Joint Frequency-domain Differential Detection and Equalization for 2-dimensional Spread/chip-interleaved DS-CDMA Uplink Transmissions

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Abstract—**The multiple-access interference (MAI) limits the performance of the DS-CDMA uplink transmission. 2 dimensional (2D) spread/chip-interleaved DS-CDMA is an MAI-free system, where 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 in a fast fading environment. In this paper, we apply joint frequency-domain differential detection and equalization (FDDDE) that requires no channel estimation to 2D spread/chip-interleaved DS-CDMA in a multiuser/multipath environment. The filter coefficient used in FDDDE is updated by estimating the normalized Doppler frequency. Computer simulation results show that 2D spread/chip-interleaved DS-CDMA using FDDDE is robust against fast fading.**

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

I. 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, the next generation mobile communication systems are required to support a wide range of high data rate services. As the chip rate increases, the frequency-selectivity of a fading channel becomes severer due to the increasing number of resolvable propagation paths with different time delays. This makes rake combining ineffective due to severe interpath interference (IPI) and too complex to implement. Recently, frequency-domain equalization (FDE) has been proposed for a single-carrier transmission [2]. More recently, it has been shown in [3], [4] that the FDE based on the minimum mean square error (MMSE) criterion can replace rake combining to significantly improve the BER performance of multicode DS-CDMA downlink transmissions in a severe frequency-selective fading channel.

However, in the uplink transmission, different users' signals go through different channels and are asynchronously received, producing multiple-access interference (MAI), which limits the uplink capacity. Although multiuser detection (MUD) schemes [5], [6] 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.

Chip interleaving has been proposed for DS-CDMA to remove the MAI for quasi-synchronous uplink transmissions [7]. In 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 including transmit timings offsets of different users are within a guard interval (GI) [8]. Recently, we have introduced 2-dimensional (2D) spreading using orthogonal variable spreading factor (OVSF) codes [9] to the chip-interleaved DS-CDMA uplink transmission [10] to offer users flexible multirate/multi-connection services.

Although single-user FDE can be used instead of MUD for the coherent reception of 2D spread/chip-interleaved DS-CDMA signals, accurate channel estimation is necessary. Imperfect channel estimation significantly degrades the bit error rate (BER) performance. If the pilot-assisted channel estimation is used, the known pilot chip blocks need to be periodically transmitted. In order to track against fast fading, pilot transmission rate must be increased, reducing the transmission efficiency. To avoid the channel estimation, differential encoding/detection can be used. Recently, we have proposed a joint frequency-domain differential detection and equalization (FDDDE) scheme for single-user DS-CDMA [11]. FDDDE is attractive owing to its simplicity and robustness against fast fading. In this paper, we apply FDDDE to 2D spread/chip-interleaved DS-CDMA. The BER performance of uplink DS-CDMA with FDDDE is evaluated by computer simulation in a doubly selective (the time- and frequency-domain) fading channel.

II. 2D Spread/chip-interleaved DS-CDMA

The DS-CDMA uplink transmission model is illustrated in Fig. 1, where only the uth user is shown (this scheme can also be applied to DS-CDMA downlink transmissions). 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. In this paper, we assume that the data rate is the same for all users and hence, the spreading factors for 2D spreading are also the same for all users.

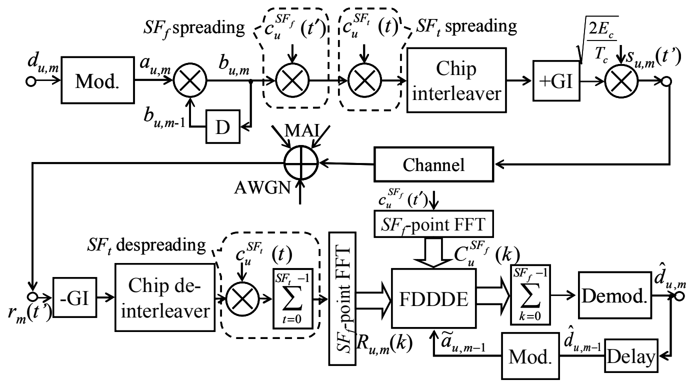

Fig. 1. Transmitter and receiver structure for the uplink transmission.

A. Transmitted signal

At the uth user's transmitter, a binary data sequence $\{d_{u,m}\}$ is first modulated into the data symbol sequence $\{a_{u,m}\}$ and then differentially encoded into $\{b_{u,m}\}$ as

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

where $|a_{u,m}|=1$ and $b_{u,0}=1$. Next, $\{b_{u,m}\}$ is spread by a spreading code $\{c_u^{SF_f}(t');t'=0\sim SF_f-1\}$ with spreading factor SF_f to obtain the chip sequence $\{s_{u,m}(t')\}$ as

$$s_{u,m}(t') = b_{u,m} c_u^{SF_f}(t')\,. \tag{2}$$

Next, the chip sequence $\{s_{u,m}(t')\}$ is spread by an OVSF code $\{c_u^{SF_t}(t);t=0\sim(SF_t-1)\}$ [9] with spreading factor SF_t. $c_u^{SF_t}(t)$ has SF_t times faster chip rate than $c_u^{SF_f}(t')$. The OVSF code tree is shown in Fig. 2.

The chip interleaving with a SF_t-by-SF_f matrix is performed with column-wise input and row-wise output, as shown in Fig. 3. The overall spreading factor of the uth user's 2D spreading is $SF=SF_t \times SF_f$. The SF_f spreading is the row-wise spreading to fully exploit the frequency-selectivity. The SF_t spreading is the column-wise spreading for orthogonal multiuser multiplexing. If there are U users, $SF_t = 2^{\lceil \log_2 U \rceil}$ can be used to allow them to access the base station without causing the MAI (if the channel is time-nonselective during SF chips) [10]. The tth block (corresponding to the tth row in Fig. 3) of the resulting 2D spread/chip-interleaved DS-CDMA signal can be expressed using equivalent lowpass representation as

$$\widetilde{s}_{u,m}(t',t) = \sqrt{2E_c/T_c}\, s_{u,m}\left(t' \bmod SF_f\right) c_u^{SF_t}(t) \tag{3}$$

for $t'=0\sim(SF_f-1)$, where E_c is the average chip energy. After an N_g-chip GI is inserted at the beginning of every SF_f-chip block to avoid inter-block interference (IBI), the 2D spread/chip-interleaved DS-CDMA signal is transmitted over a frequency- and time-selective fading channel.

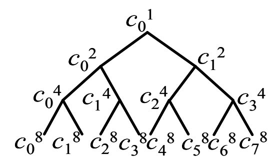

Fig. 2. OVSF code tree.

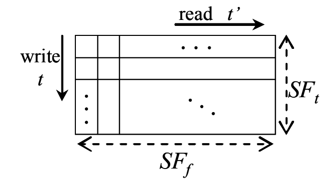

Fig. 3. 2D spreading for the mth symbol.

B. Channel

Assuming that the channel has L independent propagation paths, the uth user's discrete-time impulse response $h_{u,m}(\tau,t)$ for the reception of the mth symbol in the tth chip block is expressed as

$$h_{u,m}(\tau,t) = \sum_{l=0}^{L-1} h_{u,m,l}(t)\delta(\tau-\tau_{u,l})\,, \tag{4}$$

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 $h_{u,m,l}(t)$ remains constant over the tth block with block interval $T=(N_c+N_g)T_c$, but varies block-by-block. $\tau_{u,l}$ is assumed to be T_c-spaced time delays and equal to $\tau_{u,\lambda}=\tau_u+l$, $l=0\sim L-1$, 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 GI (we assume some transmit timing control).

C. Received signal and despreading

The superposition of U users' faded signals is received by a base station receiver. The received signal is sampled at the chip rate $1/T_c$ and the GI is removed first. The GI-removed received signal of the tth chip block of the mth symbol can be written as

$$r_m(t',t) = \sum_{u=0}^{U-1}\sum_{l=0}^{L-1} h_{u,m,l}(t)\widetilde{s}_{u,m}(t'-\tau_{u,l},t) + n_m(t',t)\,, \tag{5}$$

where $n_m(t',t)$ is the additive white Gaussian noise (AWGN) with zero-mean and the variance of $2N_0/T_c$ (N_0 is the single-sided power spectrum density). Then, $r_m(t',t)$ is chip-deinterleaved and the SF_t despreading is performed using the uth user's OVSF code $c_u^{SF_t}(t)$ to remove the MAI as

$$\hat{s}_{u,m}(t') = \frac{1}{SF_t}\sum_{t=0}^{SF_t-1} r_m(t',t)\left[c_u^{SF_t}(t)\right]^* \tag{6}$$

for $t'=0\sim SF_f-1$. After despreading, an SF_f-point FFT is applied to decompose $\hat{s}_{u,m}(t')$ into SF_f frequency components as

$$\hat{R}_{u,m}(k) = \frac{1}{\sqrt{SF_f}}\sum_{t'=0}^{SF_f-1} \hat{s}_{u,m}(t')\exp\left(-j2\pi k\frac{t'}{N_c}\right) \tag{7}$$

for $t'=0\sim SF_f-1$. Substituting Eqs. (2)-(5) into Eq. (6) and then into Eq. (7), we get

$$\begin{aligned}\hat{R}_{u,m}(k) &= \sqrt{2E_c/T_c}\, d_{u,m} C_u^{SF_f}(k) \hat{H}_{u,m}(k) \\ &+ \sqrt{2E_c/T_c} \sum_{\substack{u'=0 \\ \neq u}}^{U-1} d_{u',m} C_{u'}^{SF_f}(k) Z_{u',m}(k) + \hat{\Pi}_m(k)\end{aligned}, \tag{8}$$

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

$$\begin{cases} C_u^{SF_f}(k) = \frac{1}{\sqrt{SF_t}} \sum_{t'=0}^{SF_t-1} c_u^{SF_f}(t') \exp\left(-j2\pi k \frac{t'}{SF_t}\right) \\ \hat{H}_{u,m}(k) = \frac{1}{SF_t} \sum_{t=0}^{SF_t-1} H_{u,m}(t,k) \\ Z_{u',m}(k) = \frac{1}{SF_t} \sum_{t=0}^{SF_t-1} \left[c_{u'}^{SF_f}(t) \left\{ c_u^{SF_f}(t) \right\}^* H_{u',m}(t,k) \right] \\ \hat{\Pi}_m(k) = \frac{1}{SF_t} \sum_{t=0}^{SF_t-1} \Pi_m(t,k) \end{cases}. \tag{9}$$

Here, $H_{u,m}(t,k)$ and $\Pi_m(t,k)$ are respectively the channel gain and the noise due to AWGN at the kth frequency in the tth block for the mth symbol. They are given by

$$\begin{cases} H_{u,m}(t,k) = \sum_{l=0}^{L-1} h_{u,m,l}(t) \exp\left(-j2\pi k \frac{\tau_l}{SF_f}\right) \\ \Pi_m(t,k) = \frac{1}{\sqrt{SF_f}} \sum_{t'=0}^{SF_f-1} n_m(t',t) \exp\left(-j2\pi k \frac{t'}{SF_f}\right) \end{cases} \tag{10}$$

for k=0~SF_f−1 and t=0~SF_t−1 with $E[|H_{u,m}(t,k)|^2]=1$, $E[|\Pi_m(t,k)|^2]= 2\sigma^2_{noise}$. If $H_{u',m}(t,k)$ remains constant for t=0~SF_t−1 (i.e., very slow fading), the MAI in Eq. (8) will disappear due to the orthogonality of the OVSF codes $\{c_u^{SF_t}(t); u=0 \sim U-1\}$. The multiuser channel is transformed into a set of orthogonal single-user channels. If $H_{u',m}(t,k)$ is time-variant, the residual MAI is present.

Next, one-tap FDDDE [11] is applied (see Sect. III). The FDDDE structure is illustrated in Fig. 4. The sum of all frequency components after FDDDE, given by

$$D_{u,m} = \sum_{k=0}^{SF_f-1} D_{u,m}(k), \tag{11}$$

is the decision variable to detect $d_{u,m}$.

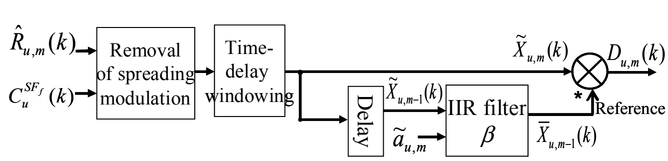

Fig. 4. FDDDE structure.

III. FDDDE

A. Removal of spreading modulation

In order to perform FDDDE, we remove the chip modulation from $\hat{R}_{u,m}(k)$ as

$$\begin{aligned} X_{u,m}(k) &= \hat{R}_{u,m}(k) / C_u^{SF_f}(k) \\ &\approx \sqrt{2E_c/T_c}\, b_{u,m} \hat{H}_{u,m}(k) + \tilde{\Pi}_{u,m}(k) \end{aligned}, \tag{12}$$

where $\tilde{\Pi}_{u,m}(k)$ can be treated as a new zero-mean Gaussian noise with variance $2\sigma^2$=2(σ^2_{noise}+σ^2_{MAI}) for given $C_u^{SF_f}(k)$, where $2\sigma^2_{noise}$ and $2\sigma^2_{MAI}$ are respectively the variances of the noise and the residual MAI. In this paper, the Zadoff-Chu sequence [12] is used for $c_u^{SF_f}(t')$, given by $c_u^{SF_f}(t') = \exp\{j\pi(t'^2 + 2ut')/SF_f\}$. Since the Zadoff-Chu sequence has a constant amplitude both in the time-domain and in the frequency-domain [12], that is, $\left|c_u^{SF_f}(t')\right| = \left|C_u^{SF_f}(k)\right| = 1$, there is no noise enhancement in $X_{u,m}(k)$.

B. Delay time-domain windowing

A delay time-domain windowing technique [13] is used to reduce the noise in $X_{u,m}(k)$. Firstly, an SF_f-point IFFT is applied to $X_{u,m}(k)$ to obtain the delay time-domain sequence, $x_{u,m}(t')$, given by

$$\begin{aligned} x_{u,m}(t') &= \frac{1}{\sqrt{SF_f}} \sum_{t=0}^{SF_f-1} X_{u,m}(k) \exp(-j2\pi t' \frac{k}{SF_f}) \\ &= \sqrt{2E_c/T_c}\, b_{u,m} \hat{h}_{u,m}(t') + \tilde{n}_{u,m}(t') \end{aligned}, \tag{13}$$

which is a noisy instantaneous channel impulse response modulated by $b_{u,m}$. Since real channel impulse response $\hat{h}_{u,m}(t')$ is assumed to be present only within the GI length (i.e., $t' = 0 \sim N_g - 1$); while the noise $\tilde{n}_{u,m}(t')$ is distributed over the entire delay-time range (i.e., $t' = 0 \sim SF_f - 1$). Therefore, the noise can be suppressed by zero-padding beyond GI. Then, after applying an SF_f-point FFT, the output $\tilde{X}_{u,m}(k)$ is obtained as

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

where $\tilde{\Pi}'_{u,m}(k)$ is a zero-mean Gaussian noise with reduced variance $(N_g/SF_f)2\sigma^2$.

C. Improved reference signal

Remembering $a_{u,m}=b_{u,m}b^*_{u,m-1}$ from Eq. (1), $\tilde{X}_{u,m-1}(k)$ can be used as the reference for FDDDE. 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) filter to reduce the noise. As shown in Fig. 4, a first-order IIR filter with forgetting factor β ($0 \le \beta \le 1$) is used to improve the reference signal. The filter output $\bar{X}_{u,m-1}(k)$ to be used as the improved reference signal is given by

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

where $\widetilde{a}_{u,m-1}$ is the feedback from the previous decision on the $(m-1)$th symbol and $\overline{X}_{u,0}(k)=\widetilde{X}_{u,0}(k)$. Here, β is an important design parameter to trade off between the noise reduction and the tracking ability against fading. There exists the optimum value in β, which depends on the received SNR and the Doppler spread.

Finally, the FDDDE output is given by

$$D_{u,m}(k)=\widetilde{X}_{u,m}(k)\overline{X}^{*}_{u,m-1}(k)\,. \quad (16)$$

D. Doppler Frequency Estimation

Assuming that the maximum Doppler frequency f_D is the same for all users, we can find the optimum β that minimizes the average BER as a function of the normalized Doppler frequency f_DT_c and SNR by computer simulation (in Sect. V). The normalized Doppler frequency f_DT_c can be estimated as follows. Assuming the Jake's fading model [14], the correlation function of $h_{u,m,l}(t)$ and $h_{u,m-1,l}(t)$ is given by

$$\rho=\frac{E[h_{u,m,l}(t)h^{*}_{u,m-1,l}(t)]}{E[|h_{u,m,l}(t)|^2]}=J_0(2\pi f_DT_cSF_t\ (SF_f+N_g))\,, \quad (17)$$

where $J_0(\cdot)$ is the zeroth-order Bessel function of the first kind. According to Eq. (17), we can use $x_{u,m}(t')$ in Eq. (13) to estimate ρ as

$$\hat{\rho}=\mathrm{Re}\left\{\sum_{t'=0}^{N_g-1}x_{u,m}(t')x^{*}_{u,m-1}(t')\right\}\Bigg/\left[\sum_{t'=0}^{N_g-1}|x_{u,m}(t')|^2-2\hat{\sigma}^2N_g\right], \quad (18)$$

where $\hat{\sigma}^2$ is the estimate of the noise variance, given as [15]

$$2\hat{\sigma}^2=\frac{1}{SF_f-N_g}\sum_{t'=N_g}^{SF_f-1}|x_{u,m}(t')|^2\,. \quad (19)$$

Since $J_0(2\pi x)\approx 1-(\pi x)^2$ if $x<0.1$, the estimate of f_DT_c, $\hat{f}_DT_c$ is obtained as

$$\hat{f}_DT_c=[\pi SF_t\ (SF_f+N_g)]^{-1}\sqrt{1-\hat{\rho}}\,, \quad (20)$$

which is used to update β for FDDDE.

IV. Simulation Model

For 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 N_g=32. In the computer simulation, the BER performance is evaluated for different values of f_DT_c ranging from 3.5×10^{-7} to 3.5×10^{-5} (corresponding to the vehicle speed of about 7.5km/h and 750km/h, respectively, with a carrier frequency of 5GHz and a chip rate of 100Mcps).

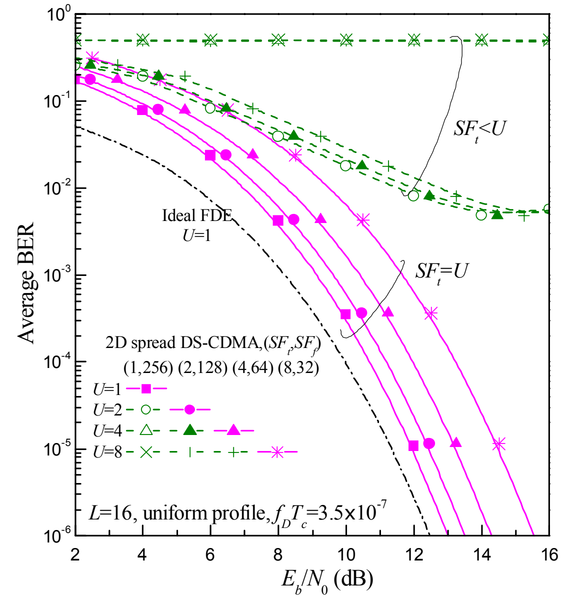

Fig. 5. BER performance in slow fading.

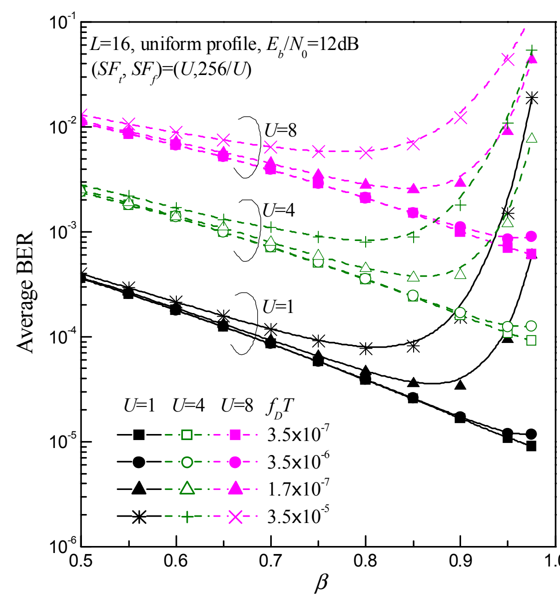

Fig. 6. Impact of β.

We first show in Fig. 5 the BER performance of 2D spread/chip-interleaved DS-CDMA as a function of E_b/N_0 for a slow Doppler fading, i.e., f_DT_c=3.5x10^{-7}. Here, the forgetting factor β is set to 0.975 for FDDDE with IIR filtering. The average received bit energy-to-the AWGN power spectrum density ratio E_b/N_0 is defined by $E_b/N_0=0.5(E_c/N_0)(SF_tSF_f)(1+N_g/SF_f)$. We assume the overall spreading factor $SF=SF_t\times SF_f=256$ for all users. If $SF_t<U$, users are partitioned into SF_t groups. Users in each group are interference-free from other groups; but the MAI is present in each group and an error floor is produced when $SF_t<U$. It can been seen that 2D spread/chip-interleaved DS-CDMA with

$(SF_t,SF_f)=(U,256/U)$ gives the best BER performance, since the MAI is removed completely when channels are time-invariant. For comparison, the BER performance of coherent FDE with ideal channel estimation (CE) for the single-user case (U=1) is also plotted. When U=1, the E_b/N_0 degradation of FDDDE with IIR filtering from coherent FDE with ideal CE is as small as 0.4dB at BER=10^{-5}. However, the E_b/N_0 loss due to the insertion of GI, equal to $10\log(1+N_g/SF_f)$ dB, gets larger as $SF_f=SF/U$ decreases.

Fig. 6 shows the impact of β on the BER performance of 2D spread/chip-interleaved DS-CDMA with $(SF_t,SF_f)=(U,256/U)$ using FDDDE with IIR filtering when E_b/N_0=12dB. It can be seen that the optimum β is different for different f_DT_c but not sensitive to U. According to the simulation results, the optimal β is found to be $\beta=0.975-10SF_t(SF_f+N_g)\hat{f}_DT_c$ if $10^{-4}<SF_t\,(N_g+SF_f)\hat{f}_DT_c\le 0.02$, where $\hat{f}_DT_c$ is the estimate of f_DT_c (see Eq. (20)).

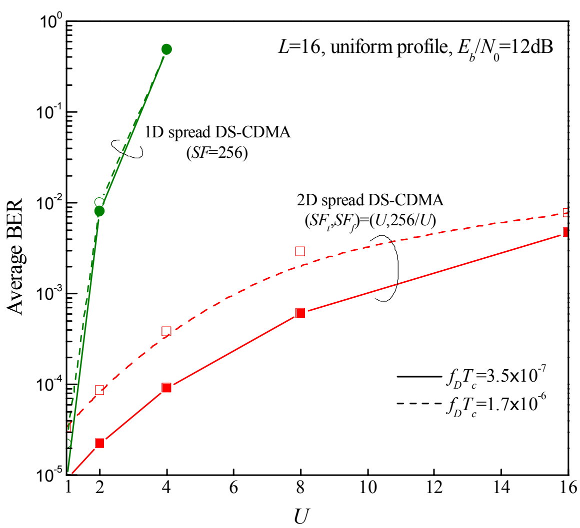

Fig. 7. Impact of U.

Fig. 7 compares the BER performances of 1D spread(conventiaonl) DS-CDMA and 2D spread DS-CDMA using FDDDE when E_b/N_0=12dB. It is observed that for 1D spread DS-CDMA using FDDDE with SF=256, which corresponds to the 2D spread/chip-interleaved DS-CDMA with (SF_t,SF_f)=(1,256), the BER performance degrades significantly when U increases due to large MAI. However, since 2D spread DS-CDMA with $(SF_t,SF_f)=(U,256/U)$ is an MAI-free system in a slow fading channel, i,e., $f_DT_c=3.5\times10^{-7}$ (7.5km/h), about U=16 users can be accommodated at BER=10^{-2}. The BER degradation for large U is the consequence of the increasing E_b/N_0 loss due to the GI insertion. As U increases, SF_f (=256/U) decreases and the E_b/N_0 loss, which is equal to $10\log(1+N_g/SF_f)=10\log(1+N_gU/256)$ dB, increases. In the case of fast fading, i.e., $f_DT_c=1.7\times10^{-6}$ (375km/h), 2D spread DS-CDMA using FDDDE with the optimum β can still accommodate U=16 users at BER=10^{-2}. In addition, only single-user FDDDE is required and the complexity is the same irrespective of the number U of users.

V. Conclusions

In this paper, we presented an application of joint frequency-domain differential detection and equalization (FDDDE) to 2-dimensional (2D) spread/chip-interleaved DS-CDMA in a quasi-synchronous 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 an IIR filter using decision feedback is applied and the filter coefficient is updated by estimating the normalized Doppler frequency. The BER performance in a time- and frequency-selective Rayleigh fading channel was evaluated by computer simulation. 2D spread DS-CDMA with FDDDE was confirmed to yield much better BER performance than 1D spread DS-CDMA with FDDDE. In this paper, transmit power control was not considered. This is left for future study.

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