社団法人 電子情報通信学会
THE INSTITUTE OF ELECTRONICS,
INFORMATION AND COMMUNICATION ENGINEERS

信学技報
TECHNICAL REPORT OF IEICE

周波数領域差動符号化 DS-CDMA

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あらまし 周波数選択性が強い無線チャネルにおいて，周波数領域等化(FDE)DS-CDMA は MC-CDMA よりも優れた特性が得られることを最近示した．また，理論検討および計算機シミュレーションにおいて，理想チャネル推定では最小二乗誤差(MMSE)規範に基づく FDE 重みが最も優れたビット誤り率(BER)特性を得られることが分かっている．しかし，平均 BER 特性はチャネル推定誤差によって劣化してしまう．ところで移動無線通信において，遅延検波は受信側でチャネル推定が必要ないことからで実用に優れていることが知られている．そこで本論文では，周波数領域差動符号化 DS-CDMA(FDDED)を提案し，同期検波を用いる FDE-MMSE 方式と比較したとき，本提案方式が同等，もしくは優れた BER 特性が得られることを示している．

キーワード DS-CDMA，FDDED, FDE, MMSE

Frequency-domain differential encoding/decoding in single-carrier DS-CDMA

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Abstract Recently, it was shown that frequency-domain equalization (FDE) for the reception of single-carrier DS-CDMA signals is more effective than that of MC-CDMA systems in a frequency-selective fading channel. Both the theoretical analysis and computer simulation have proved that FDE based on minimum mean square error (MMSE) criterion provides the best BER performance with perfect channel information. However, the BER performances suffer from inevitable channel estimation error. For applications like mobile radio, the differential detection is preferable because of its simple implementation without any channel estimation at the receiver end. In this paper, we propose frequency-domain differential encoding and decoding (FDDED) for single-carrier DS-CDMA signal transmission in a frequency-selective fading channel. Compared with that of coherent FDE-MMSE, FDDED provides similar or even better BER performance especially for small spreading factor, SF (i.e., higher data rate for the given chip rate).

Keyword DS-CDMA，FDDED, FDE, MMSE

1. Introduction

Increasing demand of providing broadband services through a wireless network requires the future wireless communication systems to support the transmission rate up to around hundred Mbps [1]. However, frequency-selective multipath fading encountered in a broadband wireless digital communication system, severely degrades the bit error rate (BER) performance of broadband data transmissions [2,3]. In direct sequence CDMA systems (DS-CDMA), coherent Rake combining can exploit the channel frequency-selectivity to improve the BER performance through path diversity effect [4]. However, due to an excessive number of resolvable multipaths appearing in a broadband channel, the complexity of Rake receiver increases and furthermore the BER performance degrades because severe inter-path interference (IPI) offsets the path diversity effect obtained by Rake combining [4].

Multi-carrier CDMA (MC-CDMA) [5] based on the combination of orthogonal frequency division multiplexing (OFDM) and CDMA has been receiving much attention. By using multicarrier modulation, data symbol to be transmitted is spread over a number of subcarriers by which frequency-diversity effect is achieved with lower complexity of frequency-domain equalization than Rake combining [6].

Recently, single-carrier wireless transmission with frequency-domain equalization (FDE) has been gaining much popularity [7]. It has been shown that FDE is more effective than that of Rake combining for the reception of single DS-CDMA signals in frequency-selective fading channel [8]. Both the theoretical analysis and computer simulation have shown that FDE based on minimum mean square error (MMSE) criterion provides the best BER performance with ideal channel information [9]. However, so far presented are only the simulation results of FDE with the assumption of ideal channel estimation. In practical systems, the BER performances suffer from inevitable channel estimation error.

There have been several pilot-based channel estimation schemes so far proposed for FDE for the reception of MC- and DS-CDMA signals [10-12]. However, the accuracy of channel estimation is sensitive to Doppler frequency spread. For applications like mobile radio, differential detection is sometimes preferable because of not only the simple implementation due to no requirement of channel estimation, but also its robustness to the high mobility of terminals. In this paper, we propose frequency-domain differential encoding and decoding (FDDED) based on the MMSE criterion for the transmission of DS-CDMA signals and compare the achievable BER performance with that of MMSE-FDE.

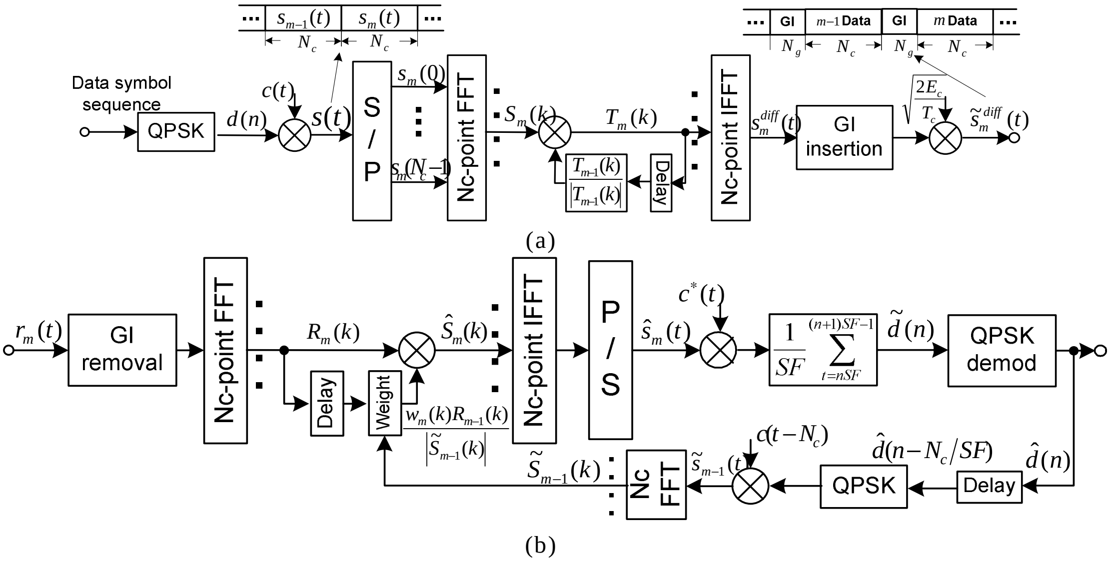


Fig.1. Transmitter and receiver

2. Transmission System Model

2.1. Transmit Signal

The data-modulated symbol sequence, $d(n)$, is spread by the user' specific spreading code, $\{c(t)\}$, with $|d(n)|=|c(t)|=1$ to obtain

$$s(t)=d(\lfloor t/SF \rfloor)c(t \bmod SF), \qquad (1)$$

where SF is the spreading factor and $\lfloor x \rfloor$ is the largest integer smaller than or equal to x. Then, the data chip sequence is then divided into blocks of N_c chips each, where N_c is the FFT window size. Each block is transformed by N_c-point FFT into N_c frequency components (hereafter we use the terminology "subcarrier" instead of frequency component) for FDDE. The k th subcarrier component of the m th chip block is given by

$$\begin{aligned} S_m(k) &= \frac{1}{\sqrt{N_c}} \sum_{t=mN_c}^{(m+1)N_c-1} s(t)\exp(-j2\pi k\frac{i}{N_c}) \\ &= \frac{1}{\sqrt{N_c}} \sum_{i=0}^{N_c-1} s_m(i)\exp(-j2\pi k\frac{i}{N_c}) \end{aligned} \qquad (2)$$

for $k=0\sim(N_c-1)$, where $s_m(i)=s(t-mN_c)$. $S_m(k)$ is a random variable with zero-mean and unity variance:

$$E\left[|S_m(k)|^2\right]=E\left[\frac{1}{N_c}\left|\sum_{i=0}^{N_c-1} s_m(i)\exp\left[-j2\pi k\frac{i}{N_c}\right]\right|^2\right]=1. \qquad (3)$$

FDDE on the k th subcarrier is performed as follows:

$$T_m(k)=S_m(k)\frac{T_{m-1}(k)}{|T_{m-1}(k)|}. \qquad (4)$$

The initial condition is set as $T_0(k)=S_p(k)$, which is known chip sequence at the receiver.

After FDDE, N_c-point IFFT is applied to obtain the time-domain DS-CDMA signal. The m th chip block of the differentially encoded DS-CDMA signal can be expressed, using the equivalent baseband representation, as:

$$s_m^{diff}(t)=\frac{1}{\sqrt{N_c}}\sum_{k=0}^{N_c-1} T_m(k)\exp(j2\pi t\frac{k}{N_c}) \qquad (5)$$

for $t=0\sim(N_c-1)$. In order to apply FFT on the received differential encoded DS-CDMA signal at the receiver, it is necessary to insert the guard interval (GI) of N_g chips before transmission [7]. In this way, the time delayed signal can be viewed as a circularly shifted version of the transmitted signal, similar to MC-CDMA. In this paper, we assume the square-root Nyquist chip shaping filter at the transmitter and the same filter at the receiver as the chip-matched filter. Also assumed is a chip-spaced L path fading channel. Ideal chip sampling time is assumed at the receiver. Therefore, the chip (T_c)-spaced discrete time signal representation is used throughout the paper. The GI-inserted chip sequence for the m th block can be expressed as

$$\tilde{s}_m^{diff}(t)=\sqrt{\frac{2E_c}{T_c}}s_m^{diff}(t) \tag{6}$$

for $t=-N_g \sim (N_c-1)$, where E_c is the average chip energy.

2.2. Channel Model

We assume a block fading channel having L independent propagation paths with T_c-spaced time delays, $\{\tau_l; l=0\sim(L-1)\}$. The assumption of the block fading means that the path gains remain constant over one block interval of N_c, but they vary block-by-block according to the terminal movement. Also the maximum time delay of the channel is assumed to be shorter than GI, i.e., $\max_l\{\tau_l\}\le N_g$. The channel impulse response during the reception of the mth block can be expressed as

$$h_m(t)=\sum_{l=0}^{L-1}h_{m,l}\delta(t-\tau_l) \tag{7}$$

where $h_{m,l}$ is the l th path gain with $E\left[\sum_{l=0}^{L}|h_{m,l}|^2\right]=1$, $\delta(x)$ is the delta function, and $E[\cdot]$ is the ensemble average operation. Consequently, the channel transfer function $H_m(k)$ can be given by

$$H_m(k)=\sum_{l=0}^{L-1}h_{m,l}\exp(-j2\pi k\frac{\tau_l}{N_c}) \tag{8}$$

with $E\left[|H_m(k)|^2\right]=1$.

2.3. Received Signal

The received signal is sampled at the chip rate to obtain

$$r_m(t)=\sum_{l=0}^{L-1}h_{m,l}\tilde{s}_m^{diff}(t-\tau_l)+\eta_m(t), \tag{9}$$

where $\eta_m(t)$ is the additive white Gaussian noise (AWGN) process with zero mean and variance $2N_0/T_c$ with N_0 being the single-sided power spectrum density. As shown in the Fig.1(b), after removal of GI, N_c-point FFT is applied to the received signal sequence $r_m(t)$ to decompose into N_c subcarrier components:

$$\begin{aligned}R_m(k)&=\frac{1}{\sqrt{N_c}}\sum_{t=0}^{N_c-1}r_m(t)\exp(-j2\pi k\frac{t}{N_c})\\&=\sqrt{\frac{2E_c}{T_c}}T_m(k)\sum_{l=0}^{L-1}h_{m,l}\exp(-j2\pi k\frac{t}{N_c})+\Pi_m(k)\\&=\sqrt{\frac{2E_c}{T_c}}T_m(k)H_m(k)+\Pi_m(k)\end{aligned} \tag{10}$$

for $k=0\sim(N_c-1)$, where $\Pi_m(k)$ is the k th subcarrier noise component with zero-mean and variance $2\sigma^2=E\left[|\Pi_m(k)|^2\right]=2N_0/T_c$.

Joint frequency-domain differential decoding (FDDD) and FDE is applied to each subcarrier. Since $S_m(k)=\{T_m(k)/T_{m-1}(k)\}|T_{m-1}(k)|=\{T_m(k)/T_{m-1}(k)\}|S_{m-1}(k)|$ from Eq.(4) and respectively replacing $T_m(k)$ and $T_{m-1}(k)$ by $R_m(k)$ and $R_{m-1}(k)$ based on the assumption of $H_m(k)\approx H_{m-1}(k)$. FDDD can be performed as follows:

$$\hat{S}_m(k)=w_m(k)\frac{R_m(k)}{R_{m-1}(k)/|S_{m-1}(k)|}, \tag{11}$$

where $w_m(k)$ is the FDE weight. Then, N_c-point IFFT is applied to obtain the time-domain chip sequence:

$$\hat{s}_m(t)=\frac{1}{\sqrt{N_c}}\sum_{k=0}^{N_c-1}\hat{S}_m(k)\exp(j2\pi t\frac{k}{N_c}). \tag{12}$$

Finally, despreading is carried out to get the decision variable for detection of $d(n)$:

$$\tilde{d}(n)=\frac{1}{SF}\sum_{t=nSF}^{(n+1)SF-1}\hat{s}_m(t)c^*(t), \tag{13}$$

based on which QPSK demodulation is performed to obtain $\hat{d}(n)$.

3. Optimal Equalization Weight

Let

$$\omega_m(k)=\frac{w_m(k)}{R_{m-1}(k)/|S_{m-1}(k)|}. \tag{14}$$

We want to find the set of FDE weights $\{\omega_m(k);k=0,\dots,N_c-1\}$ for the m th block signal according to the MMSE criterion. $\hat{S}_m(k)$ can be expressed as

$$\begin{aligned}\hat{S}_m(k)&=\omega_m(k)R_m(k)\\&=\sqrt{\frac{2E_c}{T_c}}\omega_m(k)H_m(k)S_m(k)\frac{T_{m-1}(k)}{|T_{m-1}(k)|}\\&\quad+\omega_m(k)\Pi_m(k),\end{aligned} \tag{15}$$

which is different from the transmitted signal component $S_m(k)$. We define the equalization error for the mth block signal at the kth subcarrier as

$$\begin{aligned}\varepsilon_m(k)&=S_m(k)-\hat{S}_m(k)\\&=S_m(k)\left[1-\frac{\omega_m(k)}{\sqrt{2E_c/T_c}}H_m(k)\frac{T_{m-1}(k)}{|T_{m-1}(k)|}\right]-\omega_m(k)\Pi_m(k).\end{aligned} \tag{16}$$

The mean square error $E[|\varepsilon_m(k)|^2]$ for the given set of $\{H_m(k)\ and\ H_{m-1}(k); k=0\sim(N_c-1)\}$ is given by

$$E[|\varepsilon_m(k)|^2]=\left[\begin{aligned}&1+\left|\frac{\omega_m(k)}{\sqrt{2E_c/T_c}}H_m(k)\frac{T_{m-1}(k)}{|T_{m-1}(k)|}\right|^2\\&-2\,\mathrm{Re}\{\frac{\omega_m(k)}{\sqrt{2E_c/T_c}}H_m(k)\frac{T_{m-1}(k)}{|T_{m-1}(k)|}\}\\&+\frac{2N_0}{T_c}|\omega_m(k)|^2\end{aligned}\right]. \tag{17}$$

The set of weights that satisfy $\partial E[|\varepsilon_m(k)|^2]/\partial\omega_m(k)=0$ is the optimum one in the MMSE sense. We obtain the MMSE weight for FDDD, called FDDD-MMSE:

$$\omega_m^{MMSE}(k)=\frac{\sqrt{\frac{2E_c}{T_c}}H_m^*(k)\frac{T_{m-1}^*(k)}{|T_{m-1}(k)|}}{\left|\sqrt{\frac{2E_c}{T_c}}H_m(k)\frac{T_{m-1}(k)}{|T_{m-1}(k)|}\right|^2+\sigma_\Pi^2}, \qquad (18)$$

where $\sigma_\Pi^2=2N_0/T_c$ is the noise power at each subcarrier. However, $\sqrt{2E_c/T_c}\,H_m(k)T_{m-1}(k)/|T_{m-1}(k)|$ is not known to the receiver. Assuming $H_m(k)\approx H_{m-1}(k)$, we replace $\sqrt{2E_c/T_c}H_m(k)T_{m-1}(k)$ by $R_{m-1}(k)$. From Eq.(4), we have $|T_{m-1}(k)|=|S_{m-1}(k)|$ which can be estimated by decision feedback of the previous block decision result; $|T_{m-1}(k)|$ can be replaced by $|\tilde{S}_{m-1}(k)|$. Consequently, we have the following weight for FDDD-MMSE:

$$\omega_m^{MMSE}(k)\approx\frac{R_{m-1}^*(k)/|\tilde{S}_{m-1}(k)|}{|R_{m-1}(k)/\tilde{S}_{m-1}(k)|^2+2\sigma_\Pi^2}. \qquad (19)$$

At the receiver, no channel estimation is required but only the noise power measurement. $\omega_m^{MMSE}(k)$ can be viewed as the reference signal for FDDD. However, this reference is the instantaneous received signal component and is noisy. Therefore, to improve the quality of the reference signal, we apply the following infinite impulse response (IIR) filtering combined with decision feedback:

$$\left(\frac{R'_{m-1}(k)}{|\tilde{S}_{m-1}(k)|}\right)=(1-\alpha)\frac{R_{m-1}(k)}{|\tilde{S}_{m-1}(k)|}+\alpha\left(\frac{R'_{m-2}(k)}{|\tilde{S}_{m-2}(k)|}\right)\frac{\tilde{S}_{m-1}(k)}{|\tilde{S}_{m-1}(k)|}. \qquad (20)$$

with $R'_0(k)/\tilde{S}_0(k)=S_p(k)/|S_p(k)|$, where α ($0\le\alpha\le1$) is the filter coefficient. $R_{m-1}(k)/\tilde{S}_{m-1}(k)$ in Eq.(20) is replaced by $R'_{m-1}(k)/\tilde{S}_{m-1}(k)$. α is an important design parameter to determine the tracking ability of the FDDED against fading. As $\alpha\rightarrow1$, the effect of noise reduction is improved but the tracking ability against faster fading tends to be lost; the noise reduction and fading tracking ability are in tradeoff relationship. According to the Doppler spread, the filter coefficient α can be adjusted to track the fading variation while achieving the good noise reduction. From the computer simulation, this scheme is confirmed to yield even better BER performance than that of the pilot-based coherent FDE-MMSE in the fast fading environment and achieves a good robustness against the Doppler spread.

4. Simulation Results

The computer simulation condition is shown in Table1. An L -path frequency-selective block Rayleigh fading channel having uniform power delay profile (i.e., $E[|h_{m,l}|^2]=1/L$ for all m and l) is assumed. The time delay τ_l of the l th path is assumed to be $\tau_l=l$ chips and $\max_l\{\tau_l\}\le N_g$. Comparison is done between FDDED-MMSE and coherent FDE-MMSE.

Table1. Simulation condition

Block Length	$N_c=256$
Guard interval	$N_g=32$
Data modulation	QPSK
Spreading code	Long Gold sequences
Spreading Factor	SF=1, 16, 64, 256
Equalization schemes	FDDD-MMSE

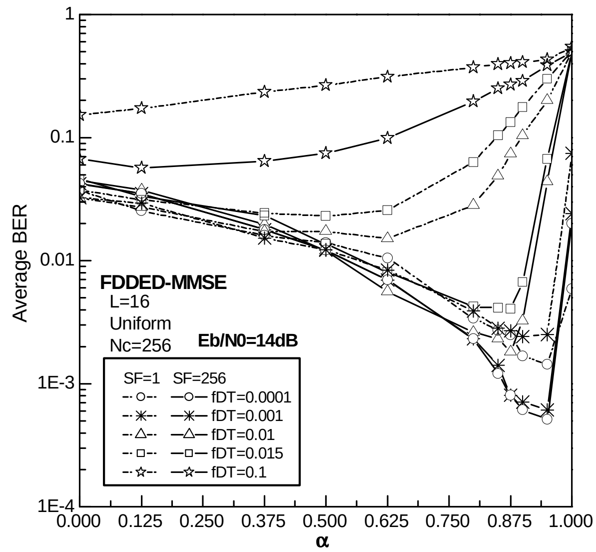


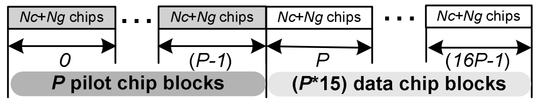
Fig.2. Effect of filter coefficient α.

As mentioned in Sec.3, the filter coefficient α should be optimized for a different channel condition. Fig.2 plots the effect of α on the BER with f_DT as a parameter for $SF=1$ and $SF=256$, where f_D is the fading maximum Doppler frequency. From Fig.2, it is shown that there exists the optimum value in α. When f_DT is small, the optimum α is close to unity because fading is slow enough and thus α can be close to unity to achieve better noise reduction. However, the optimum α becomes smaller as f_DT increases to get better tracking ability against fading variations.

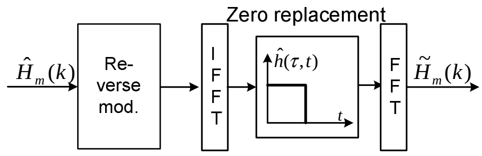
The BER performance is compared between FDDED-MMSE and coherent FDE-MMSE. A pilot-based channel estimation using delay-time domain windowing [11], [12] is employed here for coherent FDE-MMSE. For channel estimation, P pilot symbols are transmitted every P*15 data chip blocks. In Fig.3(b), the noisy estimate $\hat{H}_m(k)$ of the channel gain for the k th subcarrier is obtained by multiplying the received pilot subcarrier component $R_m(k)$ with the complex conjugate of pilot subcarrier component $P_m(k)$, i.e., $\hat{H}_m(k) = R_m(k)P_m^*(k)$, where $0 \le m < P$. Then, IFFT is applied to the noisy estimate $\hat{H}_m(k)$ for obtaining the noisy channel impulse response $\hat{h}_m(\tau)$. Since the channel impulse response $\hat{h}_m(\tau)$ is assumed to be present only within GI length, the AWGN noise over the entire range can be suppressed by replacing $\hat{h}_m(\tau)$ beyond the GI with zeros (or zero-padding). After applying FFT, the improved estimate $\tilde{H}_m(k)$ is obtained. Channel estimation can be further improved by averaging P estimates to reduce the noise effect as follows:

$$\overline{H}(k) = \frac{1}{P}\sum_{p=0}^{P-1} \tilde{H}_p(k). \quad (21)$$

The BER performance comparison between FDDED-MMSE and coherent FDE-MMSE is shown in Fig.4 for $SF = 1, 16, 64, 256$. It can be seen that the BER of FDE-MMSE using more pilot chip blocks can approach that using the ideal channel estimation when $f_D T = 0$.

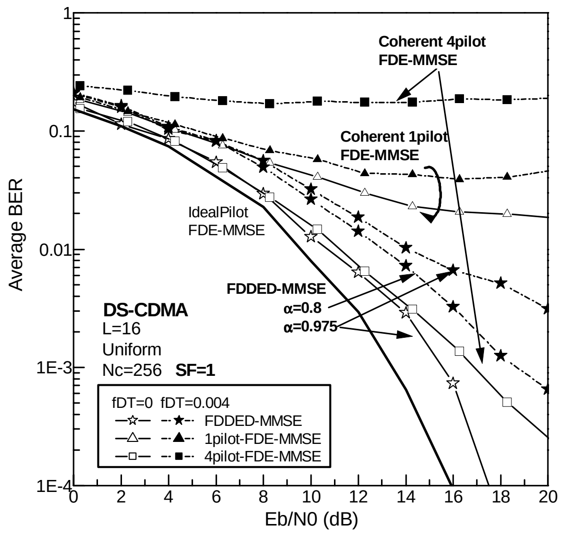


(a) Frame structure

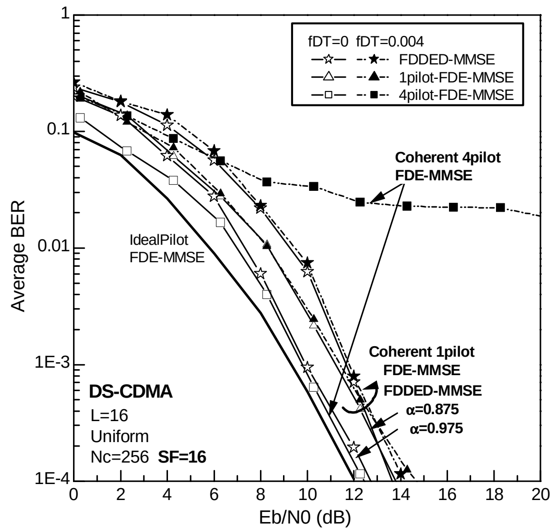


(b) Channel estimation using delay-time domain windowing

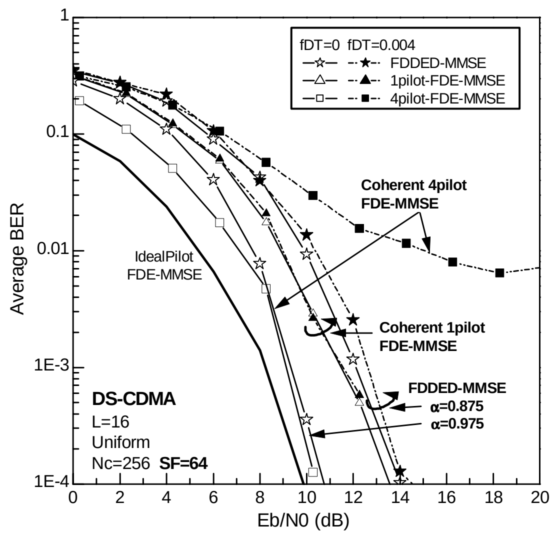
Fig.3. Pilot-based channel estimation.



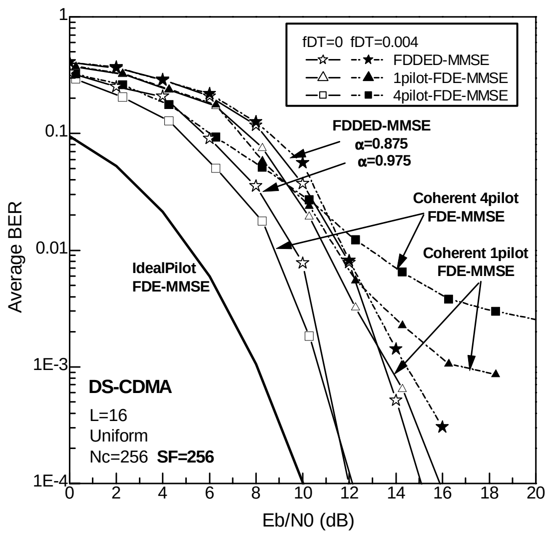
(a) SF=1



(b) SF=16



(c) SF=64



(d) SF=256

Fig.4. Comparison between FDDED-MMSE and Coherent FDE-MMSE in DS-CDMA systems with different SF

However, with the increase in Doppler spread, FDE-MMSE tends to lose the tracking ability, thereby exhibiting the significant performance degradation as the number of pilot symbols increases. 4-pilot coherent FDE is very sensitive to Doppler spread and it exhibits an error floor when $f_D T = 0.004$ (corresponding to a terminal traveling speed of 3km/h for *SF*=256 and 768km/h for *SF*=1 when the carrier frequency is 5GHz and a bit rate is 100Mbps). On the other hand, when $f_D T = 0$ our proposed FDDED-MMSE with $\alpha = 0.975$ can achieve similar performance with coherent 4pilot-based FDE-MMSE and much better performance than 1-pilot coherent FDE. Using FDDED-MMSE, channel estimation on each subcarrier required for coherent FDE is replaced by the simple noise measurement. If we set $\alpha = 0.875$, FDDED performs slightly worse but almost insensitive to Doppler spread, just similar to 1-pilot coherent FDE when $SF > 1$. For the case of *SF*=1, FDDED with $\alpha = 0.875$ can outperform 4-pilot FDE-MMSE with larger E_b/N_0 but for a large $f_D T$, α should be slightly reduced in order to get a good tracking ability against fading.

Fig.5 presents the effect of $f_D T$ on the BER performance for *SF*=8 as an example. FDDED with fixed $\alpha = 0.875$ can be robust against large Doppler spread, i.e., very high mobility environments, and provides better BER performance than pilot-based coherent FDE-MMSE.

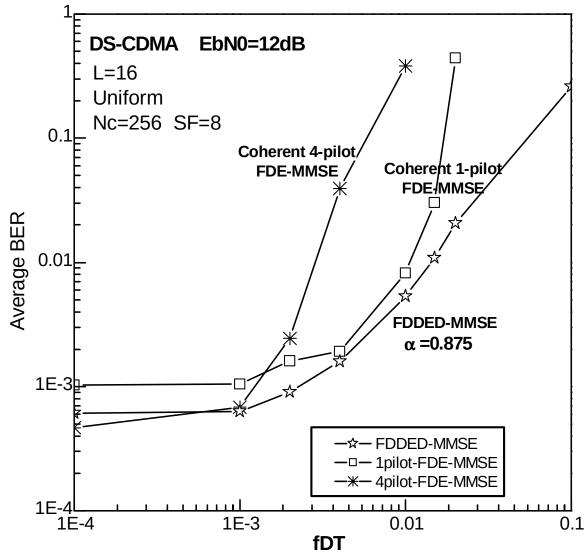


Fig.5. Effect of fDT

5. Conclusions

In this paper, we proposed frequency-domain differential encoding and decoding (FDDED) for single-carrier DS-CDMA signal transmission in a frequency-selective fading channel with high Doppler frequency spreads. Suboptimum equalization weight was theoretically derived and average BER performance was evaluated by computer simulation. It was confirmed by simulation results that the proposed FDDED-MMSE is very robust against the Doppler spread and outperforms coherent FDE-MMSE for the case of large Doppler spread. Channel estimation required for coherent FDE is replaced by the simple noise power measurement. Compared with coherent FDE, FDDED is an efficient solution by trading off between good system performance and simple implementation complexity.

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