ICI Canceling Space-Frequency Block Code for MISO-OFDM in Fast Fading Channels

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Abstract— **An intercarrier interference (ICI) canceling technique for multiple-input single-output (MISO) orthogonal frequency division multiplexing (OFDM) systems in fast fading channels is proposed. The proposed scheme consists of a linear space-frequency block code (SFBC) at the transmitter and a receive frequency block code (RFBC) at the receiver. The code design is based on the upper bound of pairwise error probability (PEP) which is derived under assumptions and approximations stated. Specifically, the proposed code is designed to minimize the upper bound of the PEP. The simulation results demonstrate that the proposed technique outperforms the conventional methods in fast fading environments.**

I. Introduction

Orthogonal frequency division multiplexing (OFDM) is a promising technique for broadband communication systems since it converts a frequency-selective broadband channel into a series of parallel narrowband channels. Therefore, the OFDM can reduce the equalization and decoding complexity [1], [2]. In fact, many modern communication standards for high data rates adopt the OFDM as a transmission technique [3]–[5]. However, it has been recognized that the performance of the OFDM can be severely degraded if the channel varies significantly within an OFDM symbol period [6]. This degradation is due to the intercarrier interference (ICI). To compensate for the ICI, various techniques have been introduced for single-input single-output (SISO) systems, including the frequency- and time-domain equalizations (see [7] and references therein), the ICI self-cancelation [8]–[11], and the frequency-domain partial response coding (PRC) [12], [13]. Among those approaches, the equalizations are the most common, but implementation requires heavy computations. The ICI self-cancelation schemes are based on the use of a frequency-domain coding or a time-domain windowing, which has a code rate of $1/k$. This method can cancel the ICI effectively with a small computational complexity at the receiver; however, the spectral efficiency decreases by a factor of $1/k$. The PRC is based on the partial response signaling for single-carrier systems [14]. It does not sacrifice spectral efficiency, but often needs a maximum-likelihood sequence estimator (MLSE) at the receiver to minimize the bit-error-rate (BER) performance loss and this increases the receiver complexity significantly. Recently, ICI canceling techniques for multiple-input multiple-output (MIMO) systems have been investigated: in [17], the space-frequency (SF) coding for MIMO-OFDM is combined with the frequency-domain ICI self-cancelation technique in [8], and in [18], the frequency-domain PRC [12] is employed for an OFDM-based spatial multiplexing system. These methods can provide either diversity and coding gains or a throughput gain in addition to the ICI cancelation ability.

This paper considers the ICI canceling method by using a space-frequency block coding (SFBC) at the transmitter and a receive-frequency block coding (RFBC) at the receiver for multiple-input single-output (MISO) systems. The code design is based on the approximated upper bound of the pairwise error probability (PEP). Specifically, the proposed SFBC and RFBC are obtained by minimizing the upper bound of the PEP among all possible pairs of the transmitted signal and erroneous signal. Through this approach, the SFBC and RFBC with an arbitrary code rate t/k ($t \leq k$) are designed. Although this paper only considers the MISO-OFDM, the proposed method can be easily extended to MIMO-OFDM by considering the use of the identical RFBC at all receive antennas. The simulation results demonstrate that the proposed method outperforms the conventional method in fast fading channels.

The notations used throughout this paper are summarized as follows: $E[\cdot]$ denotes expectation; $\mathrm{diag}(\mathbf{a})$ is an n-by-n diagonal matrix whose (p,p)-th element is the p-th element of an n-dimensional vector $\mathbf{a}$; $[\mathbf{A}]_{p,q}$ denotes the (p,q)-th element of matrix $\mathbf{A}$; $\mathbf{I}_t$ indicates a t-by-t identity matrix; $\mathbf{0}_{t\times k}$ denotes a t-by-k null matrix; and $\mathbf{1}_{t\times k}$ represents a t-by-k matrix with all one entries.

II. System Model

Figure 1 illustrates the MISO-OFDM system with frequency-domain coding and M transmit antennas. Since

This work was supported in part by the Brain Korea 21 Project, The school of information technology, KAIST in 2007.

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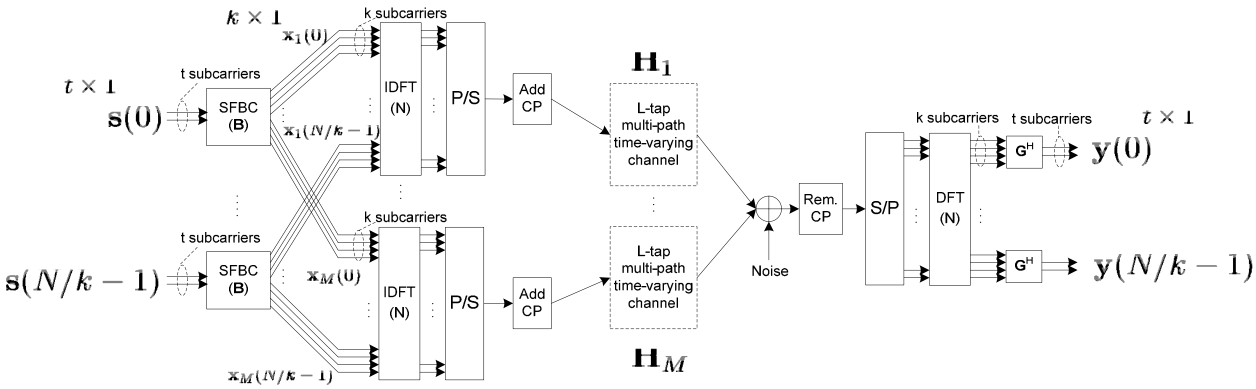

Fig. 1. MISO-OFDM system employing rate t/k ICI canceling SFBC and RFBC.

coding at the transmitter is performed across space (antenna domain) and frequency (subcarrier domain) in a block-wise manner, the scheme is referred to as SFBC. There are N/k t-dimensional input vectors $\{\mathbf{s}(u)\}_{u=0}^{N/k-1}$ at the transmitter, where N is the number of OFDM subcarriers, and k divides N. Each $\{\mathbf{s}(u)\}$ is coded by the SFBC block, i.e. multiplied by the Mk-by-t matrix $\mathbf{B}$ (t/k is the code rate). The resulting Mk-dimensional vector $\mathbf{x}(u)$ is partitioned into M blocks $\{\mathbf{x}_m(u)|\mathbf{x}_m(u) = \mathbf{B}_m\mathbf{s}(u)\}_{m=1}^{M}$ where $\{\mathbf{B}_m\}_{m=1}^{M}$ are k-by-t submatrices defined as $\mathbf{B} = [\mathbf{B}_1^T \cdots \mathbf{B}_M^T]^T$, and $\{\mathbf{x}_m(u)\}_{m=1}^{M}$ are distributed to M IDFT blocks of M antennas, respectively. Now by stacking the signals of m-th antenna, $\{\mathbf{x}_m(u)\}_{u=0}^{N/k-1}$ along the frequency subcarriers, the augmented N-dimensional vectors are obtained as $\{\mathbf{x}_m\}_{m=1}^{M}$ where $\mathbf{x}_m = [\mathbf{x}_m^T(0) \cdots \mathbf{x}_m^T(N/k-1)]^T$. $\{\mathbf{x}_m\}_{m=1}^{M}$ are multiplied by the N-point IDFT matrix $\mathbf{Q}^H$, where $[\mathbf{Q}]_{p,q} = (1/\sqrt{N}) \cdot e^{-j\frac{2\pi}{N}pq}$, and passed through M L-tap multi-path fading channels, respectively. At the receiver, after the DFT operation, the received signal vector is partitioned into N/k k-dimensional vectors, and each of these is multiplied by the t-by-k receiver coding matrix $\mathbf{G}^H$. The receiver coding is called RFBC. Before expressing the received t-dimensional signal vector $\mathbf{y}(u)$, the following notations are defined: $\mathbf{H}_m = \mathbf{Q}\mathbb{H}_m\mathbf{Q}^H$, which is the frequency-domain channel matrix from the m-th transmit antenna to the receiver. Here, $\mathbb{H}_m$ is an N-by-N time-domain channel matrix from the m-th transmit antenna to the receiver, whose (p,q)-th element is given by

$$[\mathbb{H}_m]_{p,q} = \begin{cases} h_m(p;(p-q)_N) & \text{if } 0 \le (p-q)_N \le L-1 \\ 0 & \text{otherwise,} \end{cases} \tag{1}$$

where $h_m(n;l)$ represents the impulse response of the channel at time n due to an impulse that is applied l time units earlier from the m-th antenna. To represent the received signal block $\mathbf{y}(u)$ briefly, the N-by-N matrices $\{\mathbf{H}_m\}$ are partitioned into k-by-k subblocks as follows:

$$\mathbf{H}_m = \begin{bmatrix} \mathbf{H}_m(0,0) & \cdots & \mathbf{H}_m(0,N/k-1) \\ \vdots & \ddots & \vdots \\ \mathbf{H}_m(N/k-1,0) & \cdots & \mathbf{H}_m(N/k-1,N/k-1) \end{bmatrix} \tag{2}$$

where $\mathbf{H}_m(u,v)$ is a k-by-k block. If the time-domain channel is fixed during an OFDM symbol period, $\mathbf{H}_m$ becomes a diagonal matrix. Otherwise, non-zero values appear in off-diagonals which cause the ICI. Using the partitioned channel matrices, $\mathbf{y}(u)$ can be written as

$$\mathbf{y}(u) = \mathbf{G}^H \sum_{m=1}^{M} \mathbf{H}_m(u,u)\mathbf{x}_m(u) + \mathbf{G}^H \sum_{m=1}^{M} \sum_{v=0,\ v\neq u}^{N/k-1} \mathbf{H}_m(u,v)\mathbf{x}_m(v) + \mathbf{G}^H\mathbf{w}(u), \tag{3}$$

where $\{\mathbf{w}(u)\}_{u=0}^{N/k-1}$ are the k-dimensional additive white Gaussian noise (AWGN) vectors. Since the off-diagonal terms of $\mathbf{H}_m(u,u)$ are also ICI terms, the desired diagonal channel elements are separated from $\mathbf{H}_m(u,u)$ as

$$\mathbf{H}_m(u,u) = \bar{\mathbf{H}}_m(u) + \tilde{\mathbf{H}}_m(u), \tag{4}$$

where $\tilde{\mathbf{H}}_m(u)$ is the same as $\mathbf{H}_m(u,u)$ except that the diagonal elements are all zeros. Hence, $\bar{\mathbf{H}}_m(u)$ is a diagonal matrix whose diagonal elements are the same as $\mathbf{H}_m(u,u)$. Then $\mathbf{y}(u)$ can be arranged as

$$\mathbf{y}(u) = \mathbf{G}^H \sum_{m=1}^{M} \bar{\mathbf{H}}_m(u)\mathbf{x}_m(u) + \mathbf{G}^H \sum_{m=1}^{M} \tilde{\mathbf{H}}_m(u)\mathbf{x}_m(u) + \mathbf{G}^H \sum_{m=1}^{M} \sum_{v=0,\ v\neq u}^{N/k-1} \mathbf{H}_m(u,v)\mathbf{x}_m(v) + \mathbf{G}^H\mathbf{w}(u). \tag{5}$$

Here, the first term is the desired signal term and the remaining terms are interference and noise terms. $\mathbf{y}(u)$ in (5) is further rewritten in a vector form as:

$$\mathbf{y}(u) = \mathbf{G}^H\mathbf{X}(u)\mathbf{h}_A(u) + \mathbf{v}_A(u), \tag{6}$$

where

$$\mathbf{X}(u) = [\mathrm{diag}(\mathbf{x}_1(u)) \ \cdots \ \mathrm{diag}(\mathbf{x}_M(u))] = [\mathrm{diag}(\mathbf{B}_1\mathbf{s}(u)) \ \cdots \ \mathrm{diag}(\mathbf{B}_M\mathbf{s}(u))], \quad (7)$$

$$\mathbf{h}_A(u) = \big[[\bar{\mathbf{H}}_1(u)]_{0,0} \ \cdots \ [\bar{\mathbf{H}}_1(u)]_{k-1,k-1} \ \cdots \ [\bar{\mathbf{H}}_M(u)]_{0,0} \ \cdots \ [\bar{\mathbf{H}}_M(u)]_{k-1,k-1}\big]^T, \text{ and} \quad (8)$$

$$\mathbf{v}_A(u) = \mathbf{G}^H \sum_{m=1}^{M} \tilde{\mathbf{H}}_m(u)\mathbf{x}_m(u) + \mathbf{G}^H \sum_{m=1}^{M} \sum_{v=0,\ v\neq u}^{N/k-1} \mathbf{H}_m(u,v)\mathbf{x}_m(v) + \mathbf{G}^H\mathbf{w}(u). \quad (9)$$

$\mathbf{X}(u)$ is a k-by-kM matrix partitioned into k-by-k diagonal matrices. The diagonal elements of the m-th partition represent the k dimensional signal vector corresponding to the m-th antenna and u-th SFBC block in Fig. 1. $\mathbf{h}_A(u)$ is a kM-by-1 vector partitioned into k-by-1 vectors where the m-th k-by-1 vector is the desired channel response for $\mathbf{x}_m(u)$, i.e. the diagonal elements of $\bar{\mathbf{H}}_m(u)$. $\mathbf{v}_A(u)$ is a t-by-1 vector representing the ICI and noise after RFBC at the receiver. It is assumed that t symbols in $\mathbf{s}(u)$ are jointly detected from $\mathbf{y}(u)$ in (6). Specifically, $\mathbf{s}(u)$ is detected by the maximum-likelihood (ML) scheme. Although the ML detector finds $\hat{\mathbf{s}}(u)$ maximizing the likelihood function over all possible candidates for $\mathbf{s}(u)$,[1] the receiver complexity can be controlled by properly selecting t which is the size of $\mathbf{s}(u)$. The SFBC and RFBC are designed so that the upper bound of the PEP is minimized. Throughout this paper, $\{\mathbf{h}_A(u)\}$ are assumed to be perfectly known at the receiver while they are unknown at the transmitter.

III. Approximated Pairwise Error Probability

The PEP is a popular measure for space-time or space-frequency code designs [15], [16]. By designing a code that minimizes the PEP, it is possible to minimize the ICI and maximize the diversity and the coding gain at the same time. Before formulating the PEP of the system in (6), the following assumptions are established:

- A.1) The channels are spatially independent.
- A.2) Time-domain channel tap coefficients are mutually uncorrelated and wide-sense stationary. The magnitudes and phases are Rayleigh-distributed and uniformly distributed, respectively. Specifically,
$$E[h_m(n;l)h_m^*(n';l')] = \sigma_{l,m}^2 r_m(n-n') \cdot \delta(l-l'), \quad (10)$$
where $r_m(n) = E[h_m(n'+n;l)h_m^*(n';l)]/\sigma_{l,m}^2$ and $\sigma_{l,m}^2$ denotes the mean power of the l-th channel tap from the m-th transmit antenna to the receiver. The channel power is normalized to satisfy $\sum_l \sigma_{l,m}^2 = 1$ for all m. In addition, the distributions of M channels from different transmit antennas are the same. Based on this assumption, the variable m can be omitted, i.e. $r_m(n) = r(n)$ and $\sum_l \sigma_{l,m}^2 = \sum_l \sigma_l^2 = 1$ for all m.
- A.3) The noise $\mathbf{w}(u)$ is circularly symmetric Gaussian random vectors with the covariance $\mathbf{R}_{w(u)} = \sigma^2\mathbf{I}_k$. Hence, the signal-to-noise ratio (SNR) is $1/\sigma^2$.
- A.4) ML detection is employed at the receiver and channel coding is not considered.
- A.5) The distribution of $\mathbf{v}_A(u)$ is approximated to Gaussian. In fact, $\mathbf{v}_A(u)$ is not Gaussian because it is a linear combination of the products of Gaussian random processes.
- A.6) The covariance of $\mathbf{v}_A(u)$ is independent of the input signal, $\mathbf{s}(u)$. Actually, $\mathbf{v}_A(u)$ is a function of $\mathbf{s}(u)$ and so is the covariance of $\mathbf{v}_A(u)$. To make $\mathbf{v}_A(u)$ independent of $\mathbf{s}(u)$, the original covariance is averaged over the distribution of $\mathbf{s}(u)$, i.e. $\mathbf{R}_{v(u)} = E[E[\mathbf{v}(u)\mathbf{v}(u)^H|\mathbf{s}(u)]]$, where $\mathbf{R}_{v(u)}$ is the averaged covariance matrix of $\mathbf{v}_A(u)$. Here, $\mathbf{s}(u)$ is assumed to be mutually uncorrelated ($E[\mathbf{s}(u)\mathbf{s}^H(u')] = 0$ for $u \neq u'$) and has a covariance of $\mathbf{R}_{s(u)} = \mathbf{I}_t$ for all u.

Based on assumptions A.1 to A.6, the approximated PEP (APEP) is derived. Now consider the input vectors $\mathbf{s}(u)$ and $\tilde{\mathbf{s}}(u)$ where $\mathbf{s}(u) \neq \tilde{\mathbf{s}}(u)$. By the SFBC at the transmitter, $\mathbf{s}(u) \to \{\mathbf{x}_m(u)\}_{m=1}^M \to \mathbf{X}(u)$ and $\tilde{\mathbf{s}}(u) \to \{\tilde{\mathbf{x}}_m(u)\}_{m=1}^M \to \tilde{\mathbf{X}}(u)$. The ML detector chooses the transmit vector that makes the observation $\mathbf{y}(u)$ most likely. Given $\mathbf{h}_A(u)$ and $\mathbf{R}_{v_A(u)}$, and under the assumption of A.5, the APEP between $\mathbf{s}(u)$ and $\tilde{\mathbf{s}}(u)$ is derived as:

$$\mathbb{P}(\mathbf{X}(u) \to \tilde{\mathbf{X}}(u)|\mathbf{h}_A(u)) = Q\left(\frac{\|\mathbf{R}_{v(u)}^{-1/2}\mathbf{G}^H(\mathbf{X}(u)-\tilde{\mathbf{X}}(u))\mathbf{h}_A(u)\|}{\sqrt{2}\sigma}\right), \quad (11)$$

where $\mathbf{R}_{v_A(u)} = \mathbf{R}_{v_A(u)}^{1/2}\mathbf{R}_{v_A(u)}^{1/2H}$. Since the Q function is difficult to deal with, the Chernoff bound, $Q(x) \leq \exp(-x^2/2)$, is introduced to yield

$$\mathbb{P}(\mathbf{X}(u) \to \tilde{\mathbf{X}}(u)|\mathbf{h}_A(u)) \leq \exp\left(-\frac{\|\mathbf{R}_{v(u)}^{-1/2}\mathbf{G}^H(\mathbf{X}(u)-\tilde{\mathbf{X}}(u))\mathbf{h}_A(u)\|^2}{4\sigma^2}\right). \quad (12)$$

Since the SFBC and RFBC should not depend on an instantaneous channel response, the expectation of the upper bound in (12) with respect to the distribution of $\mathbf{h}_A(u)$ is used. After some calculations, the upper bound of the averaged APEP, P_E, is written as:

$$\mathbb{P}(\mathbf{s}(u) \to \tilde{\mathbf{s}}(u)) \leq \frac{1}{\prod_{i=1}^{rank(\Delta)}\left(1+\lambda_i(\Delta)\cdot\frac{1}{4\sigma^2}\right)} \equiv P_E, \quad (13)$$

where

$$\Delta = \sigma^2\mathbf{R}_{v_A(u)}^{-1/2}\mathbf{G}^H(\mathbf{X}(u)-\tilde{\mathbf{X}}(u))\mathbf{R}_{h_A(u)}(\mathbf{X}(u)-\tilde{\mathbf{X}}(u))^H\mathbf{G}\mathbf{R}_{v_A(u)}^{-1/2H}, \quad (14)$$

$\mathbf{R}_{h_A(u)}$ is the covariance matrix of $\mathbf{h}_A(u)$, i.e. $E[\mathbf{h}_A\mathbf{h}_A^H]$, and $rank(\Delta)$ and $\{\lambda_i(\Delta)\}_{i=1}^{r(\Delta)}$ denote the rank and nonzero eigenvalues of Δ, respectively. The SFBC matrix $\mathbf{B}$ and the RFBC matrix $\mathbf{G}$ are designed to minimize P_E. To proceed to designing $\mathbf{B}$ and $\mathbf{G}$, $\mathbf{R}_{h_A(u)}$ and $\mathbf{R}_{v_A(u)}$ need to be represented in terms of $\mathbf{B}$ and $\mathbf{G}$. Since $E[[\bar{\mathbf{H}}_m(u)]_{k,k}[\bar{\mathbf{H}}_{m'}(u)]_{k',k'}^*] = 0$ if $m \neq m'$ for all k and k'

[1] Note that $\mathbf{X}(u)$ is linked to $\mathbf{s}(u)$ by the relation in (7).

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by A.1, $\mathbf{R}_{h_A(u)}$ becomes a block-diagonal matrix as $\mathbf{R}_{h_A(u)} = \mathrm{diag}(\mathbf{R}_h^{(1)}, \cdots, \mathbf{R}_h^{(M)})$ where $\mathbf{R}_h^{(m)}$ is given by:

$$\mathbf{R}_h^{(m)} = \frac{1}{N^2}\left(\mathbf{1}_{N,1}^H \mathbf{R}_T \mathbf{1}_{N,1}\right)\mathbf{R}_F, \tag{15}$$

where $\mathbf{R}_F$ is a k-by-k matrix defined as $[\mathbf{R}_F]_{p,q} = \sum_{l=0}^{L-1} \sigma_l^2 e^{-j\frac{2\pi}{N}l(p-q)}$ and $\mathbf{R}_T$ is an N-by-N matrix given by $[\mathbf{R}_T]_{p,q} = r(p-q)$. Expressing $\mathbf{R}_{v_A(u)}$ in $\mathbf{B}$ and $\mathbf{G}$ is rather complex. Under assumption A.1, $\mathbf{R}_{v_A(u)}$ can be written as:

$$\mathbf{R}_{v_A(u)} = \mathbf{G}^H \sum_{m=1}^{M} \mathbf{K}_p^{(m)} \mathbf{G} + \sigma^2 \mathbf{G}^H \mathbf{G}, \tag{16}$$

where $\mathbf{K}_p^{(m)}$ is the covariance of $\mathbf{p}^{(m)}(u)$ and:

$$\mathbf{p}^{(m)}(u) = \tilde{\mathbf{H}}_m(u)\mathbf{x}_m(u) + \sum_{v=0,\ v\neq u}^{N/k-1} \mathbf{H}_m(u,v)\mathbf{x}_m(v). \tag{17}$$

After some calculations, $\mathbf{R}_{v_A(u)}$ can be expressed as

$$\begin{aligned}\mathbf{R}_{v_A(u)} = \mathbf{G}^H \sum_{m=1}^{M} \Bigg[& \tilde{\mathbf{Q}} \left\{ \frac{N}{k} \mathbf{R}_T^{\prime(m)} \otimes (\mathbf{R}_F^{(m)} \circ \mathbf{R}_{x_m}) \right\} \tilde{\mathbf{Q}}^H \\ & - \frac{1}{N} \left\{ \left(\mathbf{1}_{1\times N} \mathbf{R}_T^{(m)}\right) \otimes \left(\mathbf{R}_F^{(m)} \circ \mathbf{R}_{x_m}\right) \right\} \tilde{\mathbf{Q}}^H \\ & - \frac{1}{N} \tilde{\mathbf{Q}} \left\{ \left(\mathbf{R}_T^{(m)} \mathbf{1}_{N\times 1}\right) \otimes \left(\mathbf{R}_F^{(m)} \circ \mathbf{R}_{x_m}\right) \right\} \\ & + \frac{1}{N^2} \mathbf{1}_{1\times N} \mathbf{R}_T^{(m)} \mathbf{1}_{N\times 1} \cdot \left(\mathbf{R}_F^{(m)} \circ \mathbf{R}_{x_m}\right) \Bigg] \mathbf{G} + \sigma^2 \mathbf{G}^H \mathbf{G}.\end{aligned} \tag{18}$$

It is observed that $\mathbf{R}_{h_A(u)}$ and $\mathbf{R}_{v_A(u)}$ are independent of u. Therefore, the notation $\mathbf{R}_{h_A}$ and $\mathbf{R}_{v_A}$ are used instead of $\mathbf{R}_{h_A(u)}$ and $\mathbf{R}_{v_A(u)}$, respectively. In summary, if the channel autocorrelation function $r(n)$, the channel power profile $\{\sigma_l^2\}$, and the received signal's SNR are given, $\mathbf{R}_{h_A}$ and $\mathbf{R}_{v_A}$ can be expressed as a function of $\mathbf{B}$ and $\mathbf{G}$. Based on this, Δ in (14) and finally, P_E in (14) can be written as a function of $\mathbf{B}$ and $\mathbf{G}$.

Before concluding this section, it is worthwhile to make the following observation.

Observation 1: If $k \ll N$ and $L \ll N$, $\mathbf{R}_F$ for a frequency-selective channel ($L > 1$) can be approximated by that for a flat fading channel ($L = 1$), i.e. $\mathbf{R}_F \approx \mathbf{1}_{k\times k}$.
This observation is verified as follows:

$$[\mathbf{R}_F]_{p,q} = \sum_{l=0}^{L-1} \sigma_{m,l}^2 e^{-j\frac{2\pi}{N}l(p-q)} \approx \sum_{l=0}^{L-1} \sigma_{m,l}^2 = 1.$$

Therefore, if we choose k and N so that $k \ll N$ and $L \ll N$, the code matrix designing problem for the frequency-selective fading channels is modified into that for flat fading channels.

IV. Designing Code Matrices

In this section, the code matrices $\mathbf{B}$ and $\mathbf{G}$ are designed to minimize the maximum P_E over all possible pairs of different input vectors $\mathbf{s}(u)$ and $\tilde{\mathbf{s}}(u)$. The problem to find the optimum $\mathbf{B}$ and $\mathbf{G}$ can be written as:

$$\arg\min_{\mathbf{B},\mathbf{G}} \max_{\mathbf{s}(u),\tilde{\mathbf{s}}(u)} P_E, \text{ subject to } tr[\mathbf{B}\mathbf{B}^H] = k. \tag{19}$$

Based on Observation 1, the flat fading channel ($L = 1$) is considered instead of the frequency-selective fading channel ($L > 1$). Then, from (15) and (18),

$$\mathbf{R}_{h_A} = \frac{1}{N^2} \cdot \mathbf{1}_{1\times N} \mathbf{R}_T \mathbf{1}_{N\times 1} \cdot \mathrm{diag}(\mathbf{1}_{k\times k}, \cdots, \mathbf{1}_{k\times k}), \tag{20}$$

and

$$\begin{aligned}\mathbf{R}_{v_A} = \; & \mathbf{G}^H \sum_{m=1}^{M} \Bigg[\tilde{\mathbf{Q}} \left\{ \frac{N}{k} \mathbf{R}_T' \otimes \mathbf{B}_m \mathbf{B}_m^H \right\} \tilde{\mathbf{Q}}^H - \frac{1}{N} \left\{ \left(\mathbf{1}_{1\times N} \mathbf{R}_T\right) \otimes \mathbf{B}_m \mathbf{B}_m^H \right\} \tilde{\mathbf{Q}}^H \\ & - \frac{1}{N} \tilde{\mathbf{Q}} \left\{ \left(\mathbf{R}_T \mathbf{1}_{N\times 1}\right) \otimes \mathbf{B}_m \mathbf{B}_m^H \right\} + \frac{1}{N^2} \mathbf{1}_{1\times N} \mathbf{R}_T \mathbf{1}_{N\times 1} \cdot \mathbf{B}_m \mathbf{B}_m^H \Bigg] \mathbf{G} \\ & + \sigma^2 \mathbf{G}^H \mathbf{G},\end{aligned} \tag{21}$$

where $\mathbf{R}_{x_m}$ is substituted by $\mathbf{B}_m \mathbf{B}_m^H$. To proceed further, a channel correlation function $r(n)$ is required which is assumed to be $J_0(2\pi f_d T_s n/N)$ [6] where $J_0(x)$ is a zeroth-order Bessel function of the first kind, f_d is the maximum Doppler frequency, and T_s is the OFDM symbol period. However, a problem remains since the maximum Doppler frequency depends on the carrier frequency and vehicle speed, which is unknown at the transmitter. Therefore, we approximate $r(n)$ by averaging $J_0(2\pi f_d T_s n/N)$ over the range $0 \leq f_d T_s \leq 0.2$. Based on this approximation and P_E in (13), the optimum SFBC and RFBC that satisfy (19) are found using a nonlinear optimization tool. The parameters for the optimization are as follows: $M = \{1, 2\}$, $N = 240$, SNR $= 25$ dB, and $(k, t) \in \{(2, 1), (2, 2), (3, 2), (3, 3), (4, 2)\}$. To guarantee an identical transmission rate, the QPSK modulation is used in the optimization for $(k, t) \in \{(2, 2), (3, 3)\}$, the 8PSK modulation is employed for $(3, 2)$, and 16QAM is applied to $(k, t) \in \{(2, 1), (4, 2)\}$. The computer program tool used to solve the nonlinear optimization problem in (19) is the function 'fmincon' in the MATLAB optimization toolbox. Since the solution may be a local minimum depending on the initial condition, the function is executed 30 times with different initial conditions, and then the solution $\{\mathbf{B}, \mathbf{G}\}$ that provides the minimum P_E from the 30 trials is chosen as the final SFBC and RFBC matrix. The best code matrices obtained through this procedure are shown in Tables I and II. It is observed that when the number of transmit antenna is one, the identical $\mathbf{B}$ and $\mathbf{G}$ can be the best solution.

V. Simulation Results

The BER performance of the proposed ICI canceling SFBC and receive codes in Tables I and II were empirically examined through computer simulations. The simulation environments were as follows: $N = 120$, $M \in \{1, 2\}$, and the time-domain channel has four taps ($L = 4$) with equal power profile. For generating a time-varying fading channel, Jakes' model [19] was used. We considered three code rates, $t/k = \{1, 2/3, 1/2\}$, and the modulation types were QPSK, 8PSK, and 16QAM, respectively, to ensure an identical transmission rate. The conventional systems were also considered for comparison. SISO-C is the conventional OFDM system without any ICI canceling codes; SFBC2-A employs the Alamouti STBC between adjacent subcarriers [20]; and SFBC2-Z employs the rate 1/2 ICI canceling SFBC code and receive code in

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TABLE I

OPTIMAL ICI CANCELING CODES $\mathbf{B}$ ($\mathbf{G} = \mathbf{B}$), FOR SINGLE TRANSMIT ANTENNA SYSTEMS ($M = 1$) WITH VARIOUS (k, t) PARAMETERS.

(2,1)	(2,2)		(4,2)	
			$0.2624 + 1.3219i$	$0.7612 + 0.6713i$
$-0.6308 - 0.0105i$	$0.5801 + 0.0340i$	$1.1009 - 0.2271i$	$-0.1570 - 0.1436i$	$0.2680 + 0.0239i$
$1.2651 + 0.0378i$	$-0.2908 - 0.0207i$	$-0.5501 + 0.1058i$	$0.2822 - 0.0899i$	$-0.8709 - 0.2996i$
			$-0.1499 - 0.2312i$	$0.1540 - 0.0313i$

(3,2)		(3,3)		
$0.8618 - 0.0741i$	$-0.4404 + 0.2886i$	$-0.0608 - 0.3984i$	$-0.1648 - 0.1745i$	$0.0177 - 0.8015i$
$0.3900 - 0.4439i$	$-0.1129 - 0.7527i$	$0.1148 + 0.4205i$	$0.1123 - 0.0147i$	$-0.0685 + 0.5150i$
$-0.4260 + 0.4663i$	$0.1135 + 0.7964i$	$-0.4012 - 0.5897i$	$0.0543 + 0.8122i$	$0.2995 + 0.6352i$

TABLE II

OPTIMAL ICI CANCELING SFBCS $\mathbf{B}$, AND RECEIVE CODES $\mathbf{G}$ FOR TWO TRANSMIT ANTENNA SYSTEMS ($M = 2$) WITH VARIOUS (k, t) PARAMETERS.

(A) SFBC $\mathbf{B}$

(2,1)	(2,2)		(4,2)	
			$-0.0241 - 0.2123j$	$-0.0650 - 0.2189j$
			$0.2440 + 0.3590j$	$-0.2787 + 0.7998j$
$0.1832 - 0.6179j$	$-0.3133 + 0.2479j$	$0.2751 - 0.5288j$	$-0.1588 - 0.2584j$	$-0.0069 - 0.2136j$
$-0.0954 + 0.3068j$	$0.5970 - 0.3796j$	$-0.0708 - 0.0093j$	$0.1424 + 0.7935j$	$0.3810 - 0.3780j$
$-0.7006 - 0.8341j$	$0.2272 + 0.5330j$	$-0.3722 - 0.0666j$	$0.2236 - 0.0189j$	$-0.1248 - 0.2311j$
$0.3437 + 0.4203j$	$-0.0364 + 0.0276j$	$0.7058 + 0.0259j$	$-0.7282 - 0.4445j$	$0.3528 + 0.2668j$
			$0.2170 + 0.0844j$	$-0.1715 - 0.1783j$
			$0.2848 + 0.3565j$	$0.0882 + 0.7343j$

(3,2)		(3,3)		
$0.0059 - 0.7936j$	$0.0154 - 0.2128j$	$0.0357 + 0.0601j$	$0.4133 - 0.0399j$	$0.2247 + 0.4053j$
$-0.1975 - 0.0644j$	$0.4691 + 0.2405j$	$-0.0187 - 0.1297j$	$-0.4388 + 0.0729j$	$-0.0920 - 0.3035j$
$0.2442 + 0.2296j$	$-0.5679 - 0.2889j$	$-0.0729 + 0.4238j$	$0.0270 - 0.2693j$	$-0.6796 - 0.2428j$
$-0.0674 - 0.0946j$	$0.4134 - 0.7443j$	$0.1608 + 0.6242j$	$0.1648 + 0.1290j$	$-0.0871 - 0.2167j$
$0.5050 + 0.1387j$	$0.0382 - 0.0122j$	$-0.1446 - 0.6146j$	$-0.1590 - 0.2155j$	$0.1331 + 0.2304j$
$-0.6356 - 0.1565j$	$-0.1465 + 0.1453j$	$0.1890 - 0.0434j$	$0.0622 + 0.5838j$	$-0.2493 - 0.1143j$

(B) Receiver codes $\mathbf{G}$

(2,1)	(2,2)		(4,2)	
			$0.0041 + 0.1761j$	$0.2898 + 0.3644j$
$-0.9914 + 0.7869j$	$-0.4027 - 0.6678j$	$0.3498 - 0.5110j$	$-0.4365 - 0.6086j$	$-0.4148 - 1.0945j$
$0.4992 - 0.3857j$	$0.5246 - 0.0419j$	$-0.8424 + 0.1472j$	$0.0424 + 0.3216j$	$0.3508 + 0.3498j$
			$0.9184 - 0.5039j$	$-0.6002 + 0.1180j$

(3,2)		(3,3)		
$-0.4730 + 0.7154j$	$0.4658 - 0.6641j$	$-0.3691 + 0.5022j$	$0.1329 - 0.5982j$	$-0.3760 - 0.0166j$
$-0.1875 + 0.5052j$	$-0.1590 + 0.5468j$	$0.3357 - 0.3255j$	$-0.2643 + 0.5810j$	$0.5274 - 0.1073j$
$0.3461 - 0.7657j$	$0.1272 - 0.5189j$	$0.3873 + 0.2208j$	$0.8640 - 0.2069j$	$0.1697 + 0.4023j$

[8]. Table III summarizes the systems that will be compared through the simulations. For all schemes, the transmitted vector $\mathbf{s}(u)$ is detected based on the ML criterion at the receiver.

Figure 2 shows the BER of the systems in Table III. In this figure, the $f_d T_s$ varies from 0 to 0.5 and SNR $= 25$ dB. The systems with ICI canceling codes perform better than the conventional SISO-C system when the $f_d T_s$ becomes larger, and the proposed SISO-1 outperforms the others when $f_d T_s \geq 0.0394$. This result indicates that when the normalized Doppler frequency is high, a smaller code rate t/k and smaller code block size t provide better ICI cancelation and BER performances. Figure 3 compares the BER of the systems in Table III (A) for $f_d T_s = 0.2$ and the SNR from 0 to 30 dB. SISO-1 also performs the best when SNR ≥ 11.32 dB. SISO-C, SISO-3, and SISO-5 show the error floors caused by the ICI. The proposed $(3, 2)$ and $(3, 3)$ codes cannot effectively cancel the ICI. From the simulation results in Figs. 2 and 3, we can conclude that using the $(2, 1)$ code is preferable for SISO-OFDM systems.

Figure 4 compares the BER of the systems in Table III for SNR = 25 dB and $f_d T_s$ from 0 to 0.5. The conventional SFBC-A performs the best for $f_d T_s < 0.1$. However, when $f_d T_s > 0.1$, the proposed SFBC2-4 provides the best BER performance. If the $f_d T_s$ is fixed at 0.2 and the SNR varies form 0 to 30 dB, almost all techniques exhibit the BER floor except the proposed SFBC2-1 and SFBC2-4, as shown in Fig. 5. Both methods have a code rate of 1/2. Thus, for MISO-OFDM systems similarly to the SISO-OFDM case, by decreasing the code rate, a more exact ICI cancelation and a lower BER at a high SNR can be achieved at the expense of spectral efficiency. The simulation results in Figs. 2–5 indicate that the proposed SFBC and RFBC can relieve the performance loss resulting from the ICI in fast fading environments.

TABLE III

LIST OF SYSTEMS CONSIDERED IN SIMULATIONS.

Name	Codes	Name	Codes
SISO-1	(2,1) in Table I	SFBC2-1	(2,1) in Table II
SISO-2	(2,2) in Table I	SFBC2-2	(2,2) in Table II
SISO-3	(3,2) in Table I	SFBC2-3	(3,2) in Table II
SISO-4	(4,2) in Table I	SFBC2-4	(4,2) in Table II
SISO-5	(3,3) in Table I	SFBC2-5	(3,3) in Table II
SISO-C	$\mathbf{B} = \mathbf{G} = 1$ without coding	SFBC2-A	Alamouti-SFBC
SISO-Z	$\mathbf{B} = \mathbf{G} = [1, -1]^T$ (2,1) code using [8]	SFBC2-Z	$\mathbf{B} = \frac{1}{\sqrt{2}}[1, -1, 1, -1]^T$ $\mathbf{G} = [1, -1]^T$ (2,1) code using [8]

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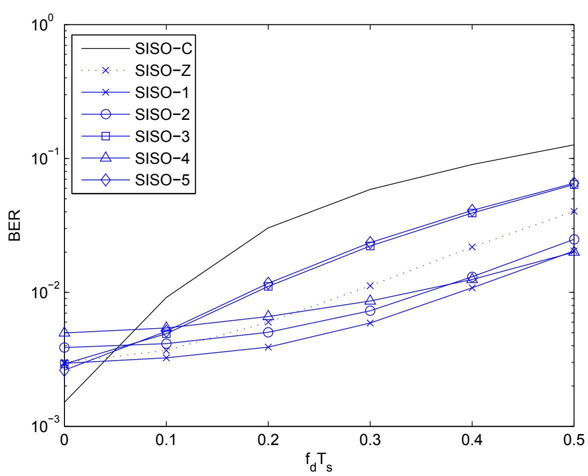

Fig. 2. BER performance of proposed SFBC code versus f_dT_s for one transmit antenna ($L = 4$ and SNR $= 25$ dB.)

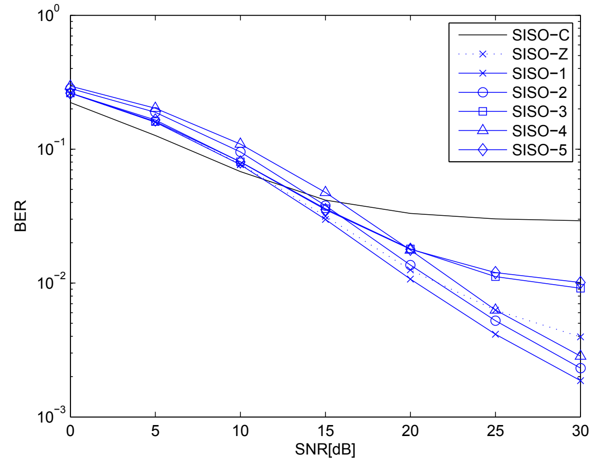

Fig. 3. BER performance of proposed SFBC code versus SNR for one transmit antenna ($L = 4$ and $f_dT_s = 0.2$).

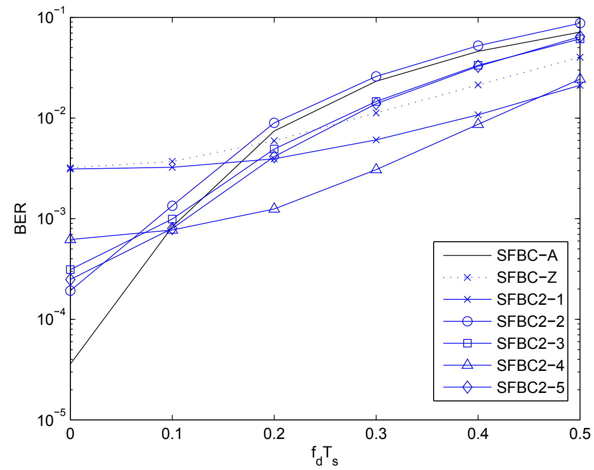

Fig. 4. BER performance of proposed SFBC code versus f_dT_s for two transmit antennas ($L = 4$ and SNR $= 25$ dB).

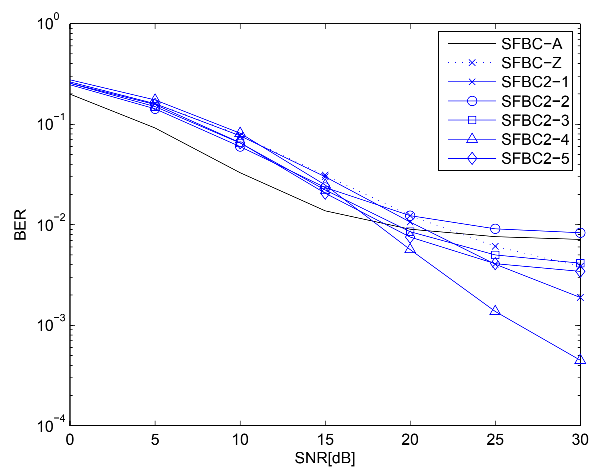

Fig. 5. BER performance of proposed SFBC code versus SNR for two transmit antennas versus ($L = 4$ and $f_dT_s = 0.2$).

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