Frequency-domain Multi-stage Inter-code Interference Cancellation for Multi-code DSSS Transmission

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***Abstract*— Frequency-domain equalization (FDE) based on minimum mean square error (MMSE) criterion can replace rake combining with much improved bit error rate (BER) performance of the multi-code direct sequence spread spectrum (DSSS) signal transmission in a severe frequency-selective fading channel. However, since the channel cannot be completely equalized, the residual inter-code interference (ICI) is present and therefore, the BER performance degrades as the number of parallel codes increases. In this paper, we propose frequency-domain multi-stage parallel interference cancellation (FD-MS-PIC). In FD-MS-PIC, interference replica is generated and subtracted from the received signal in the frequency-domain and then, MMSE-FDE is carried out. The achievable BER performance is evaluated by computer simulation.**

Keywords-component; *Direct sequence spread spectrum (DSSS), Frequency-domain equalization (FDE), ICI, PIC*

I. INTRODUCTION

In the next generation mobile communication systems, much higher rate data transmission (e.g., higher than several 10Mbps) than present systems is required. For such high-speed data transmission, the channel becomes severely frequency-selective [1] and the bit error rate (BER) performance of the multi-code direct sequence spread spectrum (DSSS) signal transmission significantly degrades due to large inter-code interference (ICI) even if coherent rake combining is used. Recently, orthogonal frequency division multiplexing (OFDM) and multi-carrier code division multiple access (MC-CDMA) [2], [3] have been attracting much attention. More recently, it has been shown [4], [5] that frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion can replace rake combining with much improved BER performance of the DSSS signal transmission in a severe frequency-selective fading channel. The computational complexity of FDE does not depend on the degree of channel frequency-selectivity (on the other hand, rake combiner requires a larger number of fingers (or correlators) as the channel selectivity becomes severer).

Multi-code DSSS has flexibility to offer various data rate services by changing the number of parallel orthogonal spreading codes [6]. However, even if MMSE-FDE is used, since the channel cannot be completely equalized, the residual ICI is present and the BER performance gradually degrades as the number of parallel codes increases [7].

In this paper, we consider DSSS with MMSE-FDE and propose frequency-domain multi-stage parallel interference cancellation (FD-MS-PIC). In FD-MS-PIC, interference replica is generated and subtracted from the received signal in the frequency-domain and then, MMSE-FDE is carried out. The remainder of this paper is organized as follows. The multi-code DSSS signal transmission system model is presented in Sect. II. FD-MS-PIC is presented in Sect. III. In Sect. IV, the BER performance of multi-code DSSS signal transmission in a frequency-selective Rayleigh fading channel is evaluated by the computer simulation. The paper is concluded in Sect. V.

II. TRANSMISSION SYSTEM MODEL

Transmitter/receiver structure for the multi-code DSSS is illustrated in Fig. 1. Throughout the paper, the chip-spaced discrete time representation is used.

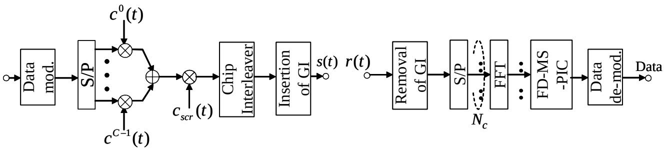
We consider the transmission of one block of N_c chips, where N_c denotes the block length of fast Fourier transform (FFT). At a transmitter, a binary data sequence is transformed into data-modulated symbol sequence and then converted to the C parallel symbol streams by serial-to-parallel (S/P) conversion. Then the qth symbol stream $d^q(n)$, n=0~(N_c/SF−1), is spread using an orthogonal spreading code $c^q_{ort}(t)$, t=0~(SF−1). The C parallel chip streams are added and multiplied by a scramble sequence $c_{scr}(t)$. The resulting multi-code DSSS signal $s(t)$, t=0~(N_c−1), can be expressed using the equivalent baseband representation as

$$s(t)=\sum_{q=0}^{C-1} d^q\left(\left\lfloor \frac{t}{SF}\right\rfloor\right)\cdot c^q(t)\ , \quad (1)$$

where

$$c^q(t)=\sqrt{\frac{2E_c}{T_c}}c^q_{ort}(t \bmod SF)\cdot c_{scr}(t)\ , \quad (2)$$

E_c and T_c are the chip energy and the chip length, respectively, SF is the spreading factor and $\lfloor x \rfloor$ represents the largest integer smaller than or equal to x. An (SF x N_c)-chip block interleaver, as shown in Fig 2, is used in order to reduce the effect of error propagation due to decision feedback for the interference replica generation. 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 the block.



(a) Transmitter (b) Receiver

Fig. 1 Transmitter/receiver structure.

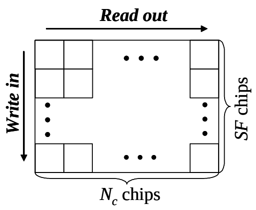


Fig. 2 Chip interleaver.

Assuming that the channel has L independent propagation paths with chip-spaced distinct time delays $\{\tau_l; l = 0 \sim (L-1)\}$, the discrete-time impulse response $h(t)$ of the channel can be expressed as

$$h(t) = \sum_{l=0}^{L-1} h_l \delta(t - \tau_l), \quad (3)$$

where h_l is the lth path gain with $\sum_{l=0}^{L-1} E[|h_l|^2] = 1$. The received signal $r(t)$ can be expressed as

$$r(t) = \sum_{l=0}^{L-1} h_l s(t - \tau_l) + \eta(t), \quad (4)$$

where $\eta(t)$ is the zero-mean noise process having variance $2N_0/T_c$ with N_0 representing the single-sided power spectrum density of the additive white Gaussian noise (AWGN). Here, we assume block fading, where path gains remain constant over the time interval of $t= -N_g \sim (N_c-1)$.

After the removal of GI, the received signal is decomposed into N_c frequency components $\{R(k); k = 0 \sim (N_c - 1)\}$ by applying N_c-point FFT. $R(k)$ is given by

$$R(k) = \sum_{t=0}^{N_c-1} r(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) = H(k)S(k) + \Pi(k), \quad (5)$$

where $S(k)$ is the kth frequency component of $s(t)$ and $H(k)$ and $\Pi(k)$ are respectively the channel gain and the noise component due to the AWGN at the kth frequency. $S(k)$, $H(k)$ and $\Pi(k)$ are given by

$$\begin{cases} S(k) = \sum_{t=0}^{N_c-1} s(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ H(k) = \sum_{l=0}^{L-1} h_l \exp\left(-j2\pi k \frac{\tau_l}{N_c}\right) \\ \Pi(k) = \sum_{t=0}^{N_c-1} \eta(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \end{cases}. \quad (6)$$

Then, FD-MS-PIC is performed to obtain a sequence of the decision variables $\{\tilde{d}_{i-1}^q(n); n = 0 \sim (N_c/SF - 1), q = 0 \sim (C-1)\}$.

III. FD-MS-PIC

The ith stage cancellation structure of FD-MS-PIC is illustrated in Fig 3.

A. PIC

The soft symbol replica $\bar{d}_{i-1}^q(n)$ for the qth parallel symbol stream is generated for PIC operation. The replica generation is as follows. At first, the log-likelihood ratio (LLR) $\lambda_{i-1,b}^q(n)$ of the bth bit in the nth symbol is computed using the decision variable $\tilde{d}_{i-1}^q(n)$:

$$\lambda_{i-1,b}^q(n) = \frac{1}{2\sigma_{i-1}^2}\left(\left| \tilde{d}_{i-1}^q(n) - \sqrt{\frac{2E_c}{T_c}} \left(\frac{1}{N_c} \sum_{k=0}^{N_c-1} \hat{H}_{i-1}^q(k) \right) \hat{d}_{0,\min} \right|^2 - \left| \tilde{d}_{i-1}^q(n) - \sqrt{\frac{2E_c}{T_c}} \left(\frac{1}{N_c} \sum_{k=0}^{N_c-1} \hat{H}_{i-1}^q(k) \right) \hat{d}_{1,\min} \right|^2 \right), \quad (7)$$

where $\hat{d}_{0,\min}$ ($\hat{d}_{1,\min}$) is the candidate symbol with 0 (or 1) in the bth bit position, for which the Euclidean distance from $\tilde{d}_{i-1}^q(n)$ is minimum. σ_{i-1}^2 is the variance of the interference plus noise component and $\hat{H}_{i-1}^q(k)$ is the equivalent channel gain, given by [5]

$$\begin{cases} \sigma_{i-1}^2 = \frac{1}{SF} \cdot \frac{E_c}{T_c} \left[\begin{array}{l} \frac{1}{N_c} \left(C - \frac{1}{N_c} \sum_{\substack{q'=0 \\ \neq q}}^{C-1} \sum_{\tau=0}^{N_c-1} \left| \bar{s}_{i-2}^{q'}(\tau) \right|^2 \right) \sum_{k=0}^{N_c-1} \left| \hat{H}_{i-1}^q(k) \right|^2 \\ - C \left| \frac{1}{N_c} \sum_{k=0}^{N_c-1} \hat{H}_{i-1}^q(k) \right|^2 \\ + \left(\frac{E_c}{N_0} \right)^{-1} \left(\frac{1}{N_c} \sum_{k=0}^{N_c-1} |w_{i-1}^q(k)|^2 \right) \end{array} \right] \\ \hat{H}_{i-1}^q(k) = H(k) w_{i-1}^{q*}(k) \end{cases} \quad (8)$$

with $w_i^q(k)$ denoting the MMSE weight (which is derived in Sect. III. *C*).

The soft symbol replica $\bar{d}_{i-1}^q(n)$ is obtained as follows [8]:

$$\bar{d}_{i-1}^q(n) = \frac{1}{\sqrt{2}} \left[\tanh\left(\frac{\lambda_{i-1,0}^q(n)}{2} \right) + j \tanh\left(\frac{\lambda_{i-1,1}^q(n)}{2} \right) \right]. \quad (9)$$

The sequence of $\bar{d}_{i-1}^{q}(n)$ is re-spread and then decomposed into N_c frequency components $\{\bar{S}_{i-1}^{q}(k); k=0\sim(N_c-1)\}$ by applying N_c-point FFT:

$$\bar{S}_{i-1}^{q}(k)=\sum_{t=0}^{N_c-1}\left\{\sqrt{\frac{2E_c}{T_c}}\bar{d}_{i-1}^{q}\left(\left\lfloor\frac{t}{SF}\right\rfloor\right)\cdot c^{q}(t)\right\}\exp\left(-j2\pi k\frac{t}{N_c}\right). \quad (10)$$

The received signal replica $\bar{R}_{i-1}^{q}(k)$ of the qth parallel chip stream in the frequency-domain is obtained from

$$\bar{R}_{i-1}^{q}(k)=H(k)\bar{S}_{i-1}^{q}(k). \quad (11)$$

Finally, PIC is carried out as

$$\tilde{R}_{i}^{q}(k)=R(k)-\sum_{\substack{q'=0\\ \neq q}}^{C-1}\bar{R}_{i-1}^{q'}(k)=R(k)-\left[\sum_{q'=0}^{C-1}\bar{R}_{i-1}^{q'}(k)-\bar{R}_{i-1}^{q}(k)\right]. \quad (12)$$

B. MMSE-FDE and Despreading

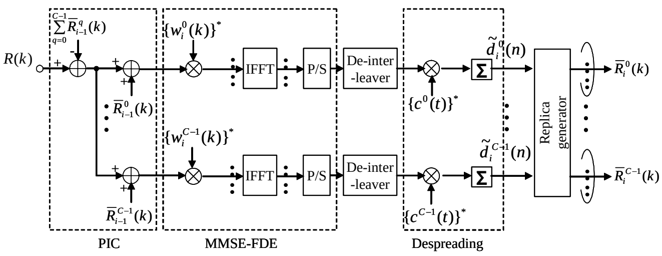
MMSE-FDE is carried out as follows:

$$\tilde{S}_{i}^{q}(k)=\left\{w_i^{q}(k)\right\}^{*}\tilde{R}_{i}^{q}(k). \quad (13)$$

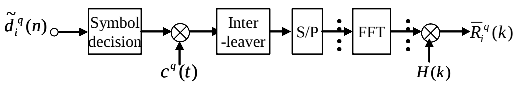
Then, after carrying out MMSE-FDE, N_c-point inverse FFT (IFFT) is performed on $\{\tilde{S}_i^{q}(k); k=0\sim(N_c-1)\}$ to produce the qth parallel chip stream $\tilde{s}_i^{q}(t)$ and despreding is carried out to obtain the decision variable $\tilde{d}_i^{q}(n)$:

$$\tilde{d}_i^{q}(n)=\frac{1}{SF}\sum_{t=nSF}^{(n+1)SF-1}\tilde{s}_i^{q}(t)\cdot\left\{c^{q}(t)\right\}^{*}. \quad (14)$$

The above procedure is carried out for all C parallel chip streams.



(a) Overall structure



(b) Replica generator

Fig. 3 FD-MS-PIC block.

C. Derivartion of MMSE Weight

The equalization error $e(k)$ at the ith stage is defined as follows:

$$e(k)=S^{q}(k)-\left\{w_i^{q}(k)\right\}^{*}\tilde{R}_i^{q}(k), \quad (15)$$

where

$$S^{q}(k)=\sum_{t=0}^{N_c-1}\left\{\sqrt{\frac{2E_c}{T_c}}d^{q}\left(\left\lfloor\frac{t}{SF}\right\rfloor\right)\cdot c^{q}(t)\right\}\exp\left(-j2\pi k\frac{t}{N_c}\right) \quad (16)$$

is the kth frequency component of the qth chip stream and $w_i^{q}(k)$ is the MMSE weight which minimizes the mean square error (MSE) $E[|e(k)|^2]$, given by, from [9]

$$w_i^{q}(k)=\left(X_i^{q}(k)\right)^{-1}p_i^{q}(k), \quad (17)$$

where

$$\begin{cases} X_i^{q}(k)=E[|\tilde{R}_i^{q}(k)|^2] \\ p_i^{q}(k)=E[\tilde{R}_i^{q}(k)S^{q*}(k)]=\dfrac{2E_c}{T_c}N_cH(k) \end{cases}. \quad (18)$$

$w_i^{q}(k)$ is derived below. Substitution of Eqs. (10) and (12) into Eq. (18) gives

$$X_i^{q}(k)\approx E[|H(k)|^2\sum_{\tau=0}^{N_c-1}|s(\tau)-\sum_{\substack{q'=0\\ \neq q}}^{C-1}\bar{s}_{i-1}^{q'}(\tau)|^2]+E[|\Pi(k)|^2], \quad (19)$$

where

$$\bar{s}_{i-1}^{q'}(t)=\sqrt{\frac{2E_c}{T_c}}\bar{d}_{i-1}^{q'}\left(\left\lfloor\frac{t}{SF}\right\rfloor\right)\cdot c^{q'}(t) \quad (20)$$

is the replica of the qth chip stream in the time-domain. Since the scramble sequence is used, the DSSS signal is white-noise like and hence,

$$\begin{aligned} &E\left[\left(s(\tau)-\sum_{\substack{q'=0\\ \neq q}}^{C-1}\bar{s}_{i-1}^{q'}(\tau)\right)\left(s(\tau')-\sum_{\substack{q''=0\\ \neq q}}^{C-1}\bar{s}_{i-1}^{q''}(\tau')\right)^{*}\right] \\ &=1-2E\left[\mathrm{Re}\left\{s(\tau)\sum_{\substack{q'=0\\ \neq q}}^{C-1}\bar{s}_{i-1}^{q'}(\tau')\right\}\right]+\sum_{\substack{q'=0\\ \neq q}}^{C-1}\left|\bar{s}_{i-1}^{q'}(\tau')\right|^2 \end{aligned}. \quad (21)$$

Let $\bar{s}_{i-1}^{q'}(\tau)$ be expressed using $s^{q'}(\tau)$ and the replica error $\Delta_{i-1}^{q'}(\tau)$ with $E[\Delta_{i-1}^{q'}(t)]=0$ as

$$\bar{s}^{q'}(\tau)=s_{i-1}^{q'}(\tau)+\Delta_{i-1}^{q'}(\tau). \quad (22)$$

We obtain

$$X_i^q(k) \approx \frac{2E_c}{T_c} N_c \left[\left\{ |H(k)|^2 + \sum_{\substack{q'=0 \\ \neq q}}^{C-1} \rho_{i-1}^{q'} |H(k)|^2 \right\} + \left(\frac{E_c}{N_0} \right)^{-1} \right], \quad (23)$$

where

$$\rho_{i-1}^{q'} = \frac{1}{N_c} \sum_{t=0}^{N_c-1} \left[1 - \left(\frac{2E_c}{T_c} \right)^{-1} \left| \bar{s}_{i-1}^{q'}(t) \right|^2 \right]. \quad (24)$$

Finally, from Eqs. (17), (18) and (23), we have

$$w_i^q(k) = \frac{H(k)}{|H(k)|^2 \left[1 + \sum_{\substack{q'=0 \\ \neq q}}^{C-1} \rho_{i-1}^{q'} \right] + \left(\frac{E_c}{N_0} \right)^{-1}}. \quad (25)$$

IV. COMPUTER SIMULATION

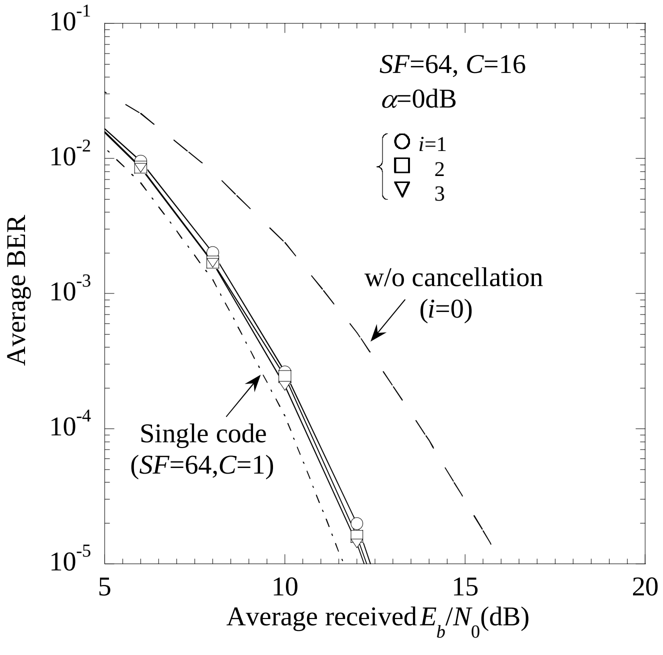
Table 1 shows the computer simulation condition. A chip-spaced 16-path (L=16) frequency-selective block Rayleigh fading channel having an exponential power delay profile with decay factor α (i.e., $h_l = h_0 \alpha^l$) is assumed. We assume an FFT block length of N_c=256 chips, a GI length of N_g=32 chips, and ideal channel estimation.

Table 1 Simulation condition.

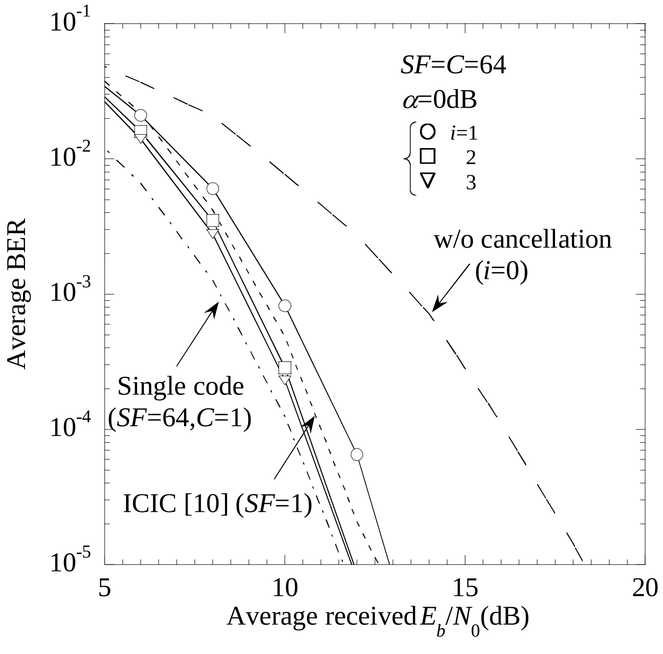
Transmitter	Modulation	QPSK
	FFT block length	N_c=256
	GI length	N_g=32 chips
	Scrambling code	Long PN sequence
	Spreading code	Walsh-Hadamard code
	Spreading factor	SF=64
	Number of codes	C=1~64
Channel	Fading	Frequency-selective block Rayleigh fading
	Power delay profile	L=16-path exponential power delay profile, decay factor α=0, 4, 8 (dB)
Receiver	Frequency-domain equalization	MMSE
	Channel estimation	Ideal

Fig. 4 shows the average BER performance of multi-code DSSS using FD-MS-PIC as a function of the average received signal energy per bit-to-the AWGN power spectrum density ratio E_b/N_0 (=0.5SF(1+N_g/N_c)(E_c/N_0)). For comparison, the result for the single-code case (C=1) is also plotted. Without cancellation (i.e., i=0), the BER performance severely degrades. However, it can be seen that, even with one stage, the proposed FD-MS-PIC can significantly improve the BER performance. For C/SF=0.5 and the full multiplexing (C=SF=64) case, the required E_b/N_0 for BER=10^{-4} approaches the single-code case by about a fractional dB at the 2nd stage. For comparison, the BER performance of nonspread single-carrier (SC) transmission (SF=1) using inter-chip interference cancellation (ICIC) is plotted in Fig. 4(b) [10]. It can be seen that the FD-MS-PIC provides better performance than SC with ICIC. This is because the interference is suppressed more effectively by the despreading operation.

Fig. 5 shows the average BER performance with FD-MS-PIC as a function of the average received E_b/N_0 with α as a parameter for the full load case (C/SF=1). The BER performance of FD-MS-PIC is superior to that of no FD-MS-PIC. When α=4 and 8dB, the BER performances of FD-MS-PIC can reduce the required E_b/N_0 by about 3.2 and 2.4dB for achieving a BER=10^{-3}.



(a) C=16



(b) C=64

Fig. 4 Average BER performance.

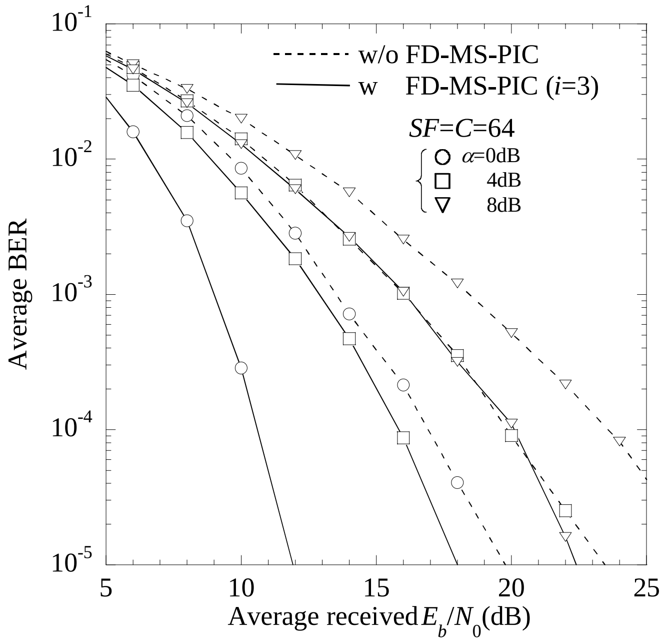


Fig. 5 Impact of decay factor α.

V. CONCLUSION

In this paper, we proposed frequency-domain multi-stage parallel interference cancellation (FD-MS-PIC) for multi-code DSSS signal transmission. The MMSE weight is updated in each cancellation stage taking into account the residual ICI. In FD-MS-PIC, the soft interference replica is generated using LLR. The BER performance with FD-MS-PIC in a frequency-selective Rayleigh fading was evaluated by computer simulation. It was shown that, even for fully loaded case (i.e., C/SF=1), the BER performance at the 2nd stage gets close to the single-code performance and is better than the SC transmission (SF=1).

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