

Fig. 3 MMSE-RPACE.

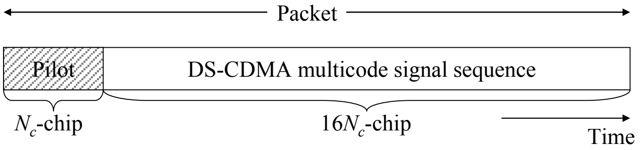

Fig. 4 Packet structure.

Table 1 Simulation condition.

Turbo coding	Info. sequence length	K=1024
	Encoder	(13, 15)RSC
	Coding rate	R=1/2, 3/4
	Interleaver	Block
	Decoder	Log-MAP with 8 iterations
ARQ	Combining	Chase
	Max. no. of retrans.	100
Data modulation	QPSK, 16QAM	
DS-CDMA	Spreading factor	SF=16
	No. of code mux.	$U=SF$
Channel model	Frequency-selective block Rayleigh	
	No. of paths	L=16
	Power delay profile	Uniform
	Time delay	$\tau_l=l$, $l=0\sim L-1$
	Doppler frequency	f_D=0
Overlap FDE	FFT/IFFT block size	N_c=256
	No. of chips to pick up	M=128
Channel estimation	Pilot sequence	PN sequence
	Channel gain	MMSE
	SNR estimation	Ideal
	Time delay estimation	Ideal

Figure 5 plots the throughput performance of DS-CDMA HARQ using overlap FDE and MMSE-RPACE. For comparison, the throughput performance assuming ideal channel estimation is also plotted. When using the latter half of the pilot chip block only, the E_s/N_0 degradation from ideal channel estimation at a throughput of 1.5 bps/Hz is 1.9dB for R=1/2 and 16QAM. If $\tau_{\max}$ is known at the receiver, the first half of the received pilot block can be used for improving the channel estimation accuracy. The required E_s/N_0 degradation at 1.5 bps/Hz is only about 1.5dB.

Figure 6 shows the throughput comparison of DS-CDMA HARQ using overlap FDE and that using conventional FDE with N_g=32-chip GI [14]. Overlap FDE does not require the GI insertion and hence, achieves higher throughput.

The conventional FDE requires GI insertion longer than the maximum time delay of the channel. This reduces the throughput. If the maximum time delay of the channel exceeds the GI length, the throughput performance degrades due to IBI. Therefore, IBI cancellation technique [16] is necessary. However, the introduction of IBI cancellation increases the complexity of the receiver which is problematic in the downlink case. It can be understood from Fig. 6 that overlap FDE provides better throughput performance than the conventional FDE at the cost of the receiver complexity increase by a factor of N_c/M.

V. CONCLUSION

In this paper, we applied MMSE-RPACE to DS-CDMA HARQ using overlap FDE and evaluated the achievable throughput performance by computer simulation. We showed that MMSE-CPACE provides a good throughput performance even if only the latter half of the pilot chip-block is used. If $\tau_{\max}$ is known at the receiver, the first half of the pilot chip-block can be used to improve the channel estimation accuracy. We also showed that even when a practical MMSE-RPACE is used, overlap FDE can provide better throughput performance than GI-inserted conventional FDE.

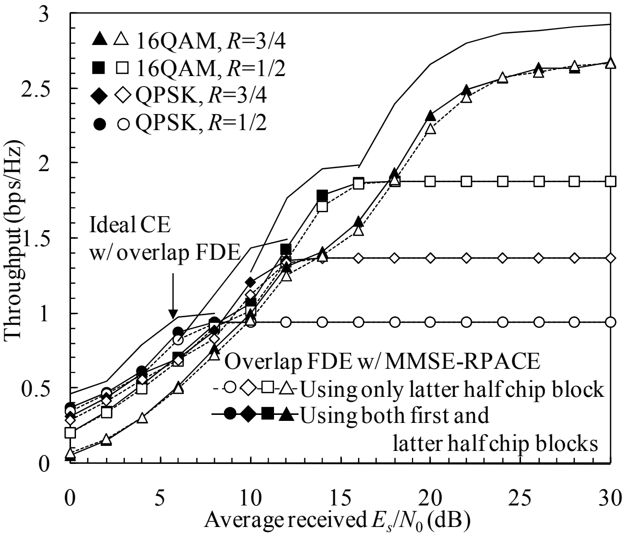

Fig. 5 Throughput performance of DS-CDMA HARQ using overlap FDE and MMSE-RPACE.

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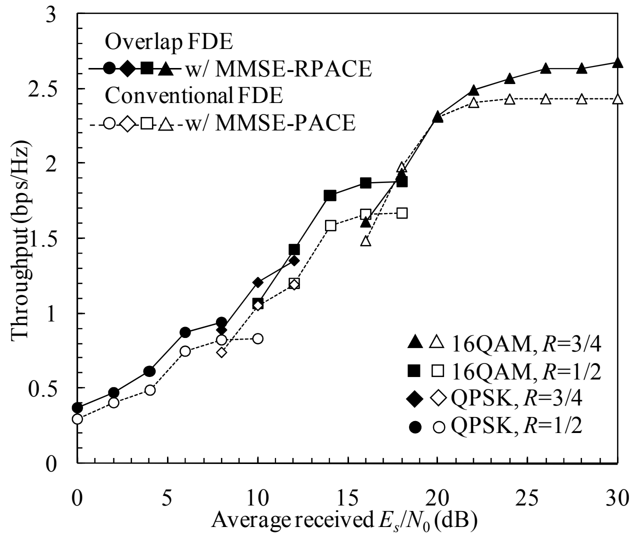

Fig. 6 Throughput comparison between overlap FDE and conventional FDE.

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