PAPER

Frequency-Domain Iterative Parallel Interference Cancellation for Multicode Spread-Spectrum MIMO Multiplexing

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SUMMARY Very high-speed data services are demanded in the next generation wireless systems. However, the available bandwidth is limited. The use of multi-input multi-output (MIMO) multiplexing can increase the transmission rate without bandwidth expansion. For high-speed data transmission, however, the channel becomes severely frequency-selective and the achievable bit error rate (BER) performance degrades. In our previous work, we proposed the joint use of iterative frequency-domain parallel interference cancellation (PIC) and two-dimensional (2D) MMSE-FDE for the non-spread single-carrier (SC) transmission in a frequency-selective fading channel. The joint use of PIC and 2D MMSE-FDE can effectively suppress the inter-path interference (IPI) and the inter-code interference (ICI), resulting from the channel frequency-selectivity, and the interference from other antennas simultaneously. An iterative PIC with 2D MMSE-FDE has a high computational complexity. In this paper, to well suppress the interference from other antennas while reducing the computational complexity, we propose to replace 2D MMSE-FDE by 1D MMSE-FDE except for the initial iteration stage and to use multicode spread-spectrum (SS) transmission instead of the non-spread SC transmission. The BER performance of the proposed scheme in a frequency-selective Rayleigh fading channel is evaluated by computer simulation to show that the proposed scheme can basically match the BER performance of 2D MMSE-FDE with lower complexity.

key words: spread-spectrum, MIMO multiplexing, frequency-domain iterative PIC, mobile communication

1. Introduction

Very high-speed data services of e.g. 100 M-1 Gbps are demanded in the next generation wireless systems [1]. However, the available bandwidth is limited. Furthermore, for such high-speed data transmissions, the channel becomes severely frequency-selective [2], [3]. Therefore, the development of spectrum efficient transmission techniques which are robust against the channel frequency-selectivity is required. One of the promising techniques is the multi-input multi-output (MIMO) multiplexing [4]. MIMO multiplexing transmits different data sequences in parallel using different transmit antennas with the same carrier frequency. At a receiver, it is necessary to separate the signals transmitted from different antennas. A lot of research attention has been paid to develop an efficient signal separation scheme. Some well-known signal separation schemes are maximum likelihood detection (MLD) [2], zero forcing (ZF) detection [2], minimum mean square error (MMSE) detection [2]

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and Vertical-bell laboratories layered space-time architecture (V-BLAST) [5].

In our previous work, we proposed an iterative frequency-domain parallel interference cancellation (PIC) with two-dimensional (2D) MMSE-FDE [6], which can achieve the space and frequency diversity gains, for the nonspread single-carrier (SC) transmission (hereafter, we simply call this the SC transmission) in a frequency-selective fading channel. The idea of iterative PIC was presented in [7] for turbo-coded SC MIMO multiplexing in a frequencynonselective fading channel; a sequence of the soft decision variables for turbo decoding is obtained by the use of PIC and antenna diversity using maximal-ratio combining (MRC). In [6], joint equalization and signal separation is simultaneously performed by applying 2D MMSE-FDE. Since a single use of 2D MMSE-FDE can not well suppress the interference from other antennas, joint PIC and 2D MMSE-FDE is repeated a sufficient number of times. However, iterative PIC with 2D MMSE-FDE has a high computational complexity. In this paper, to well suppress the interference from other antennas while reducing the computational complexity, we propose to replace 2D MMSE-FDE by 1D MMSE-FDE except for the initial iteration stage and to use multicode spread-spectrum (SS) transmission instead of the SC transmission.

The principle of the frequency-domain iterative PIC is similar to the conventional frequency-domain turbo equalization [8], which also performs the joint use of 2D MMSE-FDE and PIC iteratively. However, there are three new technical contributions in this paper: (1) the 2D MMSE weight with the inter-code interference (ICI) taken into account is derived theoretically. (2) By the joint use of PIC and MMSE-FDE weight with ICI taken into account, the BER performance with four iterations can be close to the BER performance of single-input multi-output (SIMO) transmission. (3) For the equivalent spreading factor SF_{eq} (=SF/C)=1, where SF is the spreading factor and C is codemultiplexing order, the full code-multiplexed SS transmission provides better performance than the SC transmission (i.e., SF=1).

The remainder of this paper is organized as follows. Section 2 describes the frequency-domain iterative PIC with 1D MMSE-FDE in multicode SS transmission. Section 3 presents the computer simulated BER performance in a frequency-selective Rayleigh fading channel and discusses the performance improvement achieved by the proposed scheme. The BER performance comparison of the proposed

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scheme and 2D MMSE-FDE [6], ZF detection, MMSE detection, and V-BLAST is presented. In Sect. 4, the comparison of the computational complexity is presented. Section 5 concludes the paper.

2. Frequency-Domain Iterative PIC Using Multicode SS MIMO Multiplexing

In FDE, the received multicode SS signal is decomposed by FFT into a number of orthogonal frequency components, each being multiplied by the complex equalization weight. MRC, ZF and MMSE weights are well known as for the equalization weight. The MRC weight enhances the frequency-selectivity, thus increasing the inter-path interference (IPI) and ICI resulting from frequency-selectivity. The ZF weight can achieve a frequency-nonselective channel, but degrades the performance due to the noise enhancement. On the other hand, the MMSE weight can reduce the IPI and ICI while suppressing the noise enhancement and hence provides the best BER performance. Therefore, in this paper, MMSE-FDE is considered. However, the BER performance of multicode SS MIMO multiplexing still degrades, if the conventional 2D MMSE-FDE weight for SC transmission is used. It is necessary to use the 2D MMSE-FDE weight taking into account the ICI. In this paper, 2D MMSE-FDE weight for multicode SS MIMO multiplexing is theoretically derived.

In frequency-domain iterative PIC, at the initial stage (i=0), 2D MMSE-FDE is performed. 2D MMSE-FDE attempts to suppress the IPI, ICI and the interference from the signals transmitted from other antennas, but the interference suppression is not sufficient. To achieve a good BER performance while keeping the computational complexity low, joint PIC and 1D MMSE-FDE is repeated a sufficient num-

ber of times after carrying out 2D MMSE-FDE at the initial stage (i=0).

2.1 Transmit and Receive Signals

Figure 1 illustrates the transmitter/receiver structure of multicode SS (N_t, N_r) MIMO multiplexing using frequencydomain iterative PIC. N_t transmit antennas and N_r receive antennas are used. Below, we introduce the codemultiplexing order $C(\leq SF)$; SF_{eq} (=SF/C)=1 is used to maximize the transmission data rate.

(a) Transmit signal

At the transmitter, a binary information sequence is QPSK modulated to get a symbol sequence $\{d(n'); n' = 0 \sim CN_tN_c/SF - 1\}$. Then, the QPSK symbol sequence is converted, by serial/parallel (S/P) conversion, to C parallel streams $\{d_c(n); c = 0 \sim C - 1\}$. Each stream is spread by using one of the SF orthogonal spreading codes, where N_c is the block length in chips for fast Fourier transform (FFT). The resulting C parallel chip streams are codemultiplexed into a single stream and multiplied by a scramble sequence for making the transmit orthogonal multicode SS signal noise-like. The multicode SS signal s(t') can be expressed as

$$s(t') = \left[\sum_{c=0}^{C-1} d_c(\lfloor t'/SF \rfloor)c_c(t' \mod SF)\right]c_{scr}(t'), \qquad (1)$$

where $c_c(t' \mod SF)$ is the *c*th orthogonal code, and $c_{scr}(t')$ is the scramble sequence.

For MIMO multiplexing, s(t') is converted to N_t parallel signal streams { $s_{n_t}(t)$; $n_t=0 \sim N_t - 1$ }, each to be transmitted from a different transmit antenna. Each signal stream is divided into a sequence of blocks of N_c chips each. As





shown in Fig. 2, the last N_g chips in each block are copied and inserted as a cyclic prefix into the guard interval (GI) placed at the beginning of each block. N_t GI-inserted blocks are then transmitted simultaneously from N_t transmit antennas using the same carrier frequency. We consider the transmission of N_t blocks of N_c -chip multicode SS signal from N_t transmit antennas (i.e., the transmission of CN_tN_c/SF QPSK symbols). The multicode SS signal transmitted from the n_t th antenna is represented as

$$\tilde{s}_{n_t}(t) = s_{n_t}(t \mod N_c), \tag{2}$$

for $t=-N_g \sim N_c - 1$. (b) Received signal

 N_t transmitted signals go through different frequencyselective fading channels. Each channel is assumed to be a chip-spaced *L*-path frequency-selective fading channel, each path being subjected to independent fading. Without loss of generality, it is assumed that the *l*th path time delay is τ_l chips with $\tau_0 = 0 < \tau_1 < \tau_2 \cdots < \tau_{L-1} < N_g$. At the receiver, a superposition of N_t transmitted signals is re-

ceived by N_r antennas. The received signal $r_{n_r}(t)$ on the n_r th antenna can be given by $r_{n_r}(t) = \sqrt{2S} \sum_{n=0}^{N_r-1} \sum_{l=0}^{L-1} h_{n_r,n_l,l} \breve{s}_{n_l}(t-\tau_l) + n_{n_r}(t), \quad (3)$

for $t = -N_g \sim N_c - 1$, where *S* is the transmit power per code. $h_{n_r,n_t,l}$ denotes the path gain of the *l*th path between the n_r th receive antenna and n_t th transmit antenna and $\sum_{l=0}^{L-1} E\left[\left|h_{n_r,n_t,l}\right|^2\right] = 1$ for all n_r and n_t . In this paper, block fading is assumed, where the path gains stay constant over at least one chip block. $n_{n_r}(t)$ is a zero-mean complex Gaussian process having variance $2N_0/T_c$ with N_0 being the one-sided power spectrum density of additive white Gaussian noise (AWGN), where T_c is the chip duration.

After the removal of the GI from the received signal, N_c -point fast Fourier transform (FFT) is applied to decompose the GI-removed received signal $r_{n_r}(t)$, $t = 0 \sim N_c - 1$, into N_c frequency components. The *k*th frequency component $R_{n_r}(k)$ can be expressed as

$$R_{n_r}(k) = \sum_{t=0}^{N_c-1} r_{n_r}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right)$$

= $\sqrt{2S} \sum_{n_r=0}^{N_r-1} H_{n_r,n_r}(k) S_{n_r}(k) + \Pi_{n_r}(k),$ (4)

where $H_{n_r,n_t}(k)$, $S_{n_t}(k)$ and $\Pi_{n_r}(k)$ are respectively the complex channel gain between the n_t th transmit antenna and the



Fig. 3 Operation of frequency-domain iterative PIC.

 n_r th receive antenna, the transmit signal component and the noise component at the *k*th frequency. They are given by

$$\begin{cases} H_{n_r,n_t}(k) = \sum_{l=0}^{L-1} h_{n_r,n_t,l} \exp\left(-j2\pi k \frac{\tau_l}{N_c}\right) \\ S_{n_t}(k) = \sum_{t=0}^{N_c-1} s_{n_t}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ \Pi_{n_r}(k) = \sum_{t=0}^{N_c-1} n_{n_r}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \end{cases}$$
(5)

2.2 Frequency-Domain Iterative PIC and MMSE-FDE

The operation principle of the proposed frequency-domain iterative PIC is illustrated in Fig. 3. In each iteration, N_c point IFFT is carried out after MMSE-FDE to obtain the N_t time-domain multicode SS signals, which are then converted to a serial multicode SS signal by parallel/serial (P/S) conversion. De-scrambling and multicode de-spreading are performed on the resulting, multicode SS signal to get a sequence of the decision variables. After soft symbol decision, the replica of transmit multicode SS chip sequence is generated by performing multicode spreading and scrambling. Then, N_c -point FFT is applied to N_t chip block replicas in order to generate the interference replicas in the frequencydomain to perform frequency-domain PIC. After performing PIC, 1D MMSE-FDE is performed again. The above procedure is repeated a sufficient number of times. In what follows, the operation principle is described in detail. (a) 2D MMSE-FDE (i=0)

The n_t th transmitted signal component $\tilde{R}_{n_t}^{(0)}(k)$ at the *k*th frequency after 2D MMSE-FDE can be expressed as

$$\tilde{R}_{n_t}^{(0)}(k) = \boldsymbol{W}_{n_t}^{(0)}(k)\boldsymbol{R}(k),$$
(6)

where $\mathbf{R}(k) = [R_0(k), \dots, R_{N_r-1}(k)]^T$ is the received signal vector at the *k*th frequency. $\mathbf{W}_{n_t}^{(0)}(k)$ is the 1-by- N_r 2D

MMSE-FDE weight vector with ICI taken into account for the n_t th transmitted signal. Since ICI can be approximated as a Gaussian process, the noise plus ICI is treated as a new Gaussian noise and therefore, the 2D MMSE-FDE weight is given, from [2], as

$$\boldsymbol{W}_{n_{t}}^{(0)}(k) = \boldsymbol{H}_{n_{t}}^{H}(k) [\boldsymbol{H}(k) \boldsymbol{H}^{H}(k) + (C \cdot E_{c}/N_{0})^{-1} \boldsymbol{I}]^{-1}, \quad (7)$$

where H(k) is the N_r -by- N_t complex channel gain matrix whose element of the n_r th column and the n_t th row is $H_{n_r,n_t}(k)$, $H_{n_t}(k)$ is the n_t th column vector of H(k). E_c/N_0 (= ST_c/N_0) represents the average received chip energy-to-AWGN power spectrum density ratio per code, I is the N_r by- N_r identity matrix, and (.)^H is the Hermitian transpose operation. In Eq. (7), the term ($C \cdot E_c/N_0$)⁻¹ is due to the ICI; if the ICI is not taken into account, the achievable BER performance degrades drastically.

(b) Soft symbol replica generation

The time-domain received signal $\tilde{s}_{n_t}^{(i)}(t)$ is obtained by performing N_c -point IFFT on { $\tilde{R}_{n_t}^{(i)}(k)$; $k = 0 \sim N_c - 1$ }, after carrying out MMSE-FDE at the *i*th iteration, as:

$$\tilde{s}_{n_t}^{(i)}(t) = \frac{1}{N_c} \sum_{k=0}^{N_c-1} \tilde{R}_{n_t}^{(i)}(k) \exp\left(j2\pi t \frac{k}{N_c}\right).$$
(8)

 N_t parallel multicode signal blocks { $\tilde{s}_{n_t}^{(i)}(t)$; $n_t = 0 \sim N_t - 1$ and $t = 0 \sim N_c - 1$ } are converted by P/S conversion to the serial multicode signal $\tilde{s}^{(i)}(t')$, $t' = 0 \sim N_t N_c - 1$. Descrambling and multicode de-spreading operation yield the decision variable $\tilde{d}_c^{(i)}(n)$ of the *n*th symbol of the *c*th stream as

$$\tilde{d}_{c}^{(i)}(n) = \frac{1}{SF} \sum_{t'=nSF}^{(n+1)SF} \left\{ \tilde{s}^{(i)}(t') c_{c}^{*}(t' \mod SF) c_{scr}^{*}(t') \right\}$$
(9)

for $c = 0 \sim C - 1$, where * is the complex conjugate operation. Then, *C* parallel streams of decision variables are multiplexed into $\{\tilde{d}^{(i)}(n'); n' = 0 \sim CN_tN_c/SF - 1\}$, where $\tilde{d}^{(i)}(n')$ is the decision variable associated with the transmit symbol d(n'). Then, soft decision is carried out by using $\tilde{d}^{(i)}(n')$ to generate the soft symbol replica $\hat{d}^{(i+1)}(n')$ to be used in the (i + 1)th iteration [7]:

$$\hat{d}^{(i+1)}(n') = \frac{1}{\sqrt{2}} \left\{ \tanh\left(\beta \frac{\operatorname{Re}[\tilde{d}^{(i)}(n')]}{\sqrt{2S}}\right) + j \tanh\left(\beta \frac{\operatorname{Im}[\tilde{d}^{(i)}(n')]}{\sqrt{2S}}\right) \right\}$$
(10)

with

$$\tanh(x) = \frac{\exp(x) - \exp(-x)}{\exp(x) + \exp(-x)},\tag{11}$$

where β is a parameter that controls the extent to which the soft decision contributes to the replica generation. After performing multicode spreading and scrambling operations, the replicas of N_t parallel transmit chip streams $\hat{s}_{n_t}^{(i+1)}(t)$ are obtained.

(c) PIC operation

 N_c -point FFT is carried out again using $\hat{s}_{n_t}^{(i+1)}(t)$ to obtain the frequency-domain signal replica $\{\hat{S}_{n_t}^{(i+1)}(k); k = 0 \sim N_c - 1\}$:

$$\hat{S}_{n_t}^{(i+1)}(k) = \sum_{t=0}^{N_c - 1} \hat{s}_{n_t}^{(i+1)}(t) \exp\left(-j2\pi k \frac{t}{N_c}\right).$$
(12)

Then, the interference replica $\sqrt{2S} \sum_{n'_t=0 \neq n_t}^{N_t-1} H_{n_r,n'_t}(k) \hat{S}_{n'_t}^{(i+1)}(k)$ is generated and subtracted from the n_r th received signal $R_{n_r}(k)$ to obtain the n_r th received signal transmitted from the n_t th antenna $\hat{R}_{n_r,n_t}^{(i+1)}(k)$. The PIC operation to extract the received signal $\hat{R}_{n_r,n_t}^{(i+1)}(k)$ transmitted from the n_t th antenna is represented as

$$\hat{R}_{n_r,n_t}^{(i+1)}(k) = R_{n_r}(k) - \sqrt{2S} \sum_{\substack{n_t'=0\\\neq n_t}}^{N_t-1} H_{n_r,n_t'}(k) \hat{S}_{n_t'}^{(i+1)}(k).$$
(13)

(d) 1D MMSE-FDE at i > 0

The interference from other antennas still remains at the PIC output. This can be suppressed by the use of 2D MMSE-FDE at each iteration stage. However, the computational complexity of 2D MMSE-FDE is high. To reduce the complexity, in this paper, we apply 1D MMSE-FDE for i > 0. Joint 1D MMSE-FDE and antenna diversity combining is performed to obtain the signal component $\tilde{R}_{n_t}^{(i+1)}(k)$ at the *k*th frequency as

$$\tilde{R}_{n_t}^{(i+1)}(k) = \boldsymbol{W}_{n_t}^{(i+1)}(k) \hat{\boldsymbol{R}}_{n_t}^{(i+1)}(k), \qquad (14)$$

where $\hat{\mathbf{R}}_{n_t}^{(i+1)}(k) = [\hat{R}_{0,n_t}^{(i+1)}(k), \cdots, \hat{R}_{N_r-1,n_t}^{(i+1)}(k)]^T$ is the received signal vector transmitted from the n_t th antenna and $\mathbf{W}_{n_t}^{(i+1)}(k)$ is the 1-by- N_r 1D MMSE equalization weight vector with ICI taken into account in the (i+1)th iteration and is given by

$$\boldsymbol{W}_{n_{t}}^{(i+1)}(k) = \boldsymbol{H}_{n_{t}}^{H}(k) \Big/ \Big[\boldsymbol{H}_{n_{t}}^{H}(k) \boldsymbol{H}_{n_{t}}(k) + (C \cdot E_{c}/N_{0})^{-1} \Big].$$
(15)

The above processes (b)–(d) are repeated a sufficient number of times and hard decision is performed to recover the transmit data symbol sequence.

3. Computer Simulation Result

The average BER performance of multicode SS (N_t, N_r) MIMO multiplexing with the frequency-domain iterative PIC is evaluated by computer simulation. The simulation parameters are given in Table 1. We assume independent N_r -by- N_t frequency-selective Raleigh fading channels having a chip-spaced exponentially decaying L=16-path power delay profile with decay factor α and time delay $\tau_l = l$ for $l = 0 \sim L - 1$. Block fading is assumed, where the channel gains stay constant over one block of $(N_c + N_g)$ chips. Ideal channel estimation is assumed at the receiver.

First, we will evaluate the impact of β on the average BER performance and then the effect of iterative PIC on multicode SS transmission to optimize the value of β and

Data Modu	QPSK		
Number of Tx, Rx antennas		$N_t = N_r = 2,4$	
Multicode spreading	Spreading modulation	BPSK	
	Spreading factor	SF=1, 8, 64	
Scramble	M-sequence with a period of 4095		
code	chips		
Number of FFT points		$N_c = 256$	
GI		$N_{g}=32$	
Channel	<i>L</i> =16-path exponential power delay profile		
	Delay time $\tau_l = l$		
	Decay factor α=0,6dB		
Channel estimation		Ideal	





the number of iterations. Next, we will evaluate the impact of spreading factor *SF* and code-multiplexing order *C* when the equivalent spreading factor $SF_{eq}=1$. The performance comparison among frequency-domain iterative PICs with 1D MMSE-FDE and with 2D MMSE-FDE [6], ZF detection, MMSE detection, and V-BLAST is also presented.

3.1 Optimization of β and the Number of Iterations

Figure 4 plots the average BER of full code-multiplexed SS (2, 2)MIMO multiplexing (SF=C=64) as a function of β . The average received signal energy per information bit-to-AWGN power spectrum density ratio E_b/N_0 per antenna is set to 10 dB. As β increases, the BER reduces rapidly and approaches a constant value for $\beta \ge 5$ when i > 1, where i is the number of iterations. For (N_t, N_r)=(4, 4), we also found that $\beta=5$ minimizes the BER as well. Even for SC MIMO multiplexing (SF=1), the optimal $\beta=5$ was found to be used to minimize the BER. In the following simulations, we use $\beta=5$.



Fig. *b* Encer of nequency domain herality file.

Figure 5 plots the uncoded BER performances of full code-multiplexed SS (N_t, N_r) MIMO multiplexing with frequency-domain iterative PIC as a function of the average received E_b/N_0 per receive antenna. For comparison, the BER performance with perfect PIC is also plotted (perfect PIC assumes no feedback errors and provides the BER performance obtainable by $(1, N_r)$ SIMO). It can be seen from Fig. 5 that as the number of iterations increases, the BER performance improves and approaches that of perfect PIC. When $(N_t, N_r) = (2, 2)$, the degradation E_b/N_0 from perfect PIC for the average BER= 10^{-4} reduces by about 1.7, 0.7, 0.5 and 0.4 dB at i=1, 2, 3 and 4, respectively. When $(N_t, N_r) = (4, 4)$, this E_b/N_0 degradation is about 3.0, 1.2, 0.4 and 0.2 dB for i=1, 2, 3 and 4, respectively. Almost no additional improvement is obtained by increasing the number of iterations from 3 to 4. As a consequence, the use of 4 iterations is sufficient to achieve a BER performance close to perfect PIC by a fractional dB. In the following simulation,





Fig.7 Impact of *SF* for full code-multiplexing case ($SF_{eq}=1$).

we will present the performances with i=0 and 4 only.

3.2 Impact of Spreading Factor *SF* and Code-Multiplexing Order *C*

Figure 6 shows the BER performances of multicode SS (N_t, N_r) MIMO multiplexing with the code multiplexing order *C* as a parameter for the case of *SF*=64. Let us first concentrate on the performance without iterations. It can be seen from Fig. 6(a) that as *C* increases, the transmission rate increases *C* times (the full code-multiplexing case (*C*=*SF*) gives the transmission rate equal to the SC transmission), but the performance gets worse. In case of *i*=0, the required E_b/N_0 for the average BER=10⁻⁴ is larger by about 3.5 and 8.3 dB when *C*=8 and 64 (full code multiplexing), respectively, than when *C*=1 (SC transmission).

With *i*=4, frequency-domain iterative PIC provides a

BER performance close to the perfect PIC. When C=64, the required E_b/N_0 for the average BER= 10^{-4} is smaller by about 5.9 dB than that without iteration. As a result, the E_b/N_0 degradation from the C=1 case is only about 2.7 dB for C=64 with i=4 in contrast to 8.3 dB with i=0. Figure 6(b) plots the BER performance of multicode transmission when $N_t = N_r=4$. It can be seen that the trend is the same as in Fig. 6(a).

Figure 7 plots the average BER performance of full code-multiplexed SS (N_t , N_r) MIMO multiplexing with SF as a parameter when C=SF (the equivalent spreading factor $SF_{eq}=1$).

First we will discuss the impact of frequencyselectivity. It can be seen from Fig. 7 that the BER performance in case of α =6 dB is worse than that in case of α =0 dB. This reason can be explained as follows. The BER performance is better for stronger frequency-selectivity as



Fig. 8 BER performance comparison between 1D MMSE-FDE and 2D MMSE-FDE.

MMSE-FDE takes advantage of the frequency-selectivity of the channel and provides a higher frequency diversity gain. On the other hand, when $\alpha = 6 \text{ dB}$ (weak selectivity), sufficient frequency diversity gain can not be obtained. Furthermore, the channel gains at many frequencies fade simultaneously. Therefore, the accuracy of interference replica generation degrades. As a result, the BER performance in case of $\alpha = 6 \text{ dB}$ is worse than $\alpha = 0 \text{ dB}$.

Next, the impact of SF will be discussed. It can be seen from Fig. 7 that when $\alpha = 0 \, dB$ (strong selectivity), the BER performance is insensitive to SF and close to that with perfect PIC. However, when $\alpha = 6 \, dB$ (weak selectivity), the BER performance dependency on SF is stronger. The BER performance is better for larger SF. In case of $N_t = N_r = 2$, the E_b/N_0 degradation for the average BER=10⁻⁴ from the perfect PIC is about 3.5, 1.5 and 0.2 dB when SF=1, 8 and 64, respectively. On the other hand, in case of $N_t = N_r = 4$, the E_b/N_0 degradation from the perfect PIC is about 5.3 and $0.3 \, dB$ when SF=1 and 64, respectively. Although the accuracy of interference replica generation degrades in case of α =6 dB, as SF increases, the BER performance approaches that with perfect PIC. This is because the accuracy of interference replica improves by averaging the decision errors by means of code-multiplexing process. For the full codemultiplexing case (C=SF), as SF increases, the number of data symbols to be code-multiplexed increases and hence, the error averaging effect improves. As a result, the accuracy of the interference replica improves. On the other hand, in the SC transmission, the decision error for each symbol affects directly the interference replica since no averaging process is involved. Therefore, the full-code multiplexed SS can better suppress the interference from other antennas. As a consequence, multicode SS MIMO multiplexing provides better performance than SC transmission irrespective of frequency-selectivity. Especially, in case of $\alpha = 6 \text{ dB}$, the superiority is significant.



Fig. 9 Performance comparisons between proposed and other conventional schemes.

3.3 BER Performance Comparison

Figure 8 shows the performance comparison between 1D MMSE-FDE and 2D MMSE-FDE [6] for the case of four iterations (*i*=4) when $N_t = N_r$ =4. When α =0 dB (strong selectivity), 1D MMSE-FDE can provide almost the same BER performance as 2D MMSE-FDE irrespective of *SF*. This is because the accuracy of the interference replica improves due to large frequency diversity gain. The achievable BER performance is very close to the case of perfect PIC.

When α =6 dB (weak selectivity), the BER performance with 1D MMSE-FDE gets worse than that with 2D MMSE-FDE if *SF*=1, because a sufficient frequency diversity gain cannot be achieved and hence, 1D MMSE-FDE cannot well suppress the interference from other antennas. The required E_b/N_0 for the average BER=10⁻⁴

	Weight computation	FFT (IFFT)	Interference cancellation	FDE
Proposed scheme	$\begin{cases} N_c \times N_r^2 \times (2N_t + N_r) , i = 0 \\ N_c \times N_r \times \{N_r \times (2N_t + N_r) + 2N_t\}, i \ge 1 \end{cases}$	$(1+2i) \times N_t \times N_c \log N_c$	$i \times N_t \times N_r \times N_c$	$(1+i) \times N_t \times N_r \times N_c$
ZF detection	$N_c \times N_t^2 \times (2N_r + N_t)$	$N \times N \log N$		$N \times N \times N$
MMSE detection	$N_c \times N_r^2 \times (2N_t + N_r)$	My XM c log M c		1. C I b
V-BLAST	$N_c \times \sum_{n_t=0}^{N_t-1} [(N_t - n_t)^2 \{2N_r + (N_t - n_t)\}]$	$(2N_t - 1) \times N_c \log N_c$	$(N_t - 1) \times N_c \times N_r$	$N_c \times N_t \times N_r$

Table 2Number, i, of multiply operations.

with 1D MMSE-FDE gets larger by about 2 dB than that of 2D MMSE-FDE. On the other hand, if SF=64, the accuracy of the replica can be improved by error averaging effect (obtained by the code-multiplexing process), and hence, 1D MMSE-FDE gives the BER performance close to 2D MMSE-FDE and also perfect PIC.

Next, the BER performance of the proposed scheme using four iterations is compared with those of ZF detection, MMSE detection, and V-BLAST based on MMSE detection [5], [9]. Figure 9 shows the BER performance of (N_t, N_r) MIMO multiplexing when $\alpha = 0$ dB. It can be seen that the proposed scheme can significantly improve the BER performance compared to the conventional schemes. When $(N_t, N_r) = (2, 2)$, the proposed scheme can reduce the required average E_b/N_0 for an average BER=10⁻⁴ by about 22 dB, 6 dB and 4 dB compared to ZF detection, MMSE detection and V-BLAST, respectively. When $(N_t, N_r) = (4, 4)$, the required average E_b/N_0 of the proposed scheme is smaller by about 27.5 dB, 9.1 dB and 5.9 dB than ZF detection, MMSE detection and V-BLAST, respectively. The superiority of the proposed scheme is because a higher order of antenna diversity, equal to the N_r th order, can be achieved. Only an $(N_r - N_t + 1)$ -order diversity gain can be obtained by ZF detection and MMSE detection [2]. Furthermore, the performance with ZF detection degrades due to the noise enhancement. On the other hand, a diversity order of between $(N_r - N_t + 1)$ and N_r can only be obtained by V-BLAST. The reason for this is as follows. In V-BLAST, the signals are separated according to the descending order of signals' reliability. After a signal is separated, its component is eliminated from the received signals by using its replica and then the corresponding channel gain column vector is deleted from the N_r -by- N_t channel gain matrix. For the *n*th signal separation, the N_r -by- $(N_t - n)$ channel gain matrix is used $(n=0 \sim N_t - 1)$. Therefore, a diversity order of $(N_r - N_t + n + 1)$, i.e., from $(N_r - N_t + 1)$ to N_r , can only be obtained by V-BLAST.

4. Complexity Comparison

The computational complexity of the proposed scheme, in terms of the number of multiply operations, is compared with that of 2D MMSE-FDE [6]. First, we discuss the number of multiply operations required for the weight computation of MMSE-FDE per iteration. In 2D MMSE-FDE, a total number of multiply operations is $N_t \times N_c \times N_r \times (N_t^2 + N_r^2 + N_r \times N_t + N_r)$ per each iteration (i > 0). On the other hand, in 1D MMSE-FDE, computing the norm of { $H_{n_t}(k)$; $n_t = 0 \sim N_t - 1$ } and dividing $H_{n_t}^H(k)$ by $H_{n_t}^H(k)H_{n_t}(k) + (C \cdot E_c/N_0)^{-1}$ are required for each frequency and each transmit antenna, but only at the first iteration (i=1). Therefore, a total number of multiply operations required for the weight computation is $2N_t \times N_c \times N_r$. When $N_t = N_r = 4$, since the use of four iterations (i=4) is sufficient, the computational complexity of 1D MMSE-FDE is only about 6% of 2D MMSE-FDE.

The computational complexity of the proposed scheme, in terms of the number of multiply operations, is compared with ZF detection, MMSE detection, V-BLAST and MLD. The number of multiply operations is given in Table 2. The complexity of the proposed scheme linearly increases in proportion to the number of iterations. Therefore, the complexity gets larger than those of ZF detection, MMSE detection and V-BLAST. When $(N_t, N_r) = (4, 4)$, the complexity of the proposed scheme with i=4 is about 2.7 times higher than those of ZF detection and MMSE detection. On the other hand, the complexity of the proposed scheme is about 1.5 times higher than that of V-BLAST. We can conclude that at the cost of the increased complexity, the proposed scheme can provide better BER performance than ZF detection, MMSE detection and V-BLAST, but the complexity is much lower than the MLD even when using four iterations. This is because MLD requires $N_t \times N_c \times N_r \times 2^{N_c M}$ multiply operations to compute the squared Euclidean distance between the received signal and the candidate signal, where *M* is the number of bits per symbol.

5. Conclusion

In this paper, a less computationally complex iterative frequency-domain PIC was presented for multicode SS MIMO multiplexing in a frequency-selective fading channel. Joint equalization and signal separation is performed simultaneously by applying 2D MMSE-FDE with ICI taken into account. 2D MMSE weight with ICI taken into account was derived theoretically. Furthermore, to well suppress the interference from other antennas while reducing the computational complexity, the full code-multiplexed SS, which can maximize the transmission rate, is used and joint PIC and 1D MMSE-FDE is repeated a sufficient number of times. We evaluated, by computer simu-

lation, the achievable BER performance with the proposed scheme in a frequency-selective Rayleigh fading channel. It was shown that the BER performance with four iterations can be close to the BER performance of SIMO transmission by using the MMSE-FDE with ICI taken into account. Even using 1D MMSE-FDE can provide almost the same BER performance as 2D MMSE-FDE while reducing the computational complexity. Furthermore, the full code-multiplexing $(SF_{eq}=1)$ provides the BER performance much better than the non-spread SC transmission in a weak frequency-selectivity channel, but only slight performance improvement is obtained in a strong frequency-selective channel. The BER performance comparison and complexity comparison were also presented to show that the proposed scheme can provide better BER performance than ZF detection, MMSE detection and V-BLAST at the cost of increased complexity.

In this paper, ideal channel estimation was assumed. Studying the iterative PIC performance with practical channel estimation is important. This is left as an interesting future study.

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