Polynomial Prediction RLS Channel Estimation for DS-CDMA Frequency-domain Equalization

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Abstract—**Recently, we have proposed an adaptive channel estimation (CE) scheme using one-tap recursive least square (RLS) algorithm (adaptive RLS-CE), where the forgetting factor is adapted to the changing channel condition by the least mean square (LMS) algorithm, for direct sequence-code division multiple access (DS-CDMA) with frequency-domain equalization (FDE). However, the tracking ability for adaptive RLS-CE is limited since the channel estimate obtained from the previously received block is used. In this paper, we introduce the polynomial prediction to improve the tracking ability. We evaluate the bit error rate (BER) performance of DS-CDMA using polynomial prediction RLS-CE in a frequency-selective fast Rayleigh fading channel by computer simulation.**

Keywords-components; DS-CDMA, frequency-domain equalization, channel estimation, RLS algorithm, prediction

I. INTRODUCTION

The 4th generation (4G) mobile communication systems [1] which provide broadband wireless services of e.g. 100Mbps to 1Gbps are expected to emerge around 2015. In the present 3rd generation (3G) systems, direct sequence-code division multiple access (DS-CDMA) is adopted as the wireless access technique [2]. However, since the broadband wireless channel is severely frequency-selective, the bit error rate (BER) performance of DS-CDMA with rake combining significantly degrades. Replacing the rake combing by frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion can provide a significantly good BER performance [3], [4]. FDE requires accurate channel estimation since the FDE weight depends on the channel transfer function.

Time-domain pilot-assisted channel estimation (CE) was proposed for single-carrier transmission in [5]. After the channel impulse response is estimated according to the least-sum-of-squared-error (LSSE) criterion, the channel transfer function is obtained by applying fast Fourier transform (FFT). Frequency-domain pilot-assisted CE was proposed in [6], [7]. The received pilot is transformed by FFT into the frequency-domain pilot and then the pilot modulation is removed using zero-forcing (ZF) or least square (LS) technique.

As the pilot signal, the Chu sequence [8] that has the constant amplitude in both time- and frequency-domain can be used. However, the available number of Chu sequences is limited. For example, it is only 128 for the case of 256-bit period [8]. A very large number of pilots can be generated by using a partial sequence taken from a long PN sequence. However, since the frequency spectrum of the partial PN sequence is not constant, the use of zero-forcing CE (ZF-CE) produces the noise enhancement [9]. The noise enhancement can be mitigated by using the MMSE-CE [9] instead of using ZF-CE. The channel estimation accuracy of MMSE-CE is almost insensitive to the pilot chip sequence.

Recently, we proposed a 2-step maximum likelihood channel estimation (MLCE) to further improve the estimation accuracy [10]. However, the 2-step MLCE has poor tracking ability against fast fading since it assumes a block fading in which the channel gains stay constant over several blocks. Adaptive channel estimation using recursive least square (RLS) algorithm (RLS-CE) was proposed to track the time-varying channels [11]. In [11], the superimposed training sequences are used for RLS-CE for orthogonal frequency division multiplexing (OFDM). However, the forgetting factor of RLS algorithm was not adapted in [11] and needs to be changed according to the change in the channel condition. More recently, we have proposed an adaptive channel estimation scheme using one-tap RLS algorithm (adaptive RLS-CE), where the forgetting factor is adapted to the changing channel condition by the least mean square (LMS) algorithm, for DS-CDMA with FDE [12]. However, the tracking ability of adaptive RLS-CE against fast fading is limited since the channel estimate obtained from the previously received block is used.

In this paper, we propose a polynomial prediction RLS-CE to improve the tracking ability of adaptive RLS-CE [12]. We evaluate, by computer simulation, the BER performance of DS-CDMA using polynomial prediction RLS-CE in a frequency-selective fast Rayleigh fading channel. The achievable BER performance is compared with those using adaptive RLS-CE [12] and using 1st order interpolation MMSE-CE.

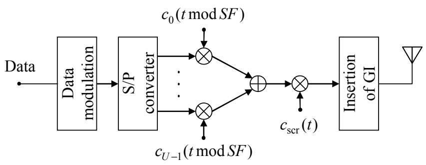
II. TRANSMISSION SYSTEM MODEL

A. Overall transmission system model

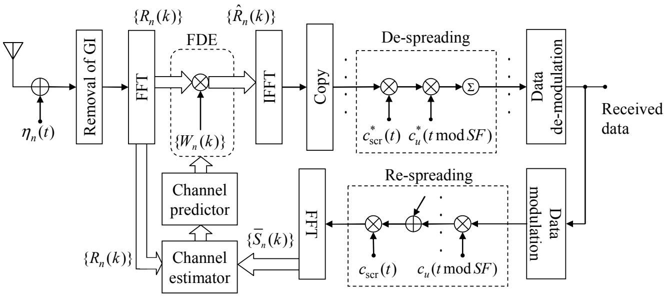
The transmission system model for multicode DS-CDMA with FDE is illustrated in Fig 1. Throughout the paper, the chip-spaced discrete-time signal representation is used.

At the transmitter, a binary data sequence is transformed into data-modulated symbol sequence and then converted to U parallel streams by serial-to-parallel (S/P) conversion. Then, each parallel stream is divided into a sequence of blocks of N_c/SF symbols each, where N_c and SF denote the size of the chip-block and the spreading factor, respectively. A sequence of N chip-blocks is defined as a frame. The mth data symbol of the nth chip-block (n=0~N−1) in the uth stream is represented by $d_{n,u}(m)$, m=0~N_c/SF−1. $d_{n,u}(m)$ is spread by multiplying it with an orthogonal spreading sequence $\{c_u(t); t=0\sim SF-1\}$. The resultant U chip-blocks of N_c chips each are added and further multiplied by a common scramble sequence $\{c_{\rm scr}(t); t=\ldots,1,0,1,\ldots\}$ to make the resultant multicode DS-CDMA chip-block like white-noise. The last N_g chips of each N_c chip-block is copied as a cyclic prefix and inserted into the guard interval (GI) placed at the beginning of each chip-block, as illustrated in Fig. 2. For channel estimation, one pilot chip-block is transmitted every N−1 data chip-blocks to constitute a frame of N chip-blocks, as shown in Fig. 3.

The GI-inserted chip-block is transmitted over a frequency-selective fading channel and is received at a receiver. After removal of the GI, the received chip-block is decomposed by an N_c-point FFT into the frequency-domain signal. At the pilot chip-block, the adaptive RLS-CE is carried out. At the data chip-block, using the channel estimate, a series of FDE, N_c-point inverse FFT (IFFT), de-spreading, and data de-modulation is performed. Then, the chip-block replica is regenerated and the adaptive RLS-CE is carried out using the chip-block replica as a pilot. The channel gain for FDE is obtained by the polynomial prediction using M past channel estimates.



(a) Transmitter



(b) Receiver

Figure 1. Transmitter/receiver structure for multicode DS-CDMA with FDE.

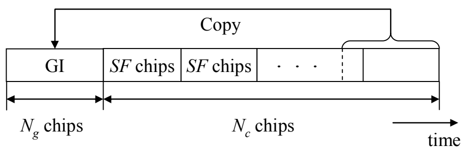


Figure 2. Chip-block structure.

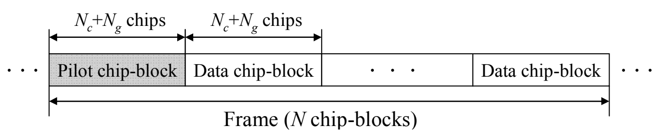


Figure 3. Frame structure.

B. Received Signal

The nth chip-block $\{\tilde{s}_n(t); t=0\sim N_c-1\}$ can be expressed, using the equivalent lowpass representation, as

$$\tilde{s}_n(t)=\sqrt{2P}s_n(t) \tag{1}$$

with

$$s_n(t)=\left[\sum_{u=0}^{U-1}d_{n,u}\left(\left\lfloor\frac{t}{SF}\right\rfloor\right)c_u(t\,\mathrm{mod}\,SF)\right]c_{\rm scr}(t)\,, \tag{2}$$

where P is the transmit power and $\lfloor x\rfloor$ represents the largest integer smaller than or equal to x. After inserting the GI of N_g chips, the nth chip-block is transmitted.

The propagation channel is assumed to be a frequency-selective fading channel having chip-spaced L discrete paths, each subjected to independent fading. The channel impulse response $h_n(\tau)$ can be expressed as

$$h_n(\tau)=\sum_{l=0}^{L-1}h_{n,l}\delta(\tau-\tau_l)\,, \tag{3}$$

where $h_{n,l}$ and τ_l are the complex-valued path gain and time delay of the lth path (l=0~L−1), respectively , with $\sum_{l=0}^{L-1}E[|h_{n,l}|^2]=1$ (E[.] denotes the ensemble average operation). In this paper, we assume that the maximum time delay difference $\tau_{L-1}-\tau_0$ of the channel is shorter than the GI length. We assume that the path gains stay constant over one chip-block but they vary block by block.

The nth received chip-block $\{r_n(t); t=0\sim N_c-1\}$ can be expressed as

$$r_n(t)=\sum_{l=0}^{L-1}h_{n,l}\tilde{s}_n(t-\tau_l)+\eta_n(t)\,, \tag{4}$$

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

C. FDE

After the removal of the GI, the received chip-block is transformed by an N_c-point FFT into the frequency-domain signal $\{R_n(k); k=0\sim N_c-1\}$. The kth frequency component $R_n(k)$ of the nth chip-block can be written as

$$\begin{aligned}R_n(k)&=\sum_{t=0}^{N_c-1}r_n(t)\exp\left(-j2\pi k\frac{t}{N_c}\right),\\&=H_n(k)S_n(k)+\Pi_n(k)\end{aligned} \tag{5}$$

where $H_n(k)$ is the channel gain, $S_n(k)$ is the signal component, and $\Pi_n(k)$ is the noise. They are given by

$$\begin{cases} S_n(k)=\sum_{t=0}^{N_c-1} s_n(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ H_n(k)=\sqrt{2P}\sum_{l=0}^{L-1} h_{n,l}\exp\left(-j2\pi k\frac{\tau_l}{N_c}\right) \\ \Pi_n(k)=\sum_{t=0}^{N_c-1}\eta_n(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \end{cases}. \quad (6)$$

One-tap FDE is carried out as

$$\hat{R}_n(k)=W_n(k)R_n(k), \quad (7)$$

where $W_n(k)$ is the MMSE-FDE weight and is given by [3], [4]

$$W_n(k)=\frac{H_n^*(k)}{UN_c|H_n(k)|^2+2\sigma^2} \quad (8)$$

with $2\sigma^2$ ($=2N_0N_c/T_c$) being the variance of $\Pi_n(k)$ and * denoting the complex conjugate operation. $H_n(k)$ and σ^2 are unknown to the receiver and need to be estimated. In this paper, $H_n(k)$ is estimated by polynomial prediction RLS-CE. σ^2 is estimated by [9].

N_c-point IFFT is applied to transform the frequency-domain signal $\{\hat{R}_n(k)$; k=0~N_c−1} into the time-domain chip-block $\{\hat{r}_n(t)$; t=0~N_c−1} as

$$\hat{r}_n(t)=\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{R}_n(k)\exp\left(j2\pi t\frac{k}{N_c}\right). \quad (9)$$

Finally, de-spreading is carried out on $\{\hat{r}_n(t)\}$, giving

$$\hat{d}_{n,u}(m)=\frac{1}{SF}\sum_{t=mSF}^{(m+1)SF-1}\hat{r}_n(t)c_u^*(t \bmod SF)c_{\mathrm{scr}}^*(t), \quad (10)$$

which is the decision variable associated with $d_{n,u}(m)$.

III. ADAPTIVE RLS-CE

We apply adaptive RLS-CE [12] to obtain the instantaneous channel estimate. The channel estimation using adaptive RLS-CE is performed as follows. The adaptive RLS-CE is carried out using the received pilot chip-block (n=0). Using the channel estimate, a series of FDE, N_c-point IFFT, de-spreading, and data de-modulation is performed on the 1st data chip-block in the frame. Then, the chip-block replica is regenerated and the RLS-CE is carried out using the chip-block replica as a pilot. The forgetting factor used in the RLS algorithm is updated using the LMS algorithm.

A. RLS-CE

Using RLS-CE, the instantaneous channel estimate $\tilde{H}_n(k)$ is obtained as [13]

$$\tilde{H}_n(k)=\overline{H}_{n-1}(k)+G_n(k)\xi_n(k), \quad (11)$$

where $\overline{H}_{n-1}(k)$ is the channel estimate obtained by RLS-CE and delay time-domain windowing technique in the (n−1)th block, $\overline{H}_{-1}(k)=0$, and $G_n(k)$ and $\xi_n(k)$ are respectively given by

$$\begin{cases} G_n(k)=S_n^*(k)/\Phi_n(k) \\ \xi_n(k)=R_n(k)-\overline{H}_{n-1}(k)S_n(k) \end{cases} \quad (12)$$

with

$$\Phi_n(k)=\lambda_{n-1}\Phi_{n-1}(k)+|S_n(k)|^2, \quad (13)$$

where $\Phi_{-1}(k)$ is a small positive constant [13]. The RLS-CE requires the knowledge of the transmitted chip-block {$S_n(k)$; k=0~N_c−1}, n≥1. However, since {$S_n(k)$; k=0~N_c−1} is unknown at the receiver, the frequency-domain signal replica {$\overline{S}_n(k)$; k=0~N_c−1} of the transmitted chip-block needs to be generated by the decision-feedback as described below.

The instantaneous channel estimate $\tilde{H}_n(k)$ is perturbed by the noise due to the AWGN. The delay time-domain windowing technique [14], [15] is applied to reduce the noise. {$\tilde{H}_n(k)$; k=0~N_c−1} is transformed by N_c-point IFFT into the instantaneous channel impulse response {$\tilde{h}_n(\tau)$; τ=0~N_c−1}. The actual channel impulse response is present only within the GI length, while the noise is spread over an entire delay-time range. Replacing $\tilde{h}_n(\tau)$ with zero's for N_g≤τ≤N_c−1 and applying N_c-point FFT, the improved channel estimate {$\overline{H}_n(k)$; k=0~N_c−1} is obtained.

B. Adaptive forgetting factor λ

Using LMS algorithm, the forgetting factor is updated as [15]

$$\lambda_n(k)=\lambda_{n-1}(k)+\mu\,\mathrm{Re}\left[\Psi_{n-1}(k)S_n(k)\xi_n^*(k)\right], \quad (14)$$

where μ is the step size and $\Psi_n(k)$ is given by

$$\Psi_n(k)=\left(1-G_n(k)S_n(k)\right)\Psi_{n-1}(k)+\Theta_n(k)S_n^*(k)\xi_n(k) \quad (15)$$

with [15]

$$\Theta_n(k)=\lambda_n^{-1}\left\{\left(1-G_n(k)S_n(k)\right)^2\Theta_{n-1}(k)+|G_n(k)|^2-\Phi_n^{-1}(k)\right\}. (16)$$

The forgetting factor $\lambda_n(k)$ depends on statistical characteristics of the fading channel. Statistical characteristics are identical for all frequencies. Therefore, in this paper, $\lambda_n=(1/N_c)\sum_{k=0}^{N_c-1}\lambda_n(k)$ is used to suppress the noise.

The adaptive RLS-CE requires the transmitted chip-block replica {$\overline{S}_n(k)$; k=0~N_c−1}, n≥1. The replica generation is done as follows. First, FDE is carried out using the predicted channel gain (the channel prediction is discussed in Sect. IV). After performing a series of FDE, N_c-point IFFT, de-spreading, and data de-modulation on the nth chip-block, the tentatively detected symbol sequence {$\overline{d}_{n,u}(m)$; m=0~N_c/SF−1}, u=0~U−1, is spread to generate the transmitted chip-block replica {$\overline{s}_n(k)$; t=0~N_c−1}:

$$\overline{s}_n(t)=\left[\sum_{u=0}^{U-1}\overline{d}_{n,u}\left(\left\lfloor\frac{t}{SF}\right\rfloor\right)c_u(t \bmod SF)\right]c_{\mathrm{scr}}(t). \quad (17)$$

The chip-block replica is transformed by an N_c-point FFT into the frequency-domain signal {$\overline{S}_n(k)$; k=0~N_c−1}. The kth frequency component $\overline{S}_n(k)$ of the transmitted chip-block replica is obtained as

$$\bar{S}_n(k)=\sum_{t=0}^{N_c-1}\bar{s}_n(t)\exp\left(-j2\pi k\frac{t}{N_c}\right). \quad (18)$$

Using $\{\bar{S}_n(k);k=0\sim N_c-1\}$ instead of $\{S_n(k); k=0\sim N_c-1\}$ in Eqs. (11)-(16), the channel estimate $\{\tilde{H}_n(k);k=0\sim N_c-1\}$ for the nth chip-block is obtained by RLS algorithm. The forgetting factor λ_n is updated by LMS algorithm.

IV. POLYNOMIAL PREDICTION

The channel may vary quickly in time in a fast fading channel. The channel estimate obtained at the reception of the nth chip-block is old at the reception of the $(n+1)$th chip-block. Therefore, we need to predict the channel at the $(n+1)$th chip-block reception from the past channel estimates. The predicted channel of $H_{n+1}(k)$ is denoted by $\hat{H}_{n+1}(k)$. We apply the polynomial prediction using M past channel estimates $\{\bar{H}_{n-m}(k);m=0\sim M-1\}$.

We use the following cost function for the polynomial prediction:

$$g_n(k)=\sum_{m=0}^{M-1}\left|\hat{H}_{n-m}(k)-\bar{H}_{n-m}(k)\right|^2, \quad (19)$$

where $\{\hat{H}_{n-m}(k);m=0\sim M-1\}$ is the channel gain approximated by the N_ith order polynomial as

$$\hat{H}_{n-m}(k)=\sum_{i=0}^{N_i}a_i(k)\cdot(n-m)^i, \quad (20)$$

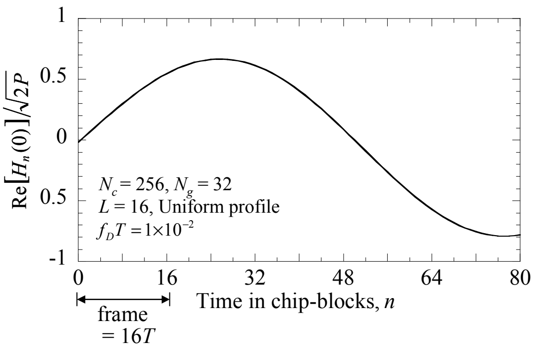


Figure 4. Time variation of fading channel.

where $\{a_i(k); i=0\sim N_i\}$ is polynomial coefficient. We want to find $\{a_i(k); i=0\sim N_i\}$ that minimizes $g_n(k)$. Solving $\{\partial g_n(k)/\partial a_i(k)=0;i=0\sim N_i\}$ gives

$$[a_0(k)\ \ a_1(k)\ \ \cdots\ \ a_{N_i}(k)]^T=$$

$$\begin{bmatrix} M & \sum_{m=0}^{M-1}(n-m) & \cdots & \sum_{m=0}^{M-1}(n-m)^{N_i} \\ \sum_{m=0}^{M-1}(n-m) & \sum_{m=0}^{M-1}(n-m)^2 & \cdots & \sum_{m=0}^{M-1}(n-m)^{N_i+1} \\ \vdots & \vdots & \ddots & \vdots \\ \sum_{m=0}^{M-1}(n-m)^{N_i} & \sum_{m=0}^{M-1}(n-m)^{N_i+1} & \cdots & \sum_{m=0}^{M-1}(n-m)^{2N_i} \end{bmatrix}^{-1}\begin{bmatrix} \sum_{m=0}^{M-1}\bar{H}_{n-m}(k) \\ \sum_{m=0}^{M-1}(n-m)\bar{H}_{n-m}(k) \\ \vdots \\ \sum_{m=0}^{M-1}(n-m)^{N_i}\bar{H}_{n-m}(k)\end{bmatrix}. \quad (21)$$

Using $\{a_i(k); i=0\sim N_i\}$, $\hat{H}_{n+1}(k)$ is predicted as

$$\hat{H}_{n+1}(k)=\sum_{i=0}^{N_i}a_i(k)\cdot(n+1)^i. \quad (22)$$

According to the Wiener theory [15], the linear prediction using M past chip-blocks needs to determine the inverse of an M by M matrix. For the proposed polynomial prediction, the size of inverse matrix does not depend on M but N_i (see Eq. (21)). Fig. 4 shows the channel variation when the normalized Doppler frequency $f_DT=1\times10^{-2}$, where T $(=(N_c+N_g)T_c)$ is the chip-block length. As an example, let assume a CDMA system of chip-rate $1/T_c$=100 Mcps (bandwidth of 100 MHz), 3.5 GHz carrier frequency, an FFT block size of N_c=256 chips, and a GI of N_g=32 chips. The value of f_DT is 1×10^{-2} when the moving speed is 1071 km/h. It is seen from Fig. 4 that the channel over an internal of 4 frames (64 blocks) can be well predicted by the N_i=2nd order polynomial for such a high fading rate as $f_DT=1\times10^{-2}$. Therefore, the polynomial prediction needs to determine the inverse of 3 by 3 matrix at most if $M\leq64$. For instance, when M=64, the proposed polynomial prediction requires a complexity that is much smaller than that of the linear prediction.

V. COMPUTER SIMULATION

We assume 16QAM data modulation, an FFT block size of N_c=256 chips and a GI of N_g=32 chips. One pilot chip-block is transmitted every 15 data chip-blocks (i.e., N=16). We assume the spreading factor $SF=U$=16 and an L=16-path frequency-selective Rayleigh fading channel having uniform power delay profile. The step size μ of LMS algorithm (see Eq. (14)) is set to $\mu=5\times10^{-6}$ to provide the minimum BER when $f_DT=1\times10^{-2}$.

Fig. 5 shows the impact of number of chip-blocks, M, for polynomial prediction at E_b/N_0=24 (dB) and $f_DT=1\times10^{-2}$. It is seen that the BER can be minimized when M=7 (19) if N_i=1 (2). Below, we use M=7 for N_i=1 and M=19 for N_i=2.

The simulated BER performance of multicode DS-CDMA with FDE is plotted in Fig. 6 as a function of the average received bit energy-to-AWGN noise power spectrum density ratio E_b/N_0 $(=0.25(P\cdot SF\cdot T_c/N_0)(1+N_g/N_c)N/(N-1))$ at $f_DT=1\times10^{-2}$. The BER performances using adaptive RLS-CE [12], 1st order interpolation MMSE-CE, and ideal CE are also plotted for comparison. It is seen from Fig. 6 that polynomial prediction RLS-CE provides better BER performance than adaptive RLS-CE [12] and 1st order interpolation MMSE-CE at E_b/N_0>18 (dB). The N_i=2nd order polynomial prediction RLS-CE provides a inferior BER performance to the N_i=1st order polynomial prediction RLS-CE at E_b/N_0<18 (dB). This is due to the error propagation.

Fig. 7 shows the impact of fading rate on the achievable BER as a function of the normalized Doppler frequency f_DT at E_b/N_0=24 (dB). For comparison, the BER performance using adaptive RLS-CE [12] and 1st order interpolation MMSE-CE are also plotted. It is seen from Fig. 7 that polynomial prediction RLS-CE provides better BER performance than adaptive RLS-CE [12]. N_i=2nd order polynomial prediction RLS-CE provides a better BER performance than 1st order interpolation MMSE-CE.

VI. CONCLUSIONS

In this paper, we proposed a polynomial prediction RLS-CE for multicode DS-CDMA with FDE to improve the tracking ability of adaptive RLS-CE [12]. In polynomial prediction RLS-CE, the channel gain for FDE is obtained by the polynomial prediction using the past channel estimates which are obtained by adaptive RLS-CE. The proposed polynomial prediction requires much lower complexity than the linear prediction based on the Wiener theory. It was shown by computer simulation that the proposed polynomial prediction RLS-CE has better tracking ability against fast fading and provides better BER performance than adaptive RLS-CE and 1st order interpolation MMSE-CE in a high E_b/N_0 region.

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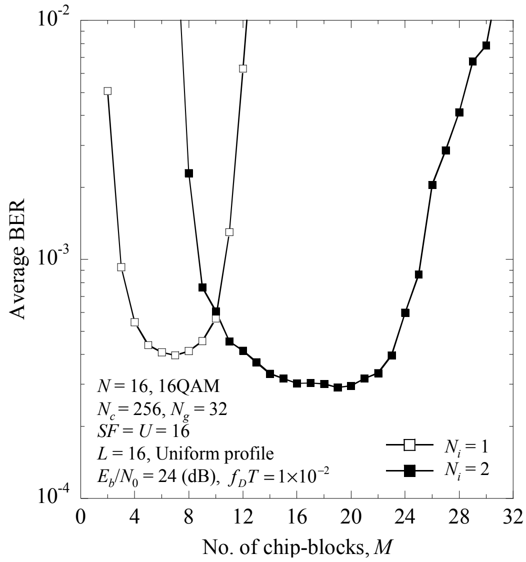

Figure 5. Impact of M.

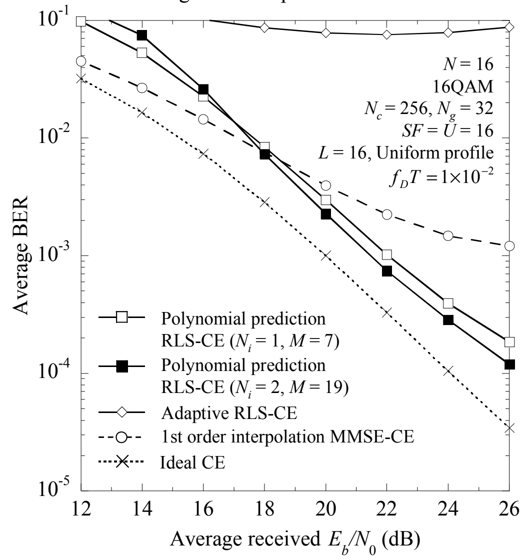

Figure 6. BER performance comparison.

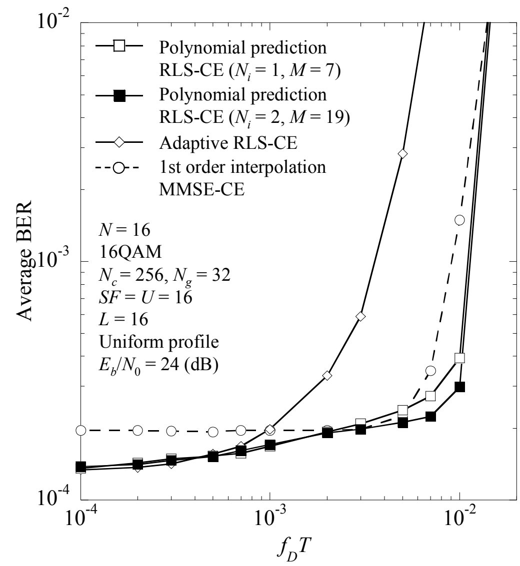

Figure 7. Impact of fading rate.