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周波数領域差動符号化を用いる 2 次元拡散/チップインターリーブ DS-CDMA

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あらまし 複数ユーザが同時に基地局にアクセスする上りリンクでは，ユーザ間の直交性が保たれず，マルチアクセス干渉（MAI）が発生し，伝送特性が大幅に劣化してしまう． 2 次元 OVSF 拡散/チップインターリーブ DS-CDMA は MAI を低減できるので，信号判定にシングルユーザのコヒーレント周波数領域等化を用いることができる．しかし，コヒーレント周波数領域等化は高精度なチャネル推定を必要とするが，高速フェージング環境下ではチャネル推定精度が劣化するという問題がある．本論文では，チャネル推定を必要としない，周波数領域差動符号等化を考える．周波数領域差動符号等化を用いる 2 次元 OVSF 拡散 DS-CDMA について，時間と周波領域の 2 重選択性フェージングチャネルにおける伝送特性を計算機シミュレーションによって明らかにしている．

キーワード 差動符号等化，DS-CDMA 上りリンク伝送，インターリーブ，2 次元拡散．

Frequency-domain Differential Detection and Equalization for 2-dimensional spread/chip-interleaved DS-CDMA

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Abstract The multiple-access interference (MAI) limits the performance of the DS-CDMA uplink transmission. 2D OVSF spread/chip-interleaved DS-CDMA is an MAI-free system, and hence single-user coherent frequency-domain equalization (FDE) can be used instead of complicated multiuser detection (MUD). However, coherent FDE needs channel estimation. Pilot-assisted channel estimation is not reliable when Doppler fading is fast. In this paper, we consider frequency-domain differential detection and equalization (FDDDE) in a multiuser/multipath environment, which doesn't require channel estimation and is robust against fast fading. The transmission performance of 2D spread/chip-interleaved DS-CDMA using FDDDE in a time- and frequency-selective fading multiuser environment is evaluated by computer simulation.

Keyword Differential detection, DS-CDMA uplink transmission, chip interleaving, 2-dimensional spreading.

1. Introduction

Direct sequence-code division multiple access (DS-CDMA) has been adopted as one of multiple access schemes in the 3rd generation (3G) wireless communication systems [1]. However, future generations of broadband wireless mobile systems are required to support a wide range of services and bit rates. As the chip rate increases, the frequency-selectivity of the fading channel becomes severer due to the increasing number of resolvable propagation paths, which makes rake combining ineffective to receive DS-CDMA signals and too complex to implement [2]. Recently, frequency-domain equalization (FDE) has been proposed for DS-CDMA [3] and it has been shown in [4] that the FDE based on the minimum mean square error (MMSE) criterion can significantly improve the BER performance of DS-CDMA downlink transmissions in a severe frequency-selective fading channel.

However, in uplink transmissions, different users' signals are asynchronous and go through different channels, producing multiple-access interference (MAI), which limits the uplink capacity. Although multiuser detection (MUD) [5], [6] schemes can be used to mitigate the detrimental effects of MAI, the MUD algorithms are relatively complex and their computational complexity increases exponentially with the number of users.

On the other hand, chip interleaving has been proposed for DS-CDMA to cancel the MAI in a quasi-synchronous multipath channel [7]. In a chip-interleaved DS-CDMA, the MUD problem is converted into a set of equivalent single-user equalization problems and single-user FDE can be used to provide good performance in a multiuser/multipath environment, provided that the propagation channel delays and transmit timings of

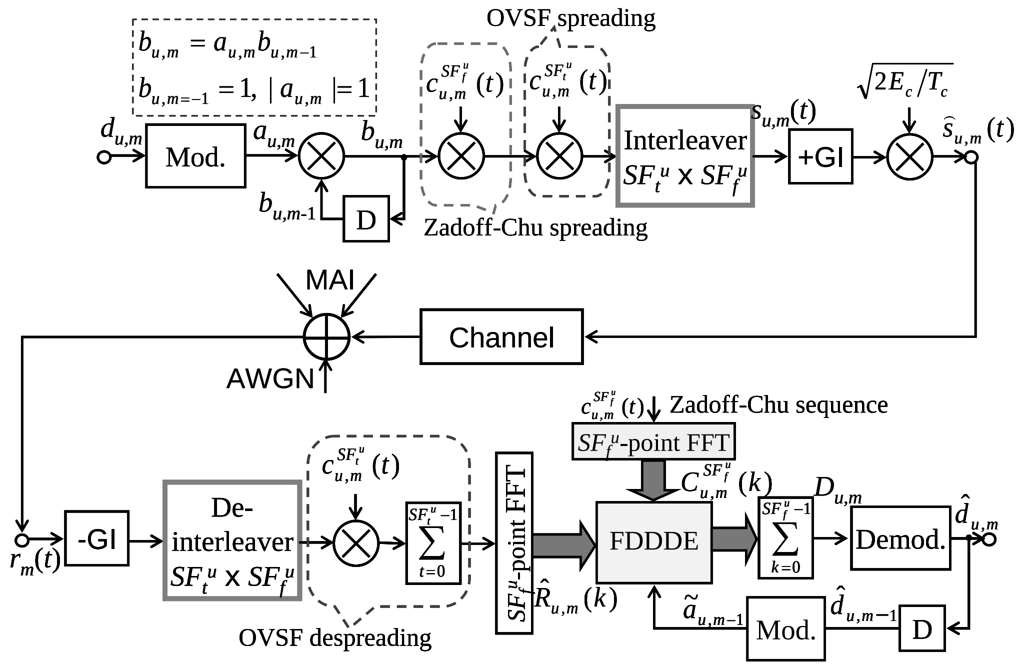


Fig.1. Transmitter and receiver structure for uplink transmissions.

different users are within a guard interval (GI) [8]. More recently, we have introduced 2-dimensional (2D) spreading using orthogonal variable spreading factor (OVSF) codes [9] for the chip-interleaved DS-CDMA uplink transmission [10], [11] to offer users flexible multirate/multi- connection services.

Although, instead of MUD, simple FDE can be applied to coherently receive 2D OVSF spread/chip-interleaved DS-CDMA signals, accurate channel estimation is necessary. The imperfect channel estimation significantly degrades the BER performance. If we use pilot-assisted channel estimation, the known pilot chip blocks need to be periodically transmitted. In order to track against fast fading, pilot block transmission rate must be increased, resulting in the transmission efficiency loss. To avoid the channel estimation, differential encoding/detection can be used. In [12], a frequency-domain differential detection and equalization (FDDDE) scheme was proposed for single-user DS-CDMA. FDDDE is attractive owing to its simplicity and robustness against fast fading. In this paper, we apply FDDDE to the uplink multiuser DS-CDMA transmission. The BER performance of 2D spread/chip-interleaved DS-CDMA with FDDDE is evaluated by computer simulation in a doublely selective (the time- and frequency-domain) fading channel and compared with that of FDE with pilot-assisted channel estimation.

2. Transmission System Model

The DS-CDMA uplink transmission model is illustrated in Fig.1, where only the uth user is shown. Here, we assume the square-root Nyquist chip shaping filter at the transmitter and the same filter at the receiver as the chip-matched filter. Throughout the paper, T_c-spaced discrete time representation is used, where T_c represents the chip duration. Here, $\lfloor x \rfloor$ is the largest integer smaller than or equal to x and $\lceil x \rceil$ is the smallest integer larger than or equal to x.

2.1. Transmitted signal

At the transmitter, the uth user's symbol sequence $\{a_{u,m}\}$ is differentially encoded into $\{b_{u,m}\}$, given by

$$b_{u,m} = a_{u,m} b_{u,m-1} \tag{1}$$

with $m \geq 1$ and $b_{u,0}$=1, where $|a_{u,m}|$=1. Next, $b_{u,m}$ is spread by a spreading sequence $\{c_{u,m}^{SF_f^u}(t); t = 0 \sim SF_f^u - 1\}$ to obtain the chip sequence as

$$s_{u,m}(t) = b_{u,m}\left(\lfloor t/SF_f^u \rfloor\right) c_{u,m}^{SF_f^u}(t \bmod SF_f^u), \tag{2}$$

where SF_f^u is the spreading factor. In this paper, the Zadoff-Chu sequence is used for $c_{u,m}^{SF_f^u}(t)$, which is given by [13]

$$c_{u,m}^{SF_f^u}(t) = \exp\left\{ j\pi(t^2 + 2mt)/SF_f^u \right\}. \tag{3}$$

Next, the SF_f^u-chip sequence $s_{u,m}(t)$ is spread by an OVSF spreading code $\{c_{u,m}^{SF_t^u}(t); t = 0 \sim SF_t^u - 1\}$ with spreading factor SF_t^u. Then, the chip interleaving is performed with column-wise input and row-wise output, as shown in Fig. 2.

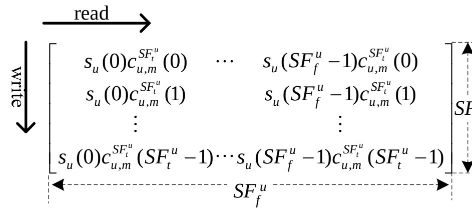


Fig. 2. Chip-interleaving.

An N_g-chip GI is inserted every N_c-chip block to avoid inter-block interference (IBI) and we set $SF_f^u=N_c$. The transmitted signal can be expressed using equivalent T_c-sampled lowpass representation as

$$\begin{aligned}\hat{s}_{u,m}(t)=&\sqrt{\frac{2E_c}{T_c}}c_{u,m}^{SF_t^u}\left(\lfloor t/(N_c+N_g)\rfloor\right)\times \\ &s_{u,m}\left((t \bmod (N_c+N_g))-N_g)\bmod N_c\right)\end{aligned} \quad (4)$$

for t=0~$SF_t^u(N_g+N_c)$-1, where E_c is the average chip energy.

The overall spreading factor of 2D codes is $SF^u=SF_t^u \times SF_f^u$. The spreading using the Zadoff-Chu sequence is used to exploit the frequency-selectivity and $SF_f^u=N_c$ should be much larger than N_g to increase the transmission efficiency. The spreading using the OVSF code is to achieve the MAI-free transmission access and SF_t^u is chosen according to the number of active users. If there are U users, $SF_t^u=2^{\lceil \log_2 U\rceil}$ can be used to allow them to access the base station without causing MAI (if the channel is time-nonselective).

2.2. Channel

The GI-inserted signal is transmitted over a frequency- and time-selective fading channel. Assuming that the channel has L independent propagation paths, the uth user's discrete-time impulse response $h_{u,m}(\tau,t)$ of the mth symbol at time t=0~$SF_t^u(N_g+N_c)$-1 is expressed as

$$h_{u,m}(\tau,t)=\sum_{l=0}^{L-1}h_{u,m,l}\left(\lfloor t/(N_c+N_g)\rfloor\right)\delta(\tau-\tau_{u,l}), \quad (5)$$

where $h_{u,m,l}(t)$ and $\tau_{u,l}$ are respectively the complex-valued path gain and time delay of the lth path with $\sum_{l=0}^{L-1}E[|h_{u,m,l}(t)|^2]=1$, and $\delta(x)$ is the delta function. We assume a block fading, where the path gain $h_{u,m,l}(\lfloor t/(N_c+N_g)\rfloor)$ remains constant over one $T=(N_c+N_g)T_c$-block interval, but vary block-by-block. $\tau_{u,l}$ is assumed to be T_c-spaced time delays and equal to $\tau_{u,l}=\tau_u+l$, l=0~L-1, where τ_u is the uth user's transmit timing offset. The maximum time delay of $\{\tau_{u,l}\}$ is assumed to be shorter than the GI (we assume some transmit timing control). Furthermore, assuming the Jake's fading model [15], we have the autocorrelation function of $h_{u,m,l}(t)$ given by $E[h_{u,m,l}(t)h_{u,m-n,l}^*(t)]=J_0(2\pi n f_D^{(u)}T)$, where $J_0(\cdot)$ is the zeroth-order Bessel function of the first kind, $f_D^{(u)}$ is the uth user's maximum Doppler frequency.

2.3. Received signal

The sum of U users' faded signals is received by a base station receiver. Assuming ideal chip sampling timing, the received signal is T_c-sampled and the GI is removed first. The GI-removed received signal can be written as

$$r_m(t)=\sum_{u=0}^{U-1}\sum_{l=0}^{L-1}h_{u,m,l}\left(\lfloor t/(N_c+N_g)\rfloor\right)\hat{s}_{u,m}(t-\tau_{u,l})+n_m(t) \quad , \quad (6)$$

where $n_m(t)$ is the additive white Gaussian noise (AWGN) with zero-mean and the variance of $2N_0/T_c$ with N_0 being the one-sided power spectrum density.

$$\begin{array}{c} \xrightarrow{\text{write}} \\ \text{read}\downarrow \begin{bmatrix} r_m(0) & r_m(1) & \cdots & r_m(N_c-1) \\ r_m(N_c) & r_m(N_c+1) & \cdots & r_m(2N_c-1) \\ \vdots & \vdots & \vdots & \vdots \end{bmatrix} \\ \longleftarrow N_c=SF_f^u \longrightarrow \end{array}$$

Fig. 3. Chip-deinterleaving.

As shown in Fig. 3, $r_m(t)$ is chip-deinterleaved and then the OVSF despreading is performed using the OVSF code $c_{u,m}^{SF_t^u}(t)$ as

$$\hat{s}_{u,m}(t)=\frac{1}{SF_t^u}\sum_{i=0}^{SF_t^u-1}r(t+iN_c)\left[c_{u,m}^{SF_t^u}(i)\right]^* \quad (7)$$

for t=0~N_c-1. After despreading, an $N_c=SF_f^u$-point FFT is applied to decompose the despread chip sequence $\{\hat{s}_{u,m}(t);t=0\sim(N_c-1)\}$ into N_c frequency components $\{\hat{R}_{u,m}(k);k=0\sim(N_c-1)\}$ as

$$\hat{R}_{u,m}(k)=\frac{1}{\sqrt{N_c}}\sum_{t=0}^{N_c-1}\hat{s}_{u,m}(t)\exp(-j2\pi k\frac{t}{N_c}). \quad (8)$$

Substituting Eqs. (2), (4), (6) into Eq. (7) and then into Eq. (8), $\hat{R}_{u,m}(k)$ can be rewritten as

$$\begin{aligned}\hat{R}_{u,m}(k)=&\sqrt{\frac{2E_c}{T_c}}d_{u,m}C_{u,m}^{SF_f^u}(k)\hat{H}_{u,m}(k) \\ &+\sqrt{\frac{2E_c}{T_c}}\sum_{\substack{u'=0\\ \neq u}}^{U-1}d_{u',m}C_{u',m}^{SF_f^{u'}}(k)Z_{u',m}(k)+\hat{\Pi}_m(k)\end{aligned} , \quad (9)$$

where the 1st, 2nd and 3rd terms represent the desired signal, the residual MAI and AWGN components, respectively, with

$$\begin{cases}\hat{H}_{u,m}(k)=\dfrac{1}{SF_t^u}\displaystyle\sum_{i=0}^{SF_t^u-1}H_{u,m}(i;k) \\ Z_{u',m}(k)=\dfrac{1}{SF_t^u}\displaystyle\sum_{i=0}^{SF_t^u-1}\left[c_{u'}^{SF_t}(i)\left\{c_u^{SF_t}(i)\right\}^*H_{u',m}(i;k)\right] \\ \hat{\Pi}_m(k)=\dfrac{1}{SF_t^u}\displaystyle\sum_{i=0}^{SF_t^u-1}\Pi_m(i;k)\end{cases}. \quad (10)$$

Here, $H_{u,m}(i;k)$ and $\Pi_m(i;k)$ are respectively the channel gain and the noise due to AWGN at the kth frequency in the $i=\lfloor t/(N_c+N_g)\rfloor$-th block of the mth symbol, and $C_{u,m}^{SF_f^u}(k)$ is the kth frequency component of $c_{u,m}^{SF_f^u}(t)$. They are given by

$$\begin{cases} H_{u,m}(i;k)=\sum_{l=0}^{L-1} h_{u,m,l}(i)\exp\{-j2\pi k\frac{\tau_l}{N_c}\} \\ \Pi_m(i;k)=\frac{1}{\sqrt{N_c}}\sum_{t=0}^{N_c-1}\exp\{-j2\pi k\frac{t}{N_c}\}\times \\ \qquad n_m[i(N_c+N_g)+t+N_g] \\ C_{u,m}^{SF_f^u}(k)=\frac{1}{\sqrt{N_c}}\sum_{t=0}^{N_c-1} c_{u,m}^{SF_f^u}(t)\exp\{-j2\pi k\frac{t}{N_c}\} \end{cases} \quad (11)$$

for k=0~N_c-1 and i=0~SF_t^u-1 with $E[|H_{u,m}(i;k)|^2]$=1, $E[|\Pi_m(i;k)|^2]$=2N_0/T_c and $E[|C_{u,m}^{SF_f^u}(k)|^2]=1$, and then

$$E[\left|\hat{H}_{u,m}(k)\right|^2]=\frac{1}{(SF_t^u)^2}\sum_{x=0}^{SF_t^u-1}\sum_{y=0}^{SF_t^u-1} J_0[2\pi|x-y|f_D^{(u)}T] \approx 1-\tfrac{1}{6}(\pi SF_t^u f_D^{(u)}T)^2\approx 1 \quad ,(12)$$

since $J_0(x)\approx 1-x^2$ if x<<1.

If the u'th mobile terminal has slow vehicle speed, $H_{u',m,i}(k)$ remains constant for i=0~SF_t^u-1 and the MAI part in Eq. (9) will disappear due to the orthogonality of the OVSF codes $\{c_{u,m}^{SF_t^u}(t);u=0\sim U-1\}$. Hence, for slow Doppler fading channels, the multiuser channel is transformed into a set of orthogonal single-user channels and the MUD problem can be converted into a set of equivalent single-user equalization problems. Therefore, single-user FDDDE [12] can be applied to such an MAI-free DS-CDMA system. As shown in Fig. 4, let

$$X_{u,m}(k)=\hat{R}_{u,m}(k)/C_{u,m}^{SF_f^u}(k) \approx \sqrt{2E_c/T_c}\,b_{u,m}\hat{H}_{u,m}(k)+\hat{\Pi}_m(k)\Big/C_{u,m}^{SF_f^u}(k) \quad ,(13)$$

where $\hat{\Pi}_m(k)/C_{u,m}^{SF_f^u}(k)$ can be treated as a new zero-mean Gaussian noise with variance $2N_0/T_c$ for given $C_{u,m}^{SF_f^u}(k)$. Since the Zadoff-Chu sequence has a constant amplitude both in the time- and in the frequency-domain [13], that is, $\left|c_{u,m}^{SF_f^u}(t)\right|=\left|C_{u,m}^{SF_f^u}(k)\right|=1$, there is no noise enhancement in $X_{u,m}(k)$.

Next, a delay-time domain windowing technique [14] is used to reduce the noise in $X_{u,m}(k)$. Firstly, an N_c-point IFFT is applied to $X_{u,m}(k)$ to obtain the delay time-domain sequence, which is a noisy instantaneous channel impulse response weighted by data information, $b_{u,m}\hat{h}_{u,m}(\tau)$. Since the actual channel impulse response is assumed to be present only within the GI length (i.e., τ=0~N_g-1), while the noise is distributed over the entire delay-time range (i.e., τ=0~N_c-1), the noise can be suppressed by zero-padding beyond GI. Then, after applying an N_c-point FFT, $\tilde{X}_{u,m}(k)$ is obtained as

$$\tilde{X}_{u,m}(k)=\sqrt{2E_c/T_c}\,b_{u,m}\hat{H}_{u,m}(k)+\tilde{\Pi}_{u,m}(k), \quad (14)$$

where $\tilde{\Pi}_{u,m}(k)$ is a zero-mean Gaussian noise with reduced variance $2\sigma^2=2N_0/T_c(N_g/N_c)$. The T_s-delayed signal $\tilde{X}_{u,m-1}(k)$ is used as the reference signal for FDDDE to recover $a_{u,m}$= $b_{u,m}b^*_{u,m-1}$ assuming $\bar{H}_{u,m}(k)\approx\bar{H}_{u,m-1}(k)$.

However, $\tilde{X}_{u,m-1}(k)$ is still noisy due to $\tilde{\Pi}_{u,m-1}(k)$ and we apply a simple infinite impulse response (IIR) fitler to further reduce the noise. As shown in Fig. 4, a first-order IIR filter with forgetting factor β ($0\le\beta\le1$) is used to improve the reference signal of FDDDE and the output of IIR filter $\bar{X}_{u,m-1}(k)$ is used for FDDDE weight and given by

$$\bar{X}_{u,m-1}(k)=\beta\bar{X}_{u,m-2}(k)\tilde{a}_{u,m-1}+(1-\beta)\tilde{X}_{u,m-1}(k) \quad (15)$$

for m>1 and $\bar{X}_{u,0}(k)=\tilde{X}_{u,0}(k)$ at m=1. Here, β is an important design parameter to trade off between the noise reduction and the tracking ability against fading. In a slow fading channel, β should be close to 1 so that the noise can be effectively reduced; while, in a fast fading channel, β should be smaller to achieve better tracking ability. When β=0, the IIR filter output $\bar{X}_{u,m-1}(k)$ becomes $\tilde{X}_{u,m-1}(k)$; a high tracking ability against fading can be achieved, but the noise reduction is insufficient. There exists an optimum value in β, which depends on the received SNR and the Doppler spread.

As shown in Fig. 4, the final data decision is based on the sum of the FDDDE outputs over all frequencies, given by

$$D_{u,m}=\sum_{k=0}^{N_c-1}D_{u,m}(k)=\sum_{k=0}^{N_c-1}\tilde{X}_{u,m}(k)\bar{X}^*_{u,m-1}(k), \quad (16)$$

based on which data demodulation is carried out.

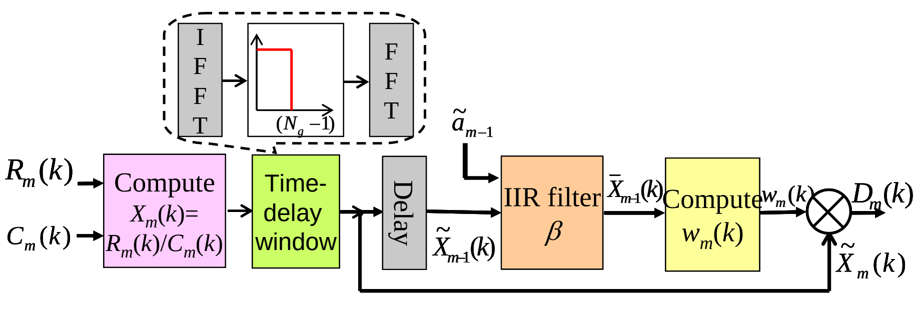


Fig. 4. FDDDE

3. Simulation Results

In the computer simulation, an L=16-path frequency-selective block Rayleigh fading channel having the uniform power delay profile is assumed for each user. The transmit timing offsets $\{\tau_u; u=0\sim(U-1)\}$ are uniformly distributed over $[-\Delta/2,\Delta/2]$ with $\Delta<(N_g-L)$ so that the maximum time delay difference is less than GI. We assume N_c=256, N_g=32 and $(SF_t^u, SF_f^u)=(2^{\lceil\log_2 U\rceil}, N_c)$ for 2D spreading. Also assumed is that each user has the same Doppler spread $f_D^{(u)}T=f_DT$ and f_DT is varied from 10^{-4} to 10^{-2} (corresponding to the vehicle speed of about 7km/h and 700km/h, respectively, with a carrier frequency of 5GHz and chip data rate of 100Mcps).

DQPSK data is received by using FDDDE; while, for comparison, the BER performance of coherent FDE using pilot-assisted channel estimation (CE) is also plotted. For coherent FDE, pilot block should be periodically inserted to estimate time-varying channel. As shown in Fig. 5, one pilot symbol is inserted every N_d data symbols and transmitted with power $R_p(E_c/T_c)$, which results in the transmission efficiency loss of $10\log(1+R_p/N_d)$.

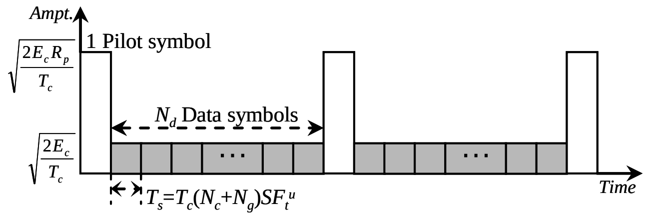
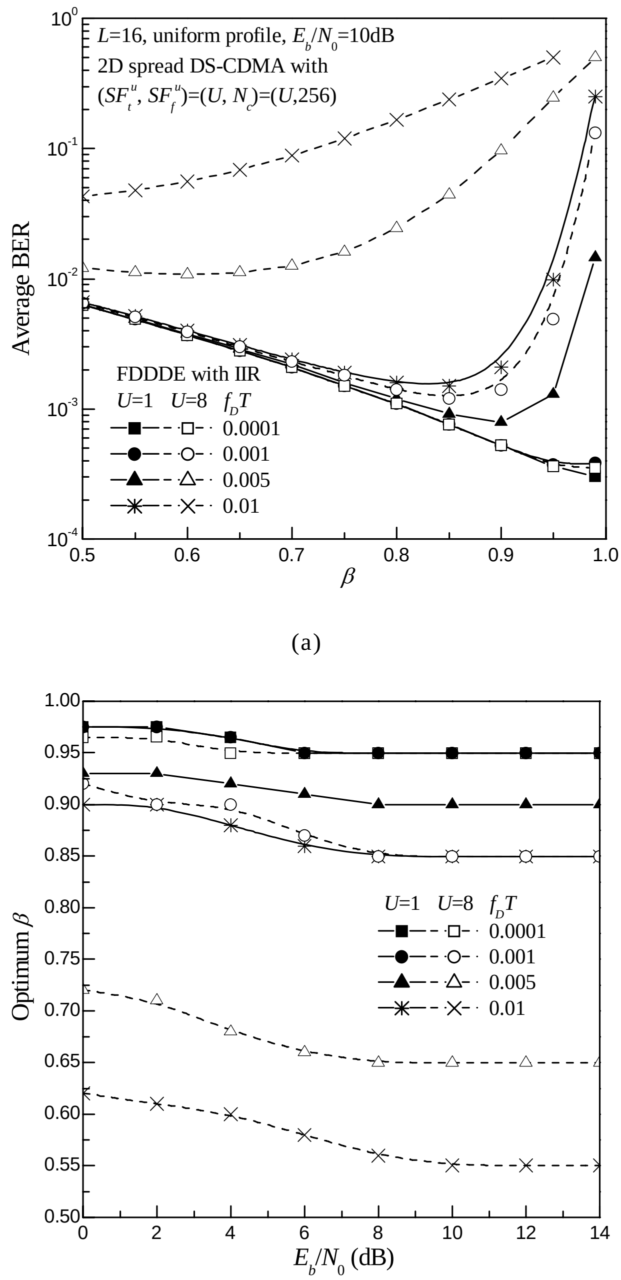


Fig. 5. Frame structures for pilot-assisted CE.

We first show in Fig. 6 the impact of β on the BER performance of 2D spread/chip-interleaved DS-CDMA using FDDDE with IIR filter under different conditions. In Fig. 6(a), we set the average received bit energy-to-the AWGN power spectrum density ratio E_b/N_0=10dB, defined by $E_b/N_0=0.5(E_c/N_0)(SF_t^uSF_f^u)\times(1+N_g/N_c)$. It can be seen that there exists an optimum β for different f_DT and U. f_DT and U. Fig. 6(b) shows the optimum β is less sensitive of E_b/N_0 than parameters f_DT and U. In the single-user case, β should be close to 1 for slow Doppler fading and the noise due to AWGN can be greatly reduced. For fast Doppler fading, β should be smaller to achieve better tracking ability. In the multiuser case, the BER performance is more sensitive to f_DT and the optimum β is much smaller for large f_DT than that of the single-user case, i.e., the optimum β is equal to 0.85 for U=1 and 0.5 for U=8. In the following simulation, the optimum β is always used.

Fig. 7 shows the BER performance of 2D spread/chip-interleaved DS-CDMA as a function of E_b/N_0. It is shown in Fig. 7 that for slow Doppler fading, i.e., $f_DT=10^{-4}$, the BER performance shows almost the same BER performance for U=1 or 8. We can see that FDDDE without IIR filter cannot work well for DQPSK-modulated signal, while the BER performance of FDDDE using IIR filter with the optimum β is better than that of coherent FDE reception using pilot-assisted CE with (R_p,N_d)=(6,31). The E_b/N_0 degradation of FDDDE with IIR filter from coherent FDE with ideal CE is as small as 0.4dB at BER=10^{-5}.



(a)

(b) Optimum β

Fig. 6. Impact of β.

How the Doppler fading rate influences the BER performance is shown in Fig. 8 for 2D spread/ chip-interleaved DS-CDMA at E_b/N_0=12dB. Although coherent FDE using pilot-assisted CE with (R_p,N_d)=(6,31) performs better than that of (R_p,N_d)=(4,15) for slow Doppler fading, the former degrades greatly as f_DT increasing when U=8. On the other hand, the BER performance of FDDDE using IIR filter with the optimum β is robust against f_DT and can approach that of coherent FDE with ideal CE when U=1.

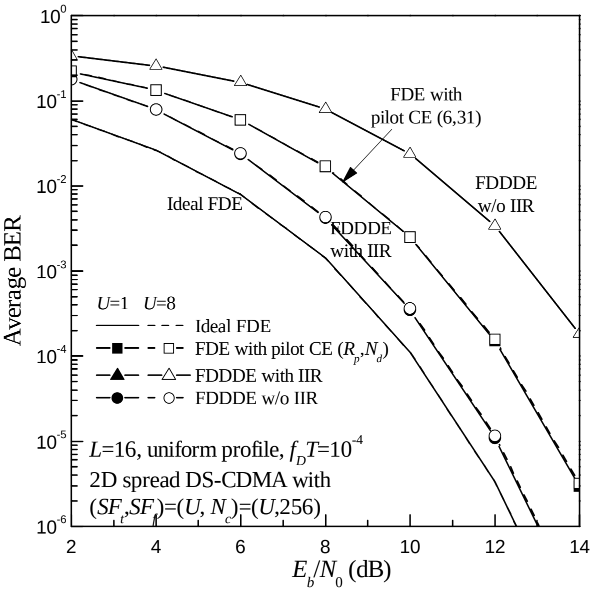


Fig. 7. Slow Doppler fading.

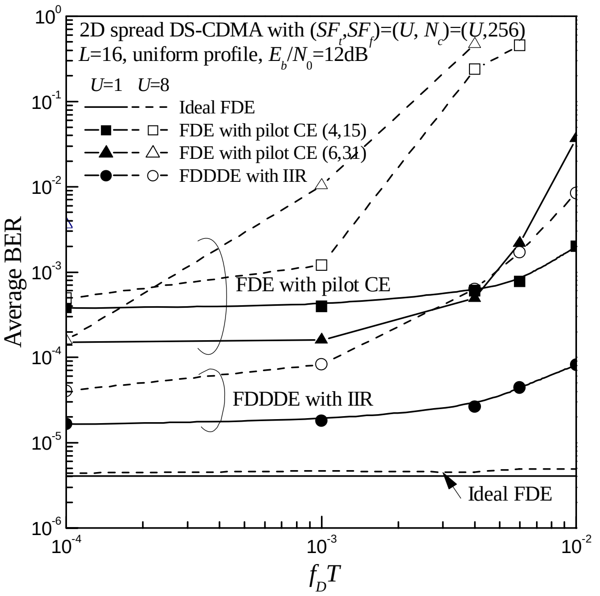


Fig. 8. Impact of f_DT.

4. Conclusions

In this paper, we presented frequency-domain differential detection and equalization (FDDDE) for 2-dimensional (2D) spread/chip-interleaved DS-CDMA in a quasi-synchronous multiuser uplink transmission. Relying on chip-interleaving and 2D spreading, a multiuser detection (MUD) problem is converted into a set of equivalent single-user equalization problems. Single-user FDDDE with IIR filter using decision feedback is applied. The BER performance in a time- and frequency-selective fading channel was evaluated by computer simulation. FDDDE was confirmed to yield much better BER performance than coherent FDE, especially in a fast fading environment.

References

[1] F. Adachi, "Wireless past and future-Evolving mobile communications systems," IEICE Trans. Fundamentals, vol.E84-A, no.1, pp.55–60, Jan. 2001.

[2] F. Adachi, D. Garg, S. Takaoka, and K. Takeda, "Broadband CDMA techniques," IEEE Wireless Commun., vol.12, no.2, pp. 8-18, Apr. 2005.

[3] D. Falconer, S. L. Ariyavisitakul, A. Benyamin-Seeyar, and B. Eidson, "Frequency domain equalization for single-carrier broadband wireless systems," IEEE Commun. Mag., vol.40, no.4, pp.58-66, April 2002.

[4] F. Adachi, K. Takeda, "Bit error rate analysis of DS-CDMA with joint frequency-domain equalization and antenna diversity combining," IEICE Trans. Commun., vol.E87-B, no.10, pp.2991-3002, Oct. 2004.

[5] H. Wei and L. Hanzo, "Genetic algorithm assisted multiuser detection for asynchronous multicarrier CDMA," IEEE Proc. VTC'04 Fall, Los Angeles, USA, Sep. 2004.

[6] X. D. Wang and H. V. Poor, "Iterative (turbo) soft interference cancellation and decoding for coded CDMA," IEEE Trans. Commun., vol.47, no.7, pp.1046-1061, July 1999.

[7] Z. Wang and G. B. Giannakis, "Block precoding for MUI/ISI-resilient generalized multicarrier CDMA with multirate capabilities," IEEE Trans. Commun., vol.49, no.11, pp.2016-2027, Nov. 2001.

[8] X. Peng, F. Chin, T. T. Tjhung and A. S. Madhukumar, "A simplified transceiver structure for cyclic extended CDMA system with frequency domain equalization," IEEE Proc. VTC'05 Spring, Sweden, pp.1565-1569, May. 2005.

[9] F. Adachi, M. Sawahashi, and K. Okawa, "Tree-structured generation of orthogonal spreading code with different lengths for foward link of DS-CDMA mobile radio," IEE Electron. Lett., vol.33, no.1, pp.27-28, Jan. 1997.

[10] L. Liu and F. Adachi, "Chip-interleaved DS-CDMA with 2-dimansional OVSF spreading codes," IEICE, RCS2004-372, pp.33-38, Mar. 2005.

[11] L. Liu and F. Adachi, "2-dimensional OVSF spreading for chip-interleaved DS-CDMA uplink transmission," Proc. IEEE, WPMC05, Alborg, Denmark, Sep. 2005.

[12] L. Liu and F. Adachi, "Frequency-domain differential detection and equalization of differentially encoded DS-CDMA signals," IEE Electron. Lett., vol.41, no.12, pp.710-712, June 2005.

[13] B. M. Propovic, "Spreading sequences for multicarrier CDMA systems," IEEE Trans. Commun., vol.47, no.6, pp.918-926, Jun. 1999.

[14] J. –J. van de Beek, O. Edfors, M. Sandell, S. K. Wilson, and P. O. Borjesson, "On channel estimation in OFDM systems," Proc. IEEE VTC'95, pp. 815-819, Chicago, Illinois, USA, July 1995.

[15] W. C. Jakes, *Microwave Mobile Communications*, IEEE Press Reissue, pp. 365-367, 1994.