LETTER

# Frequency-Domain Pre-Equalization for Multicode Direct Sequence Spread Spectrum Signal Transmission

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**SUMMARY** Severe frequency-selective fading, encountered in a broadband wireless mobile communication, significantly degrades the bit error rate (BER) performance of direct sequence spread spectrum (DSSS) signal transmission with rake combining. In this paper, frequency-domain pre-equalization transmission, called pre-FDE transmission, is presented for orthogonal multicode DSSS signal transmission. It is confirmed by the computer simulation that pre-FDE transmission can achieve a BER performance almost identical to that attainable by FDE reception.

***key words:*** *pre-equalization, frequency-domain equalization, direct sequence spread spectrum*

## 1. Introduction

Demands for broadband services in mobile wireless communications systems are becoming higher and higher. Next generation mobile wireless communications systems, called 4th generation (4G) systems, require very high-speed data transmissions capability of more than few tens of Mbps [1]. For such high speed data transmissions, wireless channel is composed of many propagation paths with different time delays and is severely frequency-selective. Direct sequence spread spectrum (DSSS) or direct sequence code division multi-access (DS-CDMA) technique improves the bit error rate (BER) performance by exploiting the channel frequency-selectivity due to the well-known rake combining [2]. High-speed data transmissions can be achieved by reducing the spreading factor for the given chip rate. However, the use of small spreading factor cannot sufficiently reduce the inter-path interference (IPI) produced by asynchronism of different propagation paths and hence, the BER performance degrades. Therefore, much attention has been paid to multicarrier (MC) technique, known as OFDM and MC-CDMA [3], [4], which uses frequency-domain spreading and despreading to exploit the channel frequency-selectivity by one-tap frequency-domain equalization (FDE). Recently, it has been shown [5], [6] that the use of FDE can improve the DSSS transmission performance as well. Since then, the DSSS technique has been looked over again as a strong candidate for high-speed data transmissions.

Recently, it has been shown [7], [8] that in MC-CDMA, frequency-domain pre-equalization transmission, called pre-FDE transmission, can be used instead of FDE reception. Meanwhile, we have shown [9] that rake combining in DSSS signal transmissions is equivalent to pre-FDE using maximal ratio (MR) weight. Therefore, in a severe frequency-selective channel, the BER performance with pre-FDE using MR weight degrades similar to rake combining. Recently, pre-FDE based on minimum mean square error (MMSE) criterion was proposed for single-carrier (SC) transmissions [10]. In this paper, we consider orthogonal multicode DSSS signal transmission, where spreading and orthogonal multiplexing are employed to achieve variable rate data transmission for the given spreading chip rate [11]. We extend the theoretical analysis presented in [10] and derive pre-FDE MMSE weight. The BER performance achievable with the pre-FDE transmission is evaluated by computer simulation and compared with that of FDE reception.

## 2. Pre-FDE Transmission

At the one transmit/receive side, preferably the base station, both pre-FDE transmission and FDE reception functions are implemented. Simple despreading and spreading functions are required at the other receive/transmit side. Transmitter and receiver structures for multicode DSSS with time division duplex (TDD) system with pre-FDE transmission are illustrated in Fig. 1. Pre-FDE requires fast Fourier transform (FFT) and inverse FFT (IFFT) operations at a transmitter. Since the same carrier frequency is used for both transmit and receive channels in the TDD system, the channel estimate for FDE reception can be reused for pre-FDE transmission. In this paper, we assume ideal channel estimation.

A sequence of data-modulated symbols $\{d(i);\ i = 0 \sim N_c(U/SF) - 1\}$ with $|d(i)|=1$ is to be transmitted, where $N_c$ is the FFT/IFFT block size in chip, $U$ is the code multiplexed order and $SF$ is the spreading factor. The multicode DSSS signal at the $t$-th chip time instance can be expressed, using the equivalent lowpass representation, as

$$s(t) = c_{scr}(t) \sum_{u=0}^{U-1} c_u(t \bmod SF) d\left(u + U \lfloor t/SF \rfloor\right) \qquad (1)$$

for $t = 0 \sim (N_c - 1)$, where $c_u(t)$ and $c_{scr}(t)$ are the $u$-th orthogonal spreading code and the scramble code sequence, respectively, with unity amplitude and $\lfloor x \rfloor$ denotes the largest integer smaller than or equal to $x$. For pre-FDE transmission, $N_c$-point FFT is applied to decompose the spread signal $s(t)$ into $N_c$ frequency components $\{S(n);\ n = 0 \sim (N_c - 1)\}$. After multiplying the complex-valued

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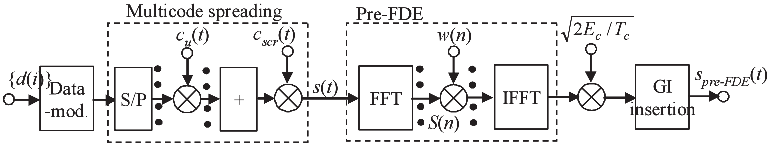
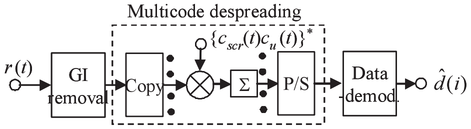


(a) transmitter



(b) receiver

**Fig. 1** Transmitter and receiver structures for pre-FDE transmission.

weights $\{w(n); n = 0 \sim (N_c - 1)\}$, $N_c$-point IFFT is applied to obtain the pre-equalized multicode DSSS signal $s_{pre-FDE}(t)$, which can be expressed as

$$s_{pre-FDE}(t) = \frac{C}{N_c}\sum_{n=0}^{N_c-1}\{w(n)S(n)\}\exp\left(j2\pi t\frac{n}{N_c}\right) \tag{2}$$

for $t = 0 \sim (N_c - 1)$, where $S(n)$ is given by

$$S(n) = \sqrt{\frac{2E_c}{T_c}}\sum_{t=0}^{N_c-1} s(t)\exp\left(-j2\pi n\frac{t}{N_c}\right) \tag{3}$$

with $E[|S(n)|^2] = (2E_c/T_c)U \cdot N_c$, where $E_c$ denotes the transmit chip energy per spreading code and $T_c$ is the chip period. $C$ is the transmit power constraint factor so as to keep the average transmit powers after and before pre-FDE the same and is given by

$$C = N_c\sqrt{\frac{(2E_c/T_c)U}{\sum_{n=0}^{N_c-1}|w(n)S(n)|^2}}. \tag{4}$$

The channel gain at the $n$-th frequency is denoted by $H(n)$. The pre-FDE MR weight $w(n)$ that maximizes the received signal-to-noise power ratio (SNR) at the receiver is $w(n) = H^*(n)$; but this enhances the channel frequency-selectivity seen at the receiver and accordingly, increases the IPI. On the other hand, the use of zero forcing (ZF) weight, i.e., $w(n) = H^*(n)/|H(n)|^2$, can restore the frequency-nonselective channel and hence, remove the IPI. But under the transmit power constraint, most of the transmit power is allocated to frequency components which experience deep fade. Hence, the received signal power averaged over $N_c$ frequencies at the receiver is significantly reduced, thereby degrading the BER performance due to the additive white Gaussian noise (AWGN).

From Eq. (2), the $n$-th frequency component of the transmitted multicode DSSS signal is $C \cdot S(n)$. The $n$-th frequency component $R(n)$ of the received signal is given by

$$R(n) = C\{w(n)S(n)\}H(n) + \Pi(n), \tag{5}$$

where $\Pi(n)$ represents the noise component due to the AWGN having the single-sided power spectrum density $N_0$. Since the use of pre-FDE alters the transmitted signal spectrum shape and the received signal power in a frequency selective fading channel, we introduce the following relative equalization error:

$$e(n) = \frac{R(n) - C \cdot S(n)}{C\sqrt{E[|S(n)|^2]}}. \tag{6}$$

The BER performance achievable with pre-FDE transmission depends on the sum of mean square relative equalization errors:

$$e^2 = \sum_{n=0}^{N_c-1} E[|e(n)|^2], \tag{7}$$

since the same data symbol is spread over all frequencies. We will find the set of MMSE weights $\{w(n)\}$ that minimizes Eq. (7). The MMSE weight $w(n)$ should satisfy

$$\partial e^2/\partial w(n) = 0 \quad \text{for} \quad n = 0 \sim (N_c - 1). \tag{8}$$

After some manipulation, we obtain the following MMSE weight for multicode DSSS (its derivation is omitted for the sake of brevity):

$$w(n) = \frac{H^*(n)}{|H(n)|^2 + (U/SF)^{-1}(E_s/N_0)^{-1}}, \tag{9}$$

where $E_s = E_c SF$ is the symbol energy. Eq. (9) with $SF$=1 (and hence $U$=1) gives the pre-FDE MMSE weight for the SC transmission derived in [10]. Assuming a chip-spaced $L$-path frequency-selective fading channel and denoting the complex-valued path gain and time delay of the $l$-th propagation path by $h_l$ and $\tau_l$ (in chip), respectively, $H(n)$ can be expressed as

$$H(n) = \sum_{l=0}^{L-1} h_l \exp\left(-j2\pi n\frac{\tau_l}{N_c}\right). \tag{10}$$

For simplicity, we have assumed a block fading, where the path gains remain constant over several $N_c$-chip blocks.

When pre-FDE transmission is applied, a cyclic prefix of $N_g$ chips needs to be inserted into the guard interval (GI), as shown in Fig. 2. The GI-inserted signal $\{s_{pre-FDE}(t); t = -N_g \sim (N_c - 1)\}$ is transmitted over a frequency-selective fading channel. The received signal can be represented as

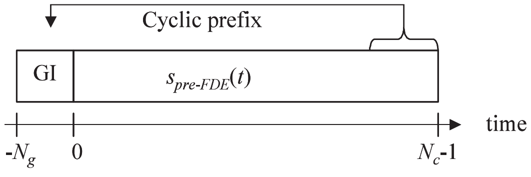


**Fig. 2** GI insertion.

$$r(t) = \sum_{l=0}^{L-1} h_l s_{pre-FDE}(t - \tau_l) + \eta(t), \tag{11}$$

where $\eta(t)$ is the zero-mean noise process due to the AWGN. At the receiver, after removal of GI, simple despreading is applied to obtain the decision variable:

$$\hat{r}(u + kU) = \frac{1}{SF} \sum_{t=kSF}^{(k+1)SF-1} r(t)\{c_{scr}(t)c_u(t)\}^* \tag{12}$$

for $u = 0 \sim (U - 1)$ and $k = 0 \sim (N_c/SF - 1)$, from which the received $i$ $(= u + kU)$-th symbol $\hat{d}(i)$ is recovered by data-demodulation.

## 3. Computer Simulation

We assume the Walsh codes for orthogonal spreading, an M-sequence of 4095 chips for scrambling, $N_c$=256, $N_g$=32, coherent quadrature phase shift keying (QPSK) data-modulation, and a chip-spaced 16-path ($L$=16) block Rayleigh fading channel with uniform power delay profile (i.e., the ensemble average of $|h_l|^2$ is $1/L$ for all $l$). Channel estimation for pre-FDE transmission is assumed to be ideal.

First, we compare the BER performances achievable with pre-FDE using MMSE, MR, and ZF weights. The simulation results are plotted for the code multiplex order $U$=1 in Fig. 3 as a function of the average transmit signal energy per bit-to-AWGN power spectrum density ratio $E_b/N_0$ (=0.5 $(E_s/N_0)(1 + N_g/N_c)$ for QPSK). As was expected, it is seen that the MMSE weight gives the best BER performance irrespective of $SF$. As $SF$ increases, the achievable BER performance improves since the residual IPI (which is produced due to imperfect channel equalization) can be better suppressed by increasing the value of $SF$ (or at the cost of increased bandwidth or lowering the data rate). The MR weight gives a very poor BER performance when $SF$=1. However, it gives almost the same BER performance as the MMSE weight for a large spreading factor (e.g., $SF$=64). This is because the increased IPI due to enhanced channel selectivity can be sufficiently suppressed by the despreading process. On the other hand, the ZF weight gives the same BER performance irrespective of $SF$, but the BER performance is very poor compared to that of the MMSE weight.

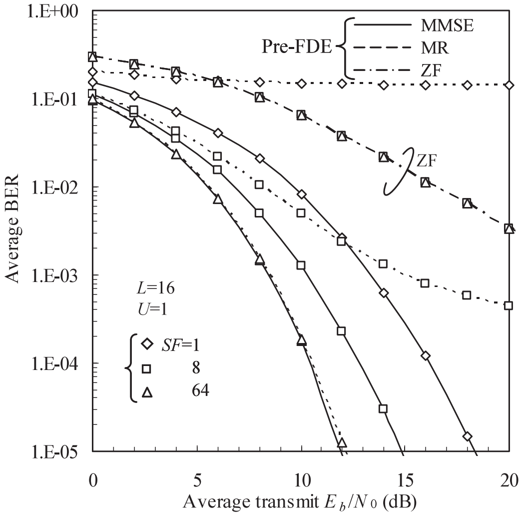


**Fig. 3** Performance comparison of pre-FDE using MMSE, MR, and ZF weights.

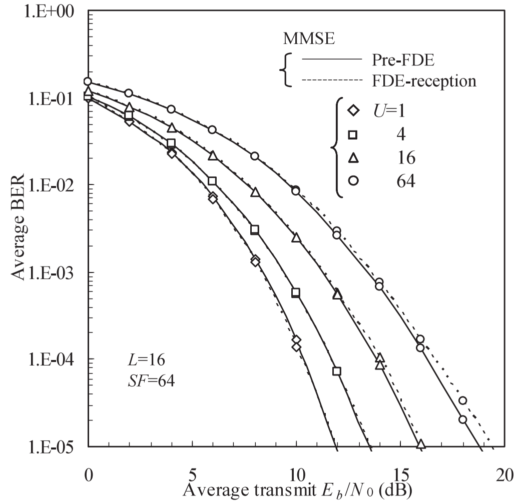


**Fig. 4** Performance comparison between pre-FDE transmission and FDE reception.

Next, we compare the pre-FDE transmission and FDE reception both using MMSE weight. When FDE reception is used, FFT/IFFT operation and MMSE equalization are required at a receiver [5], [6]. The BER performances of pre-FDE transmission and FDE reception are compared in Fig. 4 for various values of code multiplex order $U$. It is seen that pre-FDE transmission and FDE reception provide almost identical performance. This clearly shows the potential of pre-FDE transmission.

## 4. Conclusion

Pre-FDE transmission was presented for multicode DSSS signal transmission and the MMSE weight was derived. The achievable BER performance in a frequency-selective Rayleigh fading channel was evaluated by computer simulation to show that pre-FDE transmission using MMSE weight can achieve a performance almost identical to the MMSE-FDE reception irrespective of the spreading factor. Therefore, for a DSSS/TDD mobile communications system, all frequency-domain processing function can be implemented at the base stations; pre-FDE transmission for the downlink (base-to-mobile) and FDE reception for the uplink (mobile-to-base). This allows simple structure of the mobile stations.

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