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DS-CDMA FREQUENCY-DOMAIN EQUALIZATION WITH RLS-BASED CHANNEL ESTIMATION

Yohei KOJIMA Kazuaki TAKEDA and Fumiyuki ADACHI
Dept. of Electrical and Communication Engineering, Graduate School of Engineering, Tohoku University
6-6-05 Aza-Aoba, Aramaki, Aoba-ku, Sendai, 980-8579 Japan

ABSTRACT

Frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion can provide a better downlink bit error rate (BER) performance of direct sequence code division multiple access (DS-CDMA) than the conventional rake combining in a frequency-selective fading channel. FDE requires accurate channel estimation. Recently, we proposed a pilot-assisted channel estimation based on the minimum mean square error (MMSE) criterion. Using MMSE-CE, the channel estimation accuracy is almost insensitive to the pilot chip sequence, and a good bit error rate (BER) performance is achieved. We also proposed a 2-step maximum likelihood channel estimation (MLCE) to further improve the estimation accuracy. However, the tracking ability is a problem in 2-step MLCE. In this paper, we propose a one-tap RLS-based channel estimation scheme for DS-CDMA with FDE. We evaluate the BER performance using one-tap RLS-based channel estimation in a frequency-selective Rayleigh fading channel by computer simulation. The BER performance using one-tap RLS-based channel estimation is compared with that using 2-step MLCE and pilot-assisted channel estimation with linear interpolation.

I. INTRODUCTION

The 4th generation (4G) mobile communication systems [1] which provide broadband wireless services of e.g. 100 Mbps to 1 Gbps are expected 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. The use of frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion can provide a better BER performance of DS-CDMA than the rake combining [3, 4].

FDE requires accurate estimation of the channel transfer function. Pilot-assisted channel estimation (CE) can be used. Time-domain pilot-assisted 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 signal is transformed into the frequency-domain pilot signal 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 is used. However, the number of Chu sequences is limited. For example, it is only 128 for the case of 256-bit period [8].

PN sequences can be used for the pilot. Using a partial sequence taken from a long PN sequence, a very large number of pilots can be generated. However, since the frequency spectrum of the partial PN sequence is not constant, the use of ZFCE produces noise enhancement [9]. The noise enhancement can be mitigated by using the minimum mean square error (MMSECE) [9]. Using MMSECE, the channel estimation accuracy is made almost insensitive to the pilot chip sequence. Recently, we proposed a 2-step maximum likelihood channel estimation (MLCE) to further improve estimation accuracy [10]. However, in the 2-step MLCE, the tracking ability is a problem because it assumes a block fading in which the channel gains stay constant over a frame. Channel estimation with RLS algorithm was proposed to track the time-varying channels [11]. In [11], RLS algorithm and the superimposed training sequences are applied for channel estimation in orthogonal frequency division multiplexing (OFDM). However, the same number of tap weights as the subcarrier block size is necessary and hence the computation complexity is very high. In this paper, we propose a one-tap RLS-based channel estimation scheme using decision feedback which is suitable for DS-CDMA with FDE. We evaluate the BER performance of DS-CDMA using one-tap RLS-based channel estimation in a frequency selective Rayleigh fading channel by computer simulation. The BER performance using one-tap RLS-based CE is compared with that using 2-step MLCE and channel estimation with interpolation.

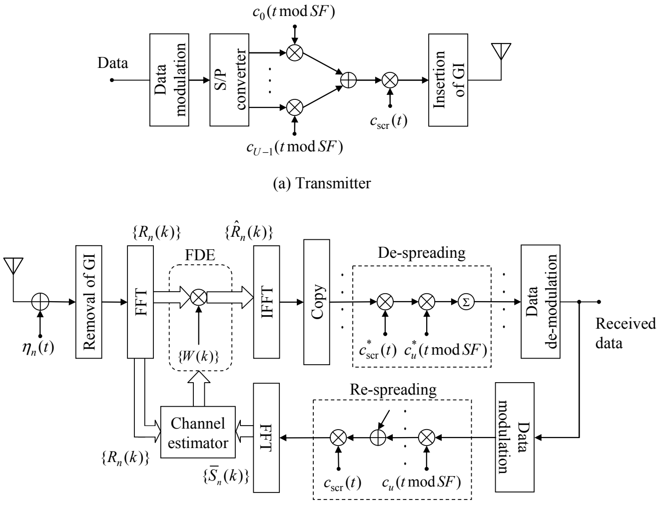
II. TRANSMISSION SYSTEM MODEL

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. The mth data symbol of the nth symbol-block (n=0~N–1) in the uth stream is represented by $d_{n,u}(m)$, m=0~N_c/SF–1, where SF is the spreading factor. $d_{n,u}(m)$ 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

N_c-point FFT into N_c frequency components and then FDE is carried out. After FDE, inverse FFT (IFFT) is applied to obtain the time-domain received chip-block for de-spreading and data-demodulation.



(a) Transmitter

(b) Receiver

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

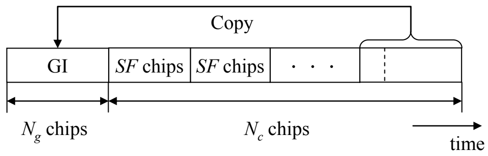


Figure 2. Chip-block structure.

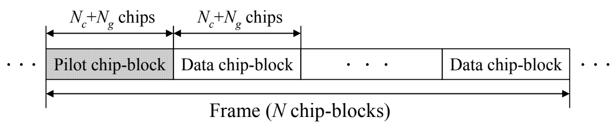


Figure 3. Frame structure.

B. Signal representation

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

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

with

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

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 change block by block.

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

$$r_n(t) = \sqrt{2P}\sum_{l=0}^{L-1} h_{n,l}\, s_n(t-\tau_l) + \eta_n(t)\,, \tag{4}$$

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. MMSE-FDE

After the removal of the GI, the received chip-block is decomposed by N_c-point FFT into N_c frequency components. The kth frequency component of the nth chip-block (n=0~N–1) 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 due to zero-mean AWGN. They are given by

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

One-tap MMSE-FDE is carried out as

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

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

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

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 at the receiver and need to be estimated. In this paper, $H_n(k)$ is estimated by RLS-based channel estimation. σ^2 can be estimated according to [9].

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

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

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

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

which is the decision variable for data demodulation on $\hat{d}_{n,u}(m)$.

III. RLS-BASED CHANNEL ESTIMATION

A. One-tap RLS algorithm

We use the following cost function for the RLS algorithm [12]:

$$\varepsilon_n(k) = \sum_{i=1}^{n} \lambda^{n-i} \mid e_i(k) \mid^2 , \tag{11}$$

where $e_i(k)$ is given by

$$e_i(k) = R_i(k) - \tilde{H}_n(k) S_i(k) , \tag{12}$$

where λ (0<λ≤1) is the forgetting factor. The channel estimate $\tilde{H}_n(k)$ is the one that minimizes $\varepsilon_n(k)$. Solving $\partial \varepsilon_n(k) / \partial \tilde{H}_n(k) = 0$ gives

$$\tilde{H}_n(k) = Z_n(k) \Phi_n^{-1}(k) , \tag{13}$$

where $Z_n(k)$ and $\Phi_n(k)$ are respectively given by

$$\begin{cases} Z_n(k) = \sum_{i=1}^{n} \lambda^{n-i} R_i(k) S_i^*(k) \\ \Phi_n(k) = \sum_{i=1}^{n} \lambda^{n-i} \left| S_i(k) \right|^2 \end{cases} . \tag{14}$$

To update $Z_n(k)$ and $\Phi_n(k)$ recursively, they are rewritten as

$$\begin{cases} Z_n(k) = \lambda Z_{n-1}(k) + R_n(k) S_n^*(k) \\ \Phi_n(k) = \lambda \Phi_{n-1}(k) + \left| S_n(k) \right|^2 \end{cases} . \tag{15}$$

$\Phi_n^{-1}(k)$ is given by [12]

$$\Phi_n^{-1}(k) = \lambda^{-1} \left(\Phi_{n-1}^{-1}(k) - G_n(k) \Phi_{n-1}^{-1}(k) S_n(k) \right), \tag{16}$$

where

$$G_n(k) = \frac{\lambda^{-1} \Phi_{n-1}^{-1}(k) S_n^*(k)}{1 + \lambda^{-1} \Phi_{n-1}^{-1}(k) \left| S_n(k) \right|^2} . \tag{17}$$

Substituting Eqs. (15), (16) and (17) into (13) gives the following update equation

$$\tilde{H}_n(k) = \tilde{H}_{n-1}(k) + G_n(k) \xi_n(k) , \tag{18}$$

where

$$\xi_n(k) = R_n(k) - \tilde{H}_{n-1}(k) S_n(k) . \tag{19}$$

B. Optimization of forgetting factor

The optimum forgetting factor λ changes according to the change in the channel condition. In this paper, λ is updated by using LMS algorithm. The following cost function is used for LMS:

$$J_n(k) = \frac{1}{2} E\left[\left| \xi_n(k) \right|^2 \right]. \tag{20}$$

Finding λ that minimizes Eq. (20) corresponds to the steepest descent method as

$$\lambda_n(k) = \lambda_{n-1}(k) + \mu \left[-\nabla \lambda_n(k) \right]. \tag{21}$$

where μ is the step size and the gradient vector $\nabla \lambda_n(k)$ is the differentiation of the cost function $J_n(k)$ with respect to λ. $\nabla \lambda_n(k)$ is given as

$$\nabla \lambda_n(k) = \frac{\partial J_n(k)}{\partial \lambda} = E\left[-\mathrm{Re}[\Psi_{n-1}(k) S_n(k) \xi_n^*(k)] \right], \tag{22}$$

where

$$\Psi_n(k) = \frac{\partial \tilde{H}_n(k)}{\partial \lambda} . \tag{23}$$

Substituting Eqs. (18) and (19) into (23) gives

$$\Psi_n(k) = \left(1 - G_n(k) S_n(k) \right) \Psi_{n-1}(k) + \frac{\partial \Phi_n^{-1}(k)}{\partial \lambda} S_n^*(k) \xi_n(k) , \tag{24}$$

$\partial \Phi_n^{-1}(k) / \partial \lambda$ in Eq. (24) can be computed recursively as follows. Differentiation of Eq. (16) with respect to λ provides the updating equation for $\partial \Phi_n^{-1}(k) / \partial \lambda$ as

$$\frac{\partial \Phi_n^{-1}(k)}{\partial \lambda} = \lambda^{-1} \begin{bmatrix} \left(1 - G_n(k) S_n(k) \right) \left(1 - G_n^*(k) S_n^*(k) \right) \\ \times \frac{\partial \Phi_{n-1}^{-1}(k)}{\partial \lambda} + \left| G_n(k) \right|^2 - \Phi_n^{-1}(k) \end{bmatrix} . \tag{25}$$

If it were possible to make exact measurements of the gradient vector $\nabla \lambda_n(k)$ at each iteration n, then the forgetting factor computed by using the steepest-decent algorithm would converge to the optimum solution. However, exact measurements of the gradient vector are not possible. Consequently, the gradient vector must be estimated from the available data when we operate in an unknown environment. The instantaneous estimate of the gradient vector $\nabla \lambda_n(k)$ on the basis of Eq. (22) is

$$\hat{\nabla} \lambda_n(k) = -\mathrm{Re}\left[\Psi_{n-1}(k) S_n(k) \xi_n^*(k) \right] . \tag{26}$$

Replacing the gradient vector $\nabla \lambda_n(k)$ in Eq. (21) by Eq. (26), we get the following recursive relation for updating the forgetting factor:

$$\begin{aligned} \lambda_n(k) &= \lambda_{n-1}(k) + \mu \left[-\hat{\nabla} \lambda_n(k) \right] \\ &= \lambda_{n-1}(k) + \mu \,\mathrm{Re}\left[\Psi_{n-1}(k) S_n(k) \xi_n^*(k) \right] \end{aligned} . \tag{27}$$

In this paper, the forgetting factor of RLS algorithm is updated with the above-described LMS algorithm.

C. Introduction of decision feedback

RLS-based CE of the nth (n≥1) block needs the transmitted chip-block replica. It is carried out as follows. After FDE of nth block, { $\hat{R}_n(k)$; k=0~N_c−1} is transformed by N_c-point IFFT into the time-domain chip-block, followed by de-spreading and tentative data symbol decision.

The tentatively detected symbol sequence { $\hat{d}_{n,u}(m)$; m=0~N_c/SF−1}, u=0~U−1, is spread to obtain the transmitted chip-block replica { $\hat{s}_n(t)$; t=0~N_c−1}:

$$\hat{s}_n(t) = \left[\sum_{u=0}^{U-1} \hat{d}_{n,u} \left(\left\lfloor \frac{t}{SF} \right\rfloor \right) c_u(t \bmod SF) \right] c_{\mathrm{scr}}(t) . \tag{28}$$

Applying N_c-point FFT to $\{\hat{s}_n(t)\}$, the kth frequency component of the transmitted chip-block replica is obtained as

$$\hat{S}_n(k) = \sum_{t=0}^{N_c-1} \hat{s}_n(t) \exp\left(-j 2\pi k \frac{t}{N_c} \right) . \tag{29}$$

For the channel estimation of the nth (n≥1) block, $G_n(k)$ and $\overline{\xi}_n(k)$ are respectively computed, using Eqs. (17) and (19), as

$$\begin{cases} G_n(k) = \dfrac{\lambda_{n-1}^{-1}(k)\Phi_{n-1}^{-1}(k)\hat{S}_n^*(k)}{1+\lambda_{n-1}^{-1}(k)\Phi_{n-1}^{-1}(k)\left|\hat{S}_n(k)\right|^2} & (30) \\ \overline{\xi}_n(k) = R_n(k) - \overline{H}_{n-1}(k)\hat{S}_n(k) \quad , & (31) \end{cases}$$

where $\overline{H}_{n-1}(k)$ is the channel estimate obtained in the (n−1)th block. Using $G_n(k)$ and $\overline{\xi}_n(k)$, the channel estimate $\widetilde{H}_n(k)$ is obtained, using Eq. (18), as

$$\widetilde{H}_n(k) = \overline{H}_{n-1}(k) + G_n(k)\overline{\xi}_n(k)\,. \tag{32}$$

The instantaneous channel gain estimate $\{\widetilde{H}_n(k)$; k=0~N_c−1} is noisy. The noise can be suppressed by applying delay time-domain windowing technique [13, 14]. $\{\widetilde{H}_n(k)$; k=0~N_c−1} is transformed by N_c-point IFFT into the instantaneous channel impulse response $\{\widetilde{h}_n(\tau)$; τ=0~N_c−1} as

$$\widetilde{h}_n(\tau) = \frac{1}{N_c}\sum_{k=0}^{N_c-1}\widetilde{H}_n(k)\exp\left(j2\pi\tau\frac{k}{N_c}\right). \tag{33}$$

The actual channel impulse response is present only within the GI length, while the noise is spread over an entire delay-time range. Replacing $\widetilde{h}_n(\tau)$ with zero's for N_g≤τ≤N_c−1 and applying N_c-point FFT, the improved channel gain estimate $\{\overline{H}_n(k)$; k=0~N_c−1} is obtained as

$$\overline{H}_n(k) = \sum_{\tau=0}^{N_g-1}\widetilde{h}_n(\tau)\exp\left(-j2\pi k\frac{\tau}{N_c}\right). \tag{34}$$

The MMSE-FDE weight of the (n+1)th block is computed using Eq.(8) by replacing $H(k)$ by $\overline{H}_n(k)$.

For the channel estimation of (n+1)th block, $\Phi_n^{-1}(k)$ is updated as

$$\Phi_n^{-1}(k) = \lambda_{n-1}^{-1}(k)\left(\Phi_{n-1}^{-1}(k) - G_n(k)\Phi_{n-1}^{-1}(k)\hat{S}_n(k)\right). \tag{35}$$

Forgetting factor is also updated as

$$\lambda_n(k) = \lambda_{n-1}(k) + \mu\,\mathrm{Re}\left[\Psi_{n-1}(k)\hat{S}_n(k)\overline{\xi}_n^*(k)\right] \,. \tag{36}$$

Finally, for updating the forgetting factor to be used for the reception of the next block, $\partial\Phi_n^{-1}(k)/\partial\lambda$ is computed as

$$\frac{\partial\Phi_n^{-1}(k)}{\partial\lambda} = \lambda_n^{-1}(k)\begin{bmatrix}\left(1-G_n(k)\hat{S}_n(k)\right)\left(1-G_n^*(k)\hat{S}_n^*(k)\right) \\ \times\dfrac{\partial\Phi_{n-1}^{-1}(k)}{\partial\lambda} + \left|G_n(k)\right|^2 - \Phi_n^{-1}(k)\end{bmatrix}. \tag{37}$$

Using $\partial\Phi_n^{-1}(k)/\partial\lambda$, $\Psi_n(k)$ is computed as

$$\Psi_n(k) = \left(1-G_n(k)\hat{S}_n(k)\right)\Psi_{n-1}(k) + \frac{\partial\Phi_n^{-1}(k)}{\partial\lambda}\hat{S}_n^*(k)\overline{\xi}_n(k)\,. \tag{38}$$

For the reception of the 0th block, replacing $\hat{S}_n(k)$ for Eqs. (30), (31), (35), (36), (37) and (38) by the pilot chip-block $\{C(k)$; k=0~N_c−1}, the channel is estimated and forgetting factor is updated. $C(k)$ is the kth frequency component of the pilot chip-block $\{\sqrt{U}c(t)$; t=0~N_c−1} with $|c(t)|$=1 (the pilot power is set to UP to keep it the same as the U-order code-multiplexed data chip-block power) and is given by

$$C(k) = \sqrt{U}\sum_{t=0}^{N_c-1}c(t)\exp\left(-j2\pi k\frac{t}{N_c}\right). \tag{39}$$

IV. 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=16 and an L=16-path frequency-selective Rayleigh fading channel having uniform power delay profile.

Fig. 4 shows the impact of fading rate on the achievable BER as a function of the normalized Doppler frequency f_DT, where T (=(N_c+N_g)T_c) is the chip-block length, at E_b/N_0=24 dB for the full code-multiplexing case (U=SF=16). It is seen from Fig. 4 that the forgetting factor λ can be optimally adjusted by LMS algorithm over all f_DT's.

The simulated BER performance of multicode DS-CDMA with MMSE-FDE is plotted in Fig. 5 for U=1 and 16 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)). The BER performances using 2-step MLCE, pilot-assisted decision feedback MMSECE, and ideal CE are also plotted for comparison. It is seen from Fig. 5 that RLS-based CE provides a better BER performance than 2-step MLCE and decision feedback MMSECE. With pilot-assisted decision feedback MMSECE, the E_b/N_0 loss from the ideal CE case for BER=10^{-4} is about 1.9 (2.4) dB when U=1 (16). This E_b/N_0 loss includes a pilot insertion loss of 0.28 dB. The use of RLS-based CE reduces the E_b/N_0 loss to about 1.2 dB for both U=1 and 16.

Fig. 6 shows the impact of fading rate on the achievable BER at E_b/N_0=24 dB as a function of the normalized Doppler frequency f_DT for the full code-multiplexing case (U=SF=16). Fig. 6 shows the results for N=16 and 64. For comparison, the BER performance using 2-step MLCE, decision feedback MMSECE and MMSECE with linear interpolation are also plotted. For RLS-based CE, the BER performance with perfect feedback (i.e., $\hat{S}_n(k) = S_n(k)$) is also plotted. It is seen from Fig. 6 that RLS-based CE provides a better BER performance than 2-step MLCE and decision feedback MMSECE. When N=16 (Fig. 6 (a)), for a very low fading rate (f_DT≤10^{-3}), RLS-based CE provides a better BER performance than MMSECE with interpolation. However, for a high fading rate (f_DT>10^{-3}), RLS-based CE provides a inferior BER performance to MMSECE with interpolation. On the other hand, when N=64 (Fig. 6 (b)), RLS-based CE provides a better BER performance than MMSECE with interpolation over a wide range of f_DT. RLS-based CE can provide a good BER performance even for a long pilot interval.

V. CONCLUSIONS

In this paper, we proposed a one-tap RLS-based CE for multicode DS-CDMA with MMSE-FDE. It was shown by computer simulation that the proposed RLS-based CE improves the BER performance compared to 2-step MLCE and decision feedback MMSECE over all f_DT's. When N=16, for a very low fading rate, RLS-based CE provides a better BER performance than MMSECE with interpolation. However, for a high fading rate, RLS-based CE provides a inferior BER performance to MMSECE with interpolation. On the other hand, when N=64, RLS-based CE provides a better BER performance than MMSECE with interpolation over a wide range of $f_D T$.

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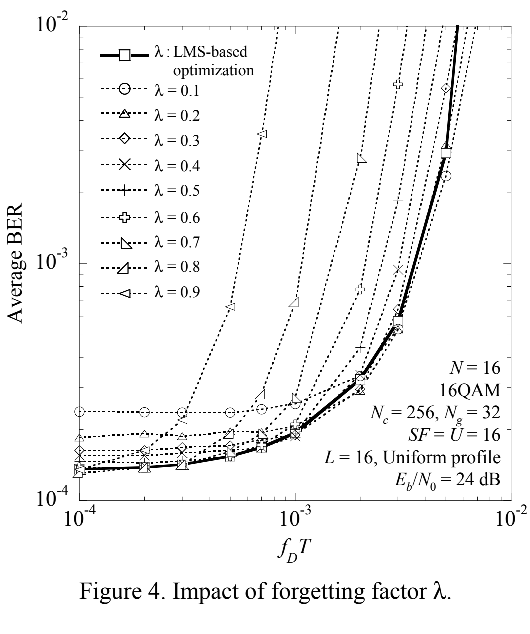


Figure 4. Impact of forgetting factor λ.

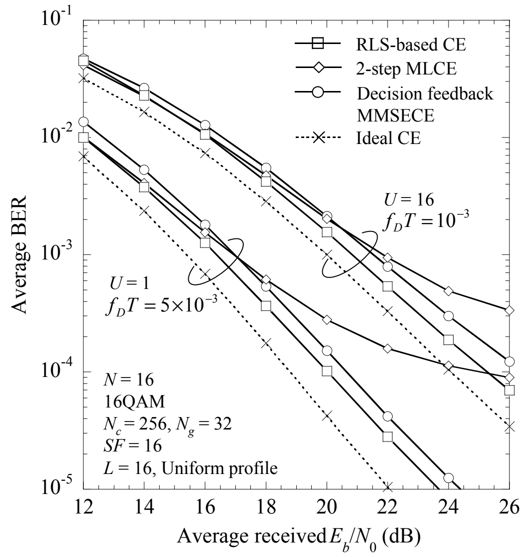
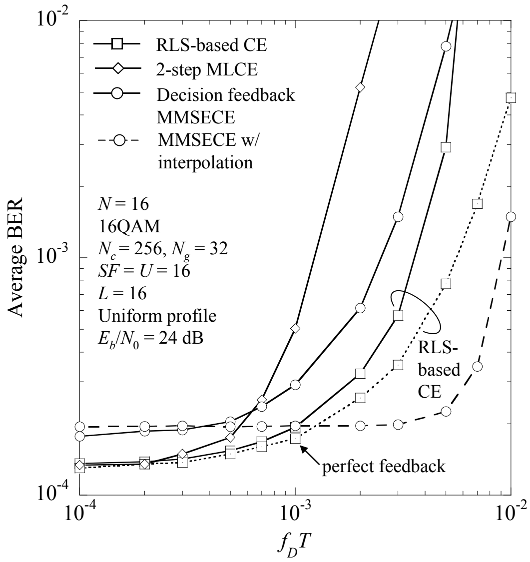
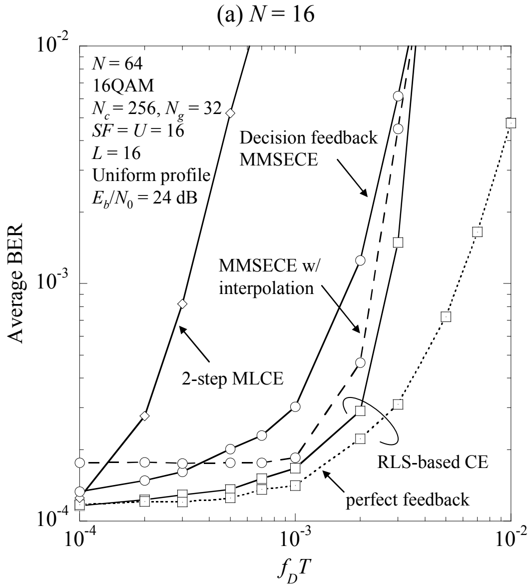


Figure 5. BER performance comparison.



(a) $N = 16$



(b) $N = 64$

Figure 6. Impact of fading rate.