Adaptive Decision Feedback Channel Estimation with Periodically Phase Correction for DS-CDMA Signal Transmission with Coherent Frequency-domain Equalization

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***Abstract*—DS-CDMA can achieve good bit error rate (BER) performance in a severe frequency-selective fading channel by using frequency-domain equalization (FDE) based on MMSE criterion instead of rake combining. FDE requires accurate channel estimation (CE). Recently, a decision feedback (DF)-MMSE-CE using a fixed DF filter coefficient has been proposed. The DF filter coefficient is an important design parameter and the optimum coefficient depends on the received signal-to-noise power ratio (SNR) and the Doppler spread. In this paper, we propose an adaptive DF (ADF)-MMSE-CE, in which the DF filter coefficient is adapted to changing channel conditions based on the recursive least square (RLS) algorithm. Furthermore, the channel estimate is phase corrected upon the reception of the periodically inserted pilot chip block. It is shown by computer simulation that the ADF-MMSE-CE with periodical phase correction is very robust against the Doppler spread.**

Keywords- DS-CDMA, frequency-domain equalization, adaptive decision feedback, channel estimation.

I. INTRODUCTION

In direct sequence code division multiple access (DS-CDMA), rake combining, that separates the copies of signal that have traveled along the different paths and coherently combining them, is applied to improve the BER performance due to path diversity effect. However, as the number of resolvable propagation paths increases, the complexity of rake combining increases and furthermore, the BER performance degrades due to severe inter-path interference (IPI). Recently, DS-CDMA with frequency-domain equalization (FDE) has been gaining much attention instead of rake combing [1]-[3]. For FDE, accurate estimation of channel transfer function is necessary. To perform channel estimation (CE) in such a fast fading channel, pilot-assisted CE [4] can be applied; the channel estimation accuracy improves by increasing the number of pilot symbols at the price of transmission efficiency. Recently, pilot-assisted CE based on the minimum mean square error (MMSE) criterion was proposed [5].

To improve the channel estimation accuracy while reducing the transmission efficiency loss due to pilot insertion, the decision feedback (DF) using the infinite impulse response (IIR) filter can be incorporated into MMSE-CE. The DF filter coefficient is an important design parameter for DF-MMSE-CE. The optimum filter coefficient depends on the signal-to-noise power ratio (SNR) and the Doppler spread. However, the SNR and the Doppler spread vary according to changes in the MS's location and traveling speed, respectively. In this paper, an adaptive DF (ADF)-MMSE-CE is proposed for the MMSE-FDE reception of DS-CDMA signals. The average BER performance of DS-CDMA with proposed ADF-MMSE-CE is evaluated by computer simulation in a frequency-selective Rayleigh fading channel.

II. DS-CDMA SIGNAL TRANSMISSION WITH FDE

A. Transmission system

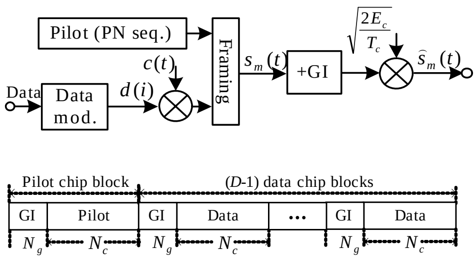
The DS-CDMA transmission system model is illustrated in Figure 1.

At the transmitter shown in Figure 1(a), the data-modulated symbol sequence $\{d(i)\}$ is spread by the spreading sequence $c(t)$ with $|d(i)|=|c(t)|=1$. Then, the resultant chip sequence $s(t)$ is divided into a sequence of N_c-chip blocks. A known pilot chip block of N_c chips is inserted every (D-1) data chip blocks; one pilot chip block and succeeding (D-1) data chip blocks make a frame. The mth block chip sequence is denoted by $s_m(t)$, t=0~(N_c-1), and the m=(nD)th chip block is the pilot chip block of the nth frame. The last N_g chips of each block is copied as a cyclic prefix and inserted into the guard interval (GI) at the beginning of each block. GI-inserted spread signal $\hat{s}_m(t)$ is expressed as

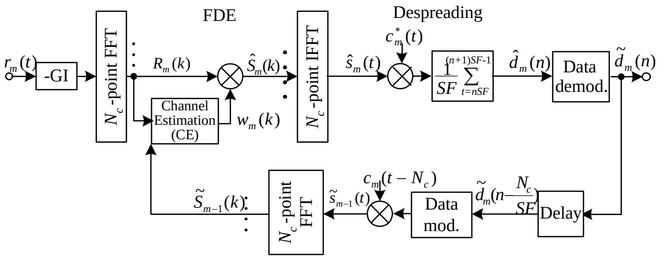
$$\hat{s}_m(t) = \sqrt{2E_c/T_c}\, s_m(t)\,, \qquad (1)$$

where E_c is the chip energy and T_c is the chip duration.

At the receiver shown in the Figure 1(b), the received signal is sampled at the chip rate to obtain



(a) Transmitter



(b) Receiver

Fig. 1 Transmitter and receiver structure.

$$r_m(t)=\sum_{l=0}^{L-1}h_{m,l}\hat{s}_m(t-\tau_l)+\eta_m(t)\,, \tag{2}$$

where $\eta_m(t)$ is the additive white Gaussian noise (AWGN) process. After removal of GI, N_c-point FFT is applied to decompose the received mth block signal sample sequence $r_m(t)$ into N_c frequency components. The kth frequency component can be expressed as

$$R_m(k)=\sqrt{\frac{2E_c}{T_c}}S_m(k)H_m(k)+\Pi_m(k)\,, \tag{3}$$

where $S_m(k)$ is the kth frequency component of $s_m(t)$ with $E[|S_m(k)|^2]=1$ and $H_m(k)$ is the channel gain with $E[|H_m(k)|^2]=1$. $\Pi_m(k)$ is the noise component, due to the AWGN, with zero-mean and the variance of $E[|\Pi_m(k)|^2]=2N_0/T_c$ with N_0 being the power spectrum density of the AWGN. They are given by

$$\begin{cases} S_m(k)=\dfrac{1}{\sqrt{N_c}}\displaystyle\sum_{t=0}^{N_c-1}s_m(t)\exp(-j2\pi k\frac{t}{N_c}) \\ H_m(k)=\displaystyle\sum_{l=0}^{L-1}h_{m,l}\exp(-j2\pi k\frac{\tau_l}{N_c}) \\ \Pi_m(k)=\dfrac{1}{\sqrt{N_c}}\displaystyle\sum_{t=0}^{N_c-1}\eta_m(t)\exp(-j2\pi k\frac{t}{N_c}) \end{cases}. \tag{4}$$

Then, one-tap MMSE-FDE is done on each frequency component as

$$\hat{S}_m(k)=w_m(k)R_m(k)\,, \tag{5}$$

where $w_m(k)$ is the MMSE-FDE weight, given by [2], [3]

$$w_m(k)=\frac{\sqrt{2E_c/T_c}H_m^*(k)}{\left|\sqrt{2E_c/T_c}H_m(k)\right|^2+2N_0/T_c}. \tag{6}$$

After FDE, N_c-point IFFT is applied to obtain the time-domain chip sequence $\hat{s}_m(t)$. Despreading is carried out to obtain the decision variable $\hat{d}_m(i)$. Finally, data-demodulation is performed to obtain $\tilde{d}_m(i)$.

B. Conventional DF-MMSE-CE

$\sqrt{2E_c/T_c}H_m(k)$ in Eq. (6) is unknown to the receiver and needs to be estimated. Since the frequency spectrum of the pilot chip sequence is not necessarily constant over the spreading bandwidth, frequency-domain MMSE-CE [5] is considered in this paper.

MMSE-CE uses the pilot chip block at $m=nD$ with n=0,1… and the channel estimate is given by [5]

$$\hat{H}_{m=nD}(k)=R_m(k)\frac{P^*(k)}{|P(k)|^2+(E_c/N_0)^{-1}}\,, \tag{7}$$

where $P(k)$ is the kth frequency component of the pilot chip sequence $p(t)$, given by

$$P(k)=\frac{1}{\sqrt{N_c}}\sum_{t=0}^{N_c-1}p(t)\exp(-j2\pi k\frac{t}{N_c}) \tag{8}$$

with $E[|P(k)|^2]=1$. The channel estimate $\hat{H}_m(k)$ is perturbed by the noise due to the AWGN. To improve the estimation accuracy, the delay-time domain windowing technique [8], [9] is applied. N_c-point IFFT is first applied to $\hat{H}_m(k)$ to obtain the instantaneous channel impulse response $\hat{h}_m(\tau)$. Since the actual channel impulse response is assumed to be present only within the GI length while the noise is distributed over the entire frequency range (i.e., k=0~N_c-1), the noise can be suppressed by setting $\hat{h}_m(\tau)=0$ (or zero-padding) beyond GI. Then, after applying N_c-point FFT, the improved estimate $\tilde{H}_m(k)$ at $m=nD$ is obtained as

$$\tilde{H}_{m=nD}(k)=\frac{1}{N_c}\sum_{\tau=0}^{N_g-1}\hat{h}_m(\tau)\exp\left(-j2\pi k\frac{\tau}{N_c}\right). \tag{9}$$

As the fading becomes faster, the pilot-assisted MMSE-CE tends to loose the tracking ability against fading. In order to improve the tracking ability, decision feedback is introduced. Pilot chip block is followed by (D-1) data chip blocks in each frame. The (m-1)th block decision $\tilde{d}_{m-1}(n)$ is fedback as a pilot. As shown in Figure 1(b), re-spreading of $\tilde{d}_{m-1}(n)$ is performed to generate the replica $\tilde{s}_{m-1}(t)$ of the transmitted (m-1)th block chip sequence. N_c-point FFT is applied to decompose $\tilde{s}_{m-1}(t)$ into N_c frequency components {$\tilde{S}_{m-1}(k)$; k=0~(N_c-1)}. Replacing $P(k)$ in Eq. (7) by $\tilde{S}_{m-1}(k)$, the DF-MMSE-CE is carried out as follows [5].

$$\hat{H}_{m-1}(k)=R_{m-1}(k)\frac{\tilde{S}_{m-1}^*(k)}{|\tilde{S}_{m-1}(k)|^2+(E_c/N_0)^{-1}}. \tag{10}$$

Also, delay time-domain windowing is performed to get the noise-reduced channel estimate $\tilde{H}_{m-1}(k)$. The first-order infinite impulse response (IIR) filter [6], [7] can be used to further improve the channel estimate as

$$\overline{H}_m(k)=\alpha\overline{H}_{m-1}(k)+(1-\alpha)\tilde{H}_{m-1}(k) \qquad (11)$$

with $m\geq 1$ and $\overline{H}_0(k)=\tilde{H}_0(k)$, where α is the IIR filter coefficient. The MMSE-FDE weight $w_m(k)$ is computed using $\overline{H}_m(k)$ instead of $\sqrt{2E_c/T_c}H_m(k)$ in Eq. (6) to perform MMSE-FDE at the mth block.

α is an important design parameter to trade off between the noise reduction and the tracking ability against fading. In a fast fading channel, α should be smaller to achieve better tracking ability. When α=0, the IIR filter output $\overline{H}_m(k)$ in Eq. (11) becomes $\tilde{H}_{m-1}(k)$; a high tracking ability against fading can be achieved, but the noise reduction is insufficient. Approximately, $\overline{H}_m(k)$ is equivalent to the average of 1/(1-α) past channel estimates, i.e., $\tilde{H}_{m-l}(k)$, l=1~$\lfloor 1/(1-\alpha)\rfloor$. Therefore, as the value of α increases, the noise can be more effectively reduced, but the tracking ability against fading tends to be lost. Therefore, there exists an optimum value in α, which depends on the received SNR and the Doppler spread [5].

III. Adaptive Decision Feedback (ADF) Filter

The fading channel statistics may vary from time to time due to the movement of the mobile terminal. Accordingly, α should be adjusted. In this paper, we apply the recursive least-square (RLS) algorithm [7]. Eq. (11) can be regarded as the single-tap least mean square (LMS) algorithm [7] with step-size parameter α. In the LMS algorithm, α is defined over the positive range of [0,1], but here we allow its value to be negative to adaptively track statistical variations of the channel gains.

α is updated so that the following exponentially weighted error Δ is minimized, which is defined as [6]

$$\Delta=\sum_{l=1}^{m-1}\beta^l\left|R_{m-l}(k)-\overline{H}_{m-l}(k)\tilde{S}_{m-l}(k)\right|^2, \qquad (12)$$

where the forgetting factor β is a positive constant close to, but less than, unity. The inverse of 1-β is, roughly speaking, a measure of the memory size of the algorithm [7]. Here, α in Eq. (11) is denoted by $\alpha_m(k)$. After some manipulations, the optimum value of $\alpha_m(k)$ that minimizes Δ is found to be

$$\alpha_m(k)=-\frac{\mathrm{Re}\left\{\sum_{l=1}^{m-1}\beta^l A_{m-l}(k)B^*_{m-l}(k)\right\}}{\sum_{l=1}^{m-1}\beta^l\left|B_{m-l}(k)\right|^2}, \qquad (13)$$

where

$$\begin{cases}A_{m-l}(k)=R_{m-l}(k)-\tilde{H}_{m-1-l}(k)\tilde{S}_{m-l}(k)\\ B_{m-l}(k)=\tilde{S}_{m-l}(k)\left\{\tilde{H}_{m-1-l}(k)-\overline{H}_{m-2-l}(k)\right\}\end{cases}. \qquad (14)$$

From the above equations, the following recursive adaptation algorithm is formulated:

$$\alpha_m(k)=\begin{cases}0, & m=1\\ -\dfrac{\Theta_m(k)}{\Omega_m(k)}, & m>1\end{cases}, \qquad (15)$$

where

$$\begin{cases}\Theta_m(k)=\beta\Theta_{m-1}(k)+\mathrm{Re}\left\{A_m(k)B^*_m(k)\right\}\\ \Omega_m(k)=\beta\Omega_{m-1}(k)+\left|B_m(k)\right|^2\end{cases} \qquad (16)$$

for k=0~(N_c-1). The initial condition is set as Θ_0=0 and $\Omega_0=\delta$ (small positive value). Upon the reception of pilot chip block at $m=nD$, n=0,1…, the output of IIR filter $\overline{H}_m(k)$ is phase rotated, so that the resultant channel estimate becomes in-phase with the pilot-assisted channel estimate $\tilde{H}_{m=nD}(k)$ obtained by the pilot-assisted MMSE-CE, see Eq. (9).

$\alpha_m(k)$ is perturbed by the noise and hence is not reliable. Remember that the fading statistics are identical at all frequencies. We use the filter coefficient α_m averaged over all the frequencies:

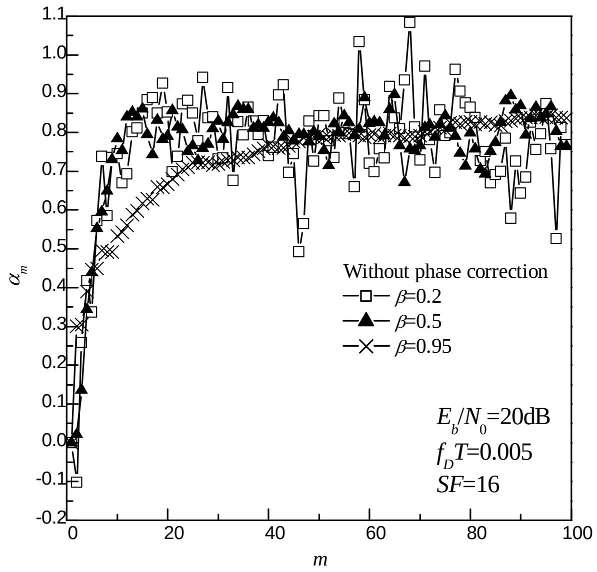
$$\alpha_m=(1/N_c)\sum_{k=0}^{N_c-1}\alpha_m(k). \qquad (17)$$

IV. Computer Simulation

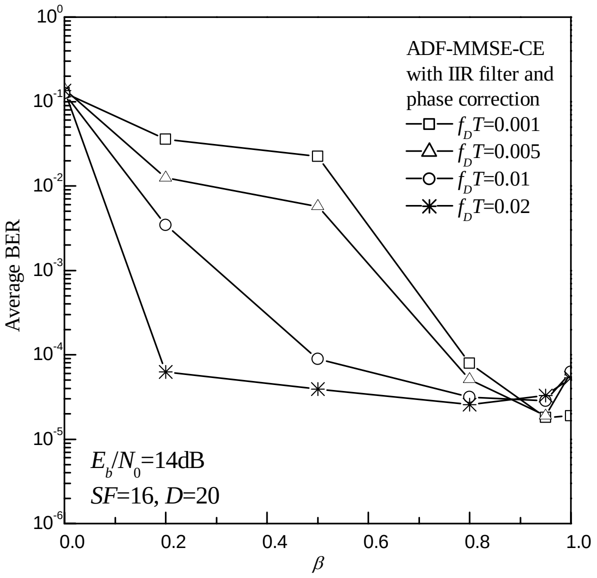
We assume N_c=256, N_g=32 and an L=16-path Rayleigh fading channel having the uniform power delay profile and the maximum Doppler frequency f_D. It is assumed that the lth path time delay τ_l is $\tau_l=l$ chips and the maximum delay difference is less than the GI length.

The impact of forgetting factor β on the convergence rate of α_m is shown in Figure 2 (a) for the average received bit energy-to-AWGN power spectrum density ratio E_b/N_0=20dB, the spreading factor SF=16, and the normalized Doppler frequency f_DT=0.005, where 1/T is the symbol rate. E_b/N_0 is defined as $E_b/N_0=0.5SF(1+N_g/N_c)(E_c/N_0)(1-D^{-1})$. The initial value of α_m is set as α_0=0. As mentioned in Sect. III, the inverse of 1-β is roughly a measure of the memory size of the RLS algorithm. It can be seen from Figure 2(a) that the use of smaller β has better tracking ability, but α_m varies widely and is not stable enough. It can also be seen that the phase correction based on pilot-assisted channel estimation is necessary due to the long convergence time. Figure 2(b) shows the impact of β on the BER at E_b/N_0=14dB. It can be seen that the use of β=0.95 for the adaptive IIR filter is suitable for f_DT=0.001~0.02.

The simulated BER performance of DS-CDMA using ADF-MMSE-CE with IIR filter is plotted as a function of E_b/N_0 for SF=16 and 64 in Figure 3. Here, the frame length D is equal to 20 and f_DT varies from 0.001 to 0.01. It can be seen that ADF-MMSE-CE provides a fairly good BER performance for all f_DT values. For comparison, the simulation result using non-adaptive DF-MMSE-CE is also plotted. Using non-adaptive DF filter with α=1, the BER performance degrades significantly for large Doppler spread (i.e., f_DT=0.01). Using α=0, $\tilde{H}_{m-1}(k)$ is always used (see Eq. (11)). A good tracking ability can be achieved but noise cannot be sufficiently suppressed; therefore, the achievable BER performance is worse than ADF-MMSE-CE (β=0.95). ADF-MMSE-CE gives the best performance and is almost insensitive to f_DT. The E_b/N_0 degradation for BER=10^{-5} from the ideal CE is less than 1dB. Part of this E_b/N_0 degradation is due to the pilot insertion loss of 10log(20/19)=0.23dB.



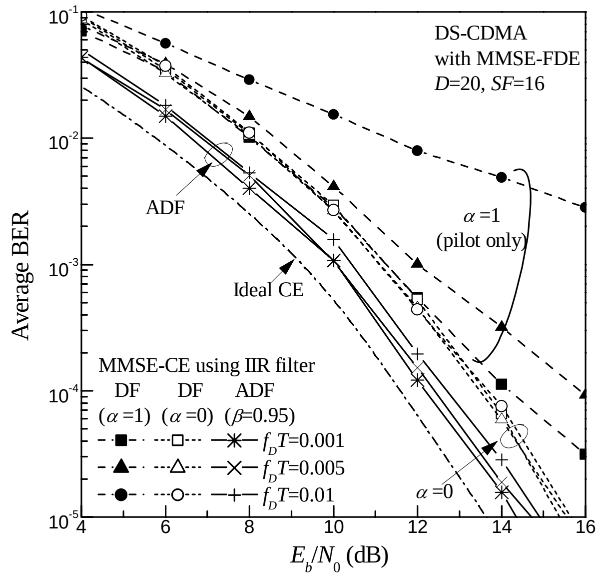
(a) Convergence rate



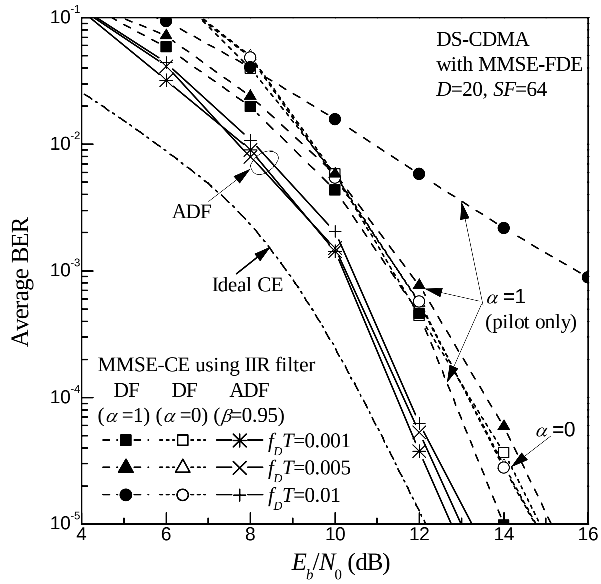
(b) BER

Figure 2. Impact of forgetting factor β on ADF-MMSE-CE with IIR filter.

Phase correction using periodically inserted pilot is used to avoid the error propagation beyond the frame. The pilot insertion loss, given by 10log$D$/(D-1) dB, can be reduced by increasing the frame length D. However, the tracking ability against fading may be lost due to decision feedback error propagation within a frame. Figure 4 shows the impact of the frame length on the BER as a function of f_DT at E_b/N_0=14dB for SF=16 and at E_b/N_0=12dB for SF=64. By increasing the value of D from 20 to 100, the pilot insertion loss can be reduced from 0.23dB to 0.04dB and the BER performance is



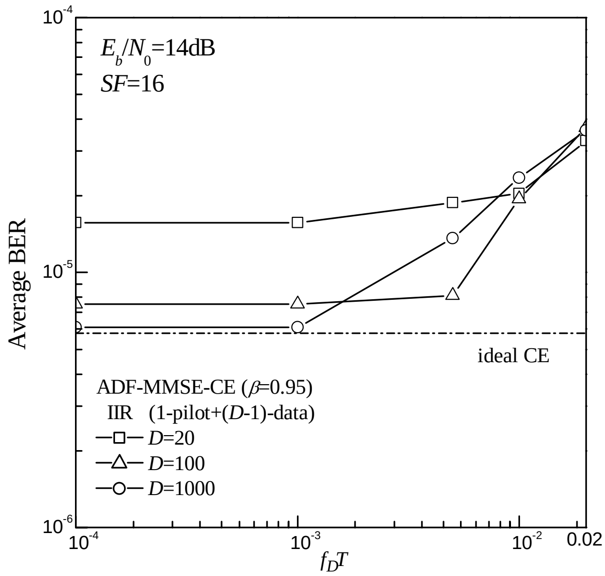
(a) SF=16



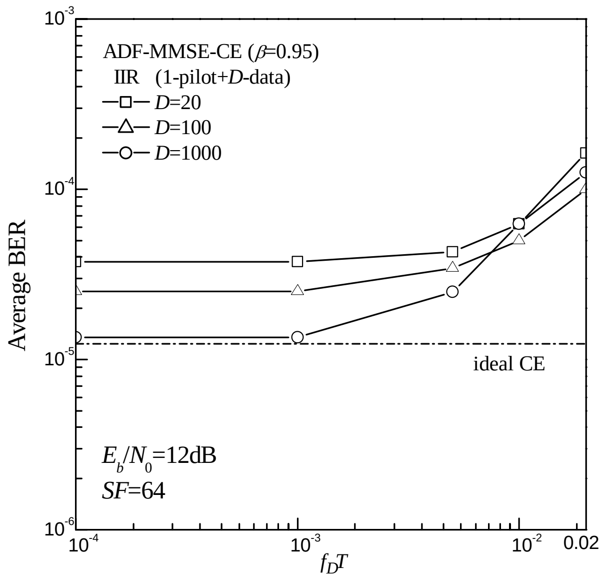
(b) SF=64

Figure 3. Performance comparison of DF- and ADF-MMSE-CE.

slightly improved. Although phase correction using periodically inserted pilot is used, the decision feedback error propagates within a frame for large Doppler spread and degrades the BER performance. Therefore, the frame length cannot be too large, e.g., D=1000. However, when D=100 is used, the achievable BER is almost the same if f_DT is less than 0.005 (f_DT=0.005 corresponds to a terminal moving speed of 375km/h for a chip rate of 100Mcps and 5GHz carrier frequency).



(a) *SF*=16



(b) *SF*=64

Figure 4. Impact of f_DT.

V. Conclusions

In this paper, an adaptive decision feedback MMSE channel estimation (ADF-MMSE-CE) scheme was proposed for DS-CDMA with coherent MMSE-FDE. The DF filter coefficient is adaptively updated based on the recursive least-square (RLS) algorithm to track the changing channel conditions. In order to stop the error propagation beyond the frame due to decision feedback, the channel gain estimate is phase corrected using the periodically received pilot chip block. It has been confirmed by the computer simulation that the average BER performance using ADF-MMSE-CE with periodical phase correction is highly robust against fast fading.

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