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Joint Frequency-Domain Equalization and Despreading for DS-CDMA Cyclic Delay Transmit Diversity

Tetsuya YAMAMOTO† Kazuki TAKEDA† and Fumiyuki ADACHI‡
Dept. of Electrical and Communications Engineering, Graduate school of Engineering, Tohoku University
6-6-05 Aza-Aoba, Aramaki, Aoba-ku, Sendai, 980-8579 Japan
E-mail: †{yamamoto, kazuki}@mobile.ecei.tohoku.ac.jp, ‡adachi@ecei.tohoku.ac.jp

***Abstract*- Cyclic delay transmit diversity (CDTD) using frequency-domain equalization (FDE) based on the minimum mean squire error (MMSE) criterion is a powerful technique to improve the bit error rate (BER) performance of direct sequence code division multiple access (DS-CDMA) signal transmission in a frequency-selective fading channel. CDTD simultaneously transmits the same signal from different antennas after adding different cyclic delays to increase the number of equivalent propagation paths and hence achieves the larger frequency diversity gain. However, the BER performance of DS-CDMA with CDTD using MMSE-FDE is limited by the residual inter-chip interference (ICI) after FDE. In this paper, we propose a new joint FDE and despreading suitable for CDTD, which can simultaneously perform equalization and despreading in the frequency-domain. We evaluate the achievable BER performance by computer simulation.**

Keywords: DS-CDMA, frequency-domain equalization, cyclic delay transmit diversity

I. INTRODUCTION

High-speed data transmissions are demanded in next generation mobile wireless communication systems. However, since the mobile wireless channel is composed of many propagation paths with different time delays, a frequency-selective fading channel is produced. In a frequency-selective fading channel, the bit error rate (BER) performance of single-carrier (SC) transmission significantly degrades due to a severe inter-symbol interference (ISI) [1]. In the third generation mobile communication systems, direct sequence code division multiple access (DS-CDMA) using coherent rake combining is adopted to obtain the path-diversity gain [2]. However, the wireless channel for the data transmissions higher than few tens of Mbps becomes severely frequency-selective and hence the BER performance of DS-CDMA with rake combining degrades due to a strong inter-path interference (IPI) [3]. Recently, it was shown that frequency-domain equalization based on the minimum mean square error criterion (MMSE-FDE) can replace rake combining to significantly improve the BER performance of DS-CDMA [3, 4].

To further improve the BER performance, the use of transmit diversity technique is effective. Recently, cyclic delay transmit diversity (CDTD) has been proposed for multi-carrier transmissions [5, 6]. CDTD can increase the number of equivalent paths by transmitting the same data from different antennas after adding different cyclic delays, and hence can achieve large frequency diversity gain. CDTD can also be applied to DS-CDMA using MMSE-FDE and can improve the BER performance [7]. In [7], the time-domain despreading is performed after MMSE-FDE. However, the residual inter-chip interference is present after MMSE-FDE and this limits the BER performance improvement [8].

In this paper, we propose a new joint FDE and despreading for CDTD in the frequency-domain. We compare, by computer simulation, the average BER performances achievable with the proposed joint FDE and despreading and the conventional MMSE-FDE. The remainder of the paper is organized as follows. Sect. II introduces CDTD. In Sect. III, the proposed joint FDE and despreading is described. In Sect. IV, the BER performance of CDTD with the proposed joint FDE and despreading is evaluated by computer simulation and compared with CDTD using the conventional MMSE-FDE. The paper is concluded in Sect. V.

II. CYCLEC DELAY TRANSMIT DIVERSITY

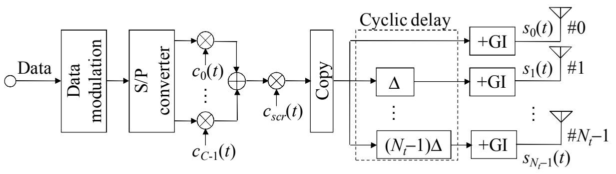


Figure 1. Transmitter structure of multi-code DS-CDMA using CDTD.

Figure 1 shows the transmitter structure of DS-CDMA using CDTD. Throughout this paper, the chip-spaced discrete time representation is used. At the transmitter, a binary data sequence is data-modulated and then, serial/parallel (S/P)-converted to C parallel data sequences. The ith data symbol d_i, i=0~C−1, is spread by an orthogonal spreading code $\{c_i(t); t=0\sim SF-1\}$ with the spreading factor SF. The resultant C chip sequences are code-multiplexed and further multiplied by a common scramble sequence $c_{scr}(t)$.

In CDTD, the same chip sequence is simultaneously transmitted from different antennas after adding different cyclic delays. N_t copies of the chip sequence are generated and then, cyclic time delay $n\Delta$ is added before the transmission from the nth antenna (n=0~N_t−1). The transmitted chip sequence from the nth antenna, $s_n(t)$,

t=0~SF−1, can be expressed using the lowpass equivalent representation as

$$s_n(t)=\sqrt{2P_n}\sum_{i=0}^{C-1}d_i\cdot\bar{c}_i((t-n\Delta)\bmod SF), \qquad (1)$$

where P_n is the transmit power of the nth antenna under the total transmit power constraint $\sum_{n=0}^{N_t-1}P_n=P$, $\bar{c}_i(t)=c_i(t\bmod SF)c_{scr}(t)$, and C is the code-multiplexing order. Finally, the last N_g chips of each block are copied as a cyclic prefix and inserted into the guard interval (GI) placed at the beginning of each block.

The signal transmitted from N_t antennas goes through different frequency-selective fading channels, each composed of L distinct paths. The received signal $r(t)$ is a superposition of N_t transmitted signal and can be expressed as

$$\begin{aligned}r(t)&=\sum_{n=0}^{N_t-1}\sum_{l=0}^{L-1}h_{n,l}s_n(t-\tau_{n,l})+\eta(t)\\&=\sum_{n=0}^{N_t-1}\sqrt{2P_n}\sum_{i=0}^{C-1}d_i\sum_{l=0}^{L-1}h_{n,l}\bar{c}_i\big((t-n\Delta-\tau_{n,l})\bmod SF\big)+\eta(t)\end{aligned}, \qquad (2)$$

where $h_{n,l}$ and $\tau_{n,l}$ are respectively the complex-valued path gain and the time delay of the lth path between the nth transmit antenna and the receiver, and $\eta(t)$ is the zero-mean additive white Gaussian noise (AWGN) having the variance $2N_0/T_c$ with N_0 being the one-sided noise power spectrum density.

At the receiver, after the GI removal, SF-point fast Fourier transform (FFT) is applied to transform the received signal $\{r(t);\ t=0\sim SF-1\}$ into the frequency-domain signal $\{R(k);\ k=0\sim SF-1\}$. $R(k)$ is given by

$$\begin{aligned}R(k)&=\sum_{t=0}^{SF-1}r(t)\exp\left(-j2\pi k\frac{t}{SF}\right)\\&=\sum_{n=0}^{N_t-1}\sqrt{2P_n}H_n(k)\sum_{i=0}^{C-1}d_i\left\{C_i(k)\exp\left(-j2\pi k\frac{n\Delta}{SF}\right)\right\}+\Pi(k)\end{aligned}, \qquad (3)$$

where $C_i(k)$, $H_n(k)$ and $\Pi(k)$ are the kth frequency component of the spreading sequence $\{\bar{c}_i(t);\ t=0\sim SF-1\}$, the channel impulse response and noise, respectively. These are given by

$$\begin{cases}C_i(k)=\sum_{t=0}^{SF-1}\bar{c}_i(t)\exp\left(-j2\pi k\frac{t}{SF}\right)\\ H_n(k)=\sum_{l=0}^{L-1}h_{n,l}\exp\left(-j2\pi k\frac{\tau_{n,l}}{SF}\right)\\ \Pi(k)=\sum_{t=0}^{SF-1}\eta(t)\exp\left(-j2\pi k\frac{t}{SF}\right)\end{cases}. \qquad (4)$$

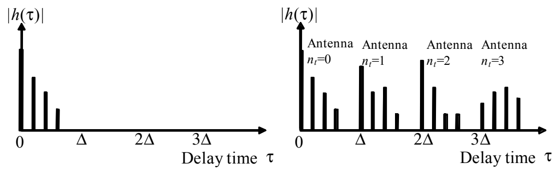
If the transmit power of each antenna is the same, Eq. (3) can be expressed as

$$R(k)=\sqrt{\frac{2P}{N_t}}H(k)\sum_{i=0}^{C-1}d_i\cdot C_i(k)+\Pi(k), \qquad (5)$$

where $H(k)$ is the composite channel gain and is given by

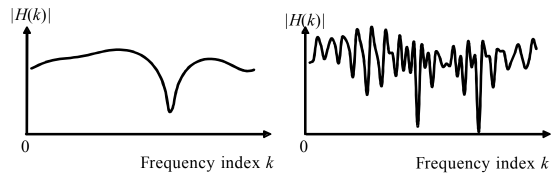
$$\begin{aligned}H(k)&=\sum_{n=0}^{N_t-1}H_n(k)\exp\left(-j2\pi k\frac{n\Delta}{SF}\right)\\&=\sum_{n=0}^{N_t-1}\sum_{l=0}^{L-1}h_{n,l}\exp\left(-j2\pi k\frac{n\Delta+\tau_{n,l}}{SF}\right)\end{aligned}. \qquad (6)$$

The impulse response $h(\tau)$ and channel gain $H(k)$ of the composite channel are shown in Fig. 2 and Fig. 3, respectively, for an L=4 path channel when N_t=1 and N_t=4. CDTD can increase the number of equivalent paths by transmitting the same data from different antennas after adding different cyclic delays and hence, the CDTD channel can be treated as a composite channel which is the sum of N_t channels. It can be understood from Fig. 3 that CDTD enhances the degree of frequency-selectivity of the channel. This can be exploited by the use of MMSE-FDE which can achieve larger diversity gain.



(a) N_t=1 (b) N_t=4

Figure 2. Impulse response $h(\tau)$ of the composite channel.



(a) N_t=1 (b) N_t=4

Figure 3. Channel gain $H(k)$ of the composite channel.

III. MMSE-FDE FOR CDTD

In this section, after reviewing the conventional MMSE-FDE used for DS-CDMA using CDTD [7, 8], we propose the joint FDE and despreading for single-code and multi-code transmissions. Fig. 4 shows the receiver structure of DS-CDMA CDTD using the joint FDE and despreading. With joint FDE and despreading, all signal processing can be simultaneously performed in the frequency-domain.

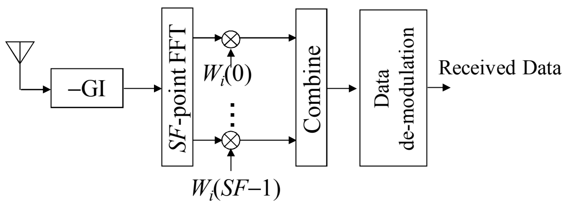


Figure 4. Receiver structure of DS-CDMA using CDTD using the joint FDE and despreading.

A. Conventional MMSE-FDE

The conventional MMSE-FDE is carried out on $R(k)$ as

$$\hat{R}(k)=R(k)W(k), \qquad (7)$$

where $W(k)$ is the MMSE-FDE weight and is given by

$$W(k)=\frac{H^*(k)}{|H(k)|^2+\left(\frac{C}{N_t}\frac{PT_c}{N_0}\right)^{-1}}. \tag{8}$$

After MMSE-FDE, the frequency-domain signal $\{\hat{R}(k); k=0\sim SF-1\}$ is transformed by SF-point inverse FFT (IFFT) into a time-domain signal $\{\hat{r}(t); t=0\sim SF-1\}$ as

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

Finally, despreading is performed, giving

$$\hat{d}_i=\frac{1}{SF}\sum_{t=0}^{SF-1}\hat{r}(t)c_i^*(t \bmod SF)c_{scr}^*(t), \tag{10}$$

which is the decision variable associated with d_i. The instantaneous signal-to-interference plus noise power ratio (SINR) for the given set of the composite channel gain $\{H(k); k=0\sim SF-1\}$ is given by [9]

$$\gamma=\frac{\frac{2}{N_t}\frac{PT_cSF}{N_0}\left|\frac{1}{SF}\sum_{k=0}^{SF-1}\hat{H}(k)\right|^2}{\left[\begin{array}{l}\frac{1}{SF}\sum_{k=0}^{SF-1}|W(k)|^2\\ +\left(\frac{C}{N_t}\frac{PT_c}{N_0}\right)\left\{\frac{1}{SF}\sum_{k=0}^{SF-1}|\hat{H}(k)|^2-\left|\frac{1}{SF}\sum_{k=0}^{SF-1}\hat{H}(k)\right|^2\right\}\end{array}\right]}, \tag{11}$$

where $\hat{H}(k)=H(k)W(k)$.

As CDTD can enhance the frequency-selectivity of the channel, CDTD can achieve larger diversity gain. However, larger residual ICI is produced after MMSE-FDE. This inter-chip interference contains the self-code interference and the inter-code interference.

B. FDE-weight for single-code case

For the single-code case, the frequency-domain signal $\{R(k); k=0\sim SF-1\}$ can be expressed as

$$R(k)=\left(\sqrt{\frac{2P}{N_t}}H(k)C_0(k)\right)d_0+\Pi(k). \tag{12}$$

From Eq. (12), the weight which maximizes the signal-to-noise ratio (SNR) at each subcarrier is derived as

$$W_{MRC}(k)=\{H(k)C_0(k)\}^*, \tag{13}$$

which is called the maximal-ratio combining (MRC) weight. The decision variable is obtained as

$$\hat{d}_0=\sum_{k=0}^{SF-1}R(k)W_{MRC}(k). \tag{14}$$

Using this joint MRC-FDE and despreading, equalization and despreading can be simultaneously performed in the frequency-domain. The conditional SNR for the given set of the composite channel gains $\{H(k); k=0\sim SF-1\}$ is given by

$$\gamma=\frac{\frac{2}{N_t}\frac{PT_c}{N_0}\left|\frac{1}{SF}\sum_{k=0}^{SF-1}|H(k)C_0(k)|^2\right|^2}{\frac{1}{SF}\sum_{k=0}^{SF-1}|H^*(k)C_0^*(k)|^2}=\frac{2}{N_t}\frac{PT_c}{N_0}\frac{1}{SF}\sum_{k=0}^{SF-1}|H(k)C_0(k)|^2. \tag{15}$$

It can be clearly seen that no ICI is present unlike the conventional MMSE-FDE.

C. FDE-weight for multi-code case

For the multi-code case, the frequency-domain signal is a sum of simultaneously transmitted C data symbols. For this reason, joint MRC-FDE and despreading increases the inter-code interference and degrades the BER performance. Therefore, we apply the joint MMSE-FDE and despreading to CDTD. After FDE, the frequency-domain signal $\{\hat{R}_i(k); k=0\sim SF-1\}$ is transformed by SF-point IFFT into a delay-time domain signal $\{y_i(\tau); \tau=0\sim SF-1\}$. We derive the MMSE-FDE weight which minimizes the MSE between $y_i(\tau)$ and $d_i\delta(\tau)$. This means that $y_i(0)$ can be the decision variable $\hat{d}_i$. The MMSE-weight is given as

$$W_i(k)=\frac{\{C_i(k)H(k)\}^*}{\frac{1}{SF}|H(k)|^2\sum_{i'=0}^{C-1}|C_{i'}(k)|^2+\left(\frac{1}{N_t}\frac{PT_c}{N_0}\right)^{-1}} \tag{16}$$

and the decision variable is obtained as

$$\hat{d}_i=y_i(0)=\sum_{k=0}^{SF-1}R(k)W_i(k). \tag{17}$$

By the use of joint MMSE-FDE and despreading, equalization and despreading can be simultaneously performed in the frequency-domain, similar to the single-code case (see Eq. (14)). All frequency components are coherently combined by multiplying the complex conjugate of $C_i(k)H(k)$.

The conditional SINR for the given set of the composite channel gains $\{H(k); k=0\sim SF-1\}$ is given by

$$\gamma=\frac{\frac{2}{N_t}\frac{PT_c}{N_0}\left|\frac{1}{SF}\sum_{k=0}^{SF-1}\hat{H}(k)C_i(k)\right|^2}{\frac{1}{SF}\sum_{k=0}^{SF-1}|W_i(k)|^2+\left(\frac{1}{N_t}\frac{PT_c}{N_0}\right)\sum_{\substack{i'=0\\ \neq i}}^{C-1}\left|\frac{1}{SF}\sum_{k=0}^{SF-1}\hat{H}(k)C_{i'}(k)\right|^2}. \tag{18}$$

IV. SIMULATION RESULTS

The simulation condition is summarized in Table 1. The channel is assumed to be an L=16-path frequency-selective block Rayleigh fading channel having exponential power delay profile with the path decay factor α. Ideal timing recovery and channel estimation are assumed.

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TABLE I
SIMULATION CONDITION.

Transmitter	Number of transmit antennas	N_t=4
	Modulation	QPSK, 16QAM, 64QAM
	GI length	N_g=16
	Spreading sequence	Walsh sequence
	Spreading factor	SF=64
	Number of parallel codes	C=1~64
	Scramble code	Long PN sequence
Channel	Fading type	Frequency-selective block Rayleigh
	Power delay profile	L=16-path exponential power delay profile
	Decay factor	α=0, 6 (dB)
Receiver	FFT block size	SF
	FDE	MMSE, MRC
	Channel estimation	Ideal

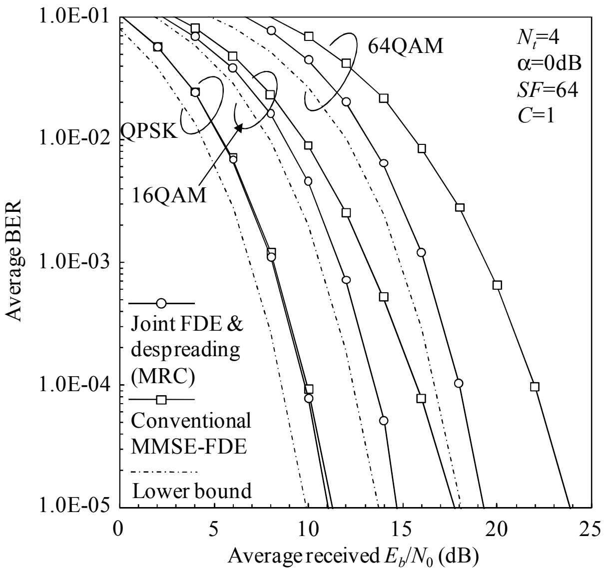
A. Single-code case

Figure 5 compares the BER performances of CDTD for the single-code case using the joint FDE and despreading and the conventional MMSE-FDE when N_t=4 as a function of average received bit energy-to noise power spectrum density ratio $E_b/N_0(=(ST_c/N_0)(SF+N_g)/\log_2 M)$, where M is the modulation level. Also plotted is the BER performance of the theoretical lower bound [9].

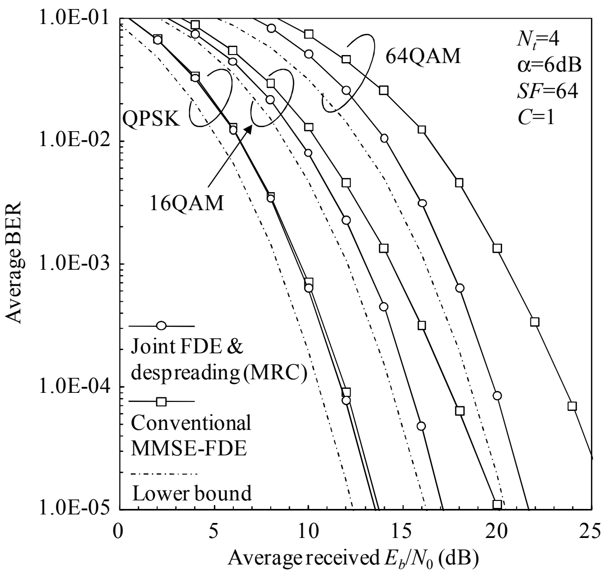
When QPSK data modulation (M=4) is used, the influence of the ICI is not significant and therefore, only a slight performance difference is seen between the joint FDE and despreading and the conventional MMSE-FDE. For higher modulation level (i.e., M=16, 64), since the Euclidian distance between the symbol constellation points becomes smaller, the BER performance using the conventional MMSE-FDE degrades due to the residual ICI. However, since the joint FDE and despreading does not produce the residual ICI at all, better BER performance can be achieved. The performance improvement gets larger for higher level modulation. When the joint FDE and despreading is used, the BER performance approaches the theoretical lower bound (a degradation from the theoretical lower bound is due to the GI insertion loss (0.97dB)).

B. Multi-code case

Figure 6 plots the average BER performances of CDTD for the multi-code case using the joint FDE and despreading and the conventional MMSE-FDE when N_t=4 with the code multiplexing order C as a parameter. 16QAM is considered. For comparison, the lower-bound BER performance is also plotted. The BER performance of CDTD degrades as C increases due to the increased residual ICI. However, the joint FDE and despreading can achieve slightly better BER performance than the conventional MMSE-FDE. This is because the proposed joint FDE and despreading minimizes the MSE between the transmitted symbol and soft decision variable, while the conventional MMSE-FDE minimizes the MSE between the transmitted and received DS-CDMA signals before despreading. In the case of C=SF (full code-multiplex), both the joint FDE and despreading and the conventional MMSE-FDE achieve almost the same BER performance.



(a) α=0 dB

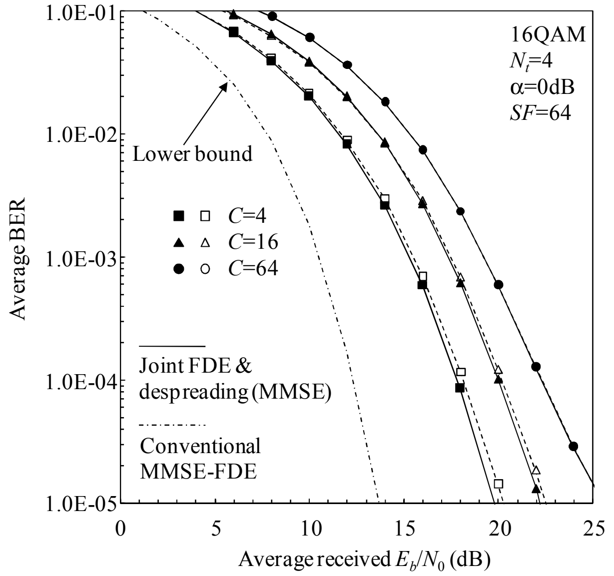


(b) α=6 dB

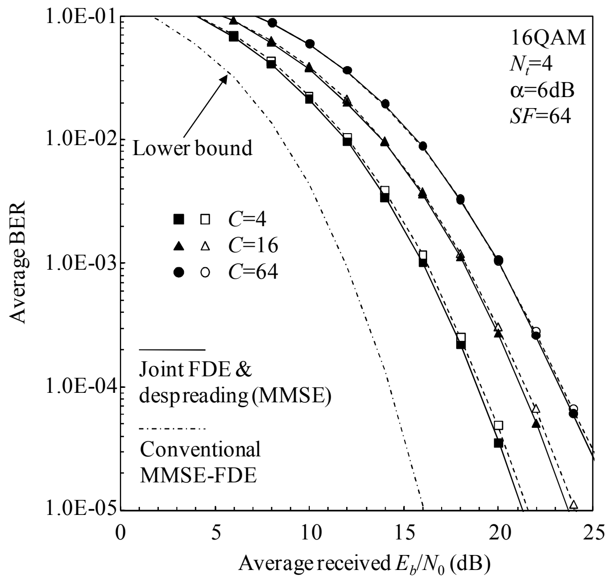
Figure 5. Simulated BER performance of CDTD for single-code case.

C. Complexity

Figure 7 shows the number of complex multiply operations required in FDE (including FFT/IFFT) and despreading using the joint FDE and despreading and the conventional MMSE-FDE as a function of the spreading factor SF. The number of complex multiply operations is $SF\log_2 SF+2SF$ for the conventional MMSE-FDE and is $(SF/2)\log_2 SF+SF$ for the joint FDE and despreading. The joint FDE and despreading requires lower computational complexity than the conventional MMSE-FDE.



(a) α=0 dB



(b) α=6 dB

Figure 6. Simulated BER performance of CDTD with the number C of parallel codes as a parameter.

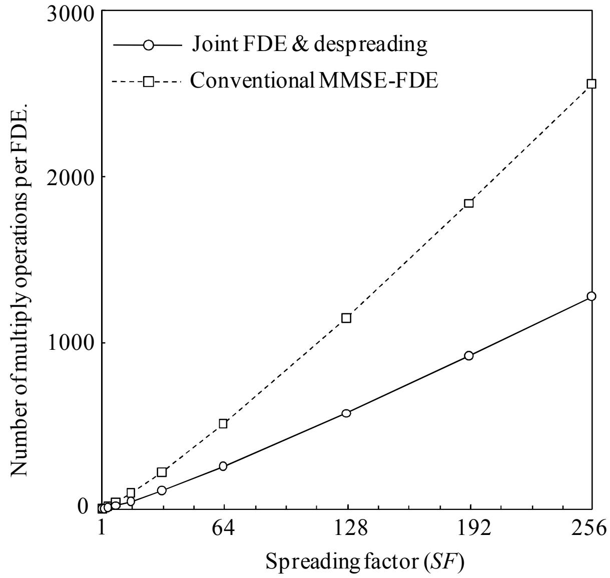


Figure 7. The number of complex multiply operations.

V. CONCLUSION

In this paper, we proposed a joint FDE and despreading suitable for CDTD which simultaneously performs equalization and despreading in the frequency-domain. We evaluated the average BER performance and compared it with the conventional MMSE-FDE. We showed that, in the case of single-code transmission, the joint FDE and despreading does not produce the ICI at all and hence provides better BER performance than the conventional MMSE-FDE. We also showed that the joint FDE and despreading has a lower computational complexity without BER performance degradation.

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