Reduced Complexity Signal Detection for OFDM Systems with Transmit Diversity

Jaekwon Kim, Robert W. Heath Jr., and Edward J. Powers

Abstract: **Orthogonal frequency division multiplexing (OFDM) systems with multiple transmit antennas can exploit space-time block coding on each subchannel for reliable data transmission. Space-time coded OFDM systems, however, are very sensitive to time variant channels because the channels need to be static over multiple OFDM symbol periods. In this paper, we propose to mitigate the channel variations in the frequency domain using a linear filter in the frequency domain that exploits the sparse structure of the system matrix in the frequency domain. Our approach has reduced complexity compared with alternative approaches based on time domain block-linear filters. Simulation results demonstrate that our proposed frequency domain block-linear filter reduces computational complexity by more than a factor of ten at the cost of small performance degradation, compared with a time domain block-linear filter.**

Index Terms: **Equalization, fast fading, orthogonal frequency division multiplexing (OFDM), transmit diversity.**

I. INTRODUCTION

Recently, transmit diversity techniques have received attention because they improve data transmission reliability without either reducing bandwidth efficiency or increasing transmit power [1], [2]. Transmit diversity is most commonly implemented using one of two types of space time codes: Space time trellis codes (STTC) [3] or space time block codes (STBC) [4], [5]. Unlike STTCs that require the quite complex vector Viterbi decoding at the receiver, STBCs can be designed such that only linear processing is required at the receiver. This paper focuses on the Alamouti code [4] - the most well known STBC - due to its simple decoding capability and acceptance in a number of wireless standards.

Although STBCs were developed assuming flat fading channels, they can be easily extended to frequency selective channels with the aid of orthogonal frequency division multiplexing (OFDM) which converts a frequency selective channel into a set of parallel flat fading channels. The basic idea is to treat each subcarrier as an independent channel and to code over multiple OFDM symbol periods, assuming that the channel is quasi-static over the space-time block codeword period. This allows the advantages of STBCs to be reaped in frequency selective channels [6]–[9].

Unfortunately, when the channels are fast fading, the quasi-static channel condition is not met. Consequently, this causes a severe performance degradation of space-time block coded OFDM systems. Note that the quasi-static requirement of STBC becomes more strict on channels when using OFDM modulation compared with single carrier modulation (SCM) systems because an OFDM symbol is longer than in single carrier systems by the factor of the fast Fourier transform (FFT) size[1]. Furthermore, as the order of STBC increases employing more transmit antennas, a space-time codeword period becomes even longer. In [10], they reported that STBC techniques can be used for single carrier systems in time-variant channels, which is possible due in part to a short symbol period.

A differential transmission scheme for the STBC-OFDM was reported [11] that would be very beneficial under the fast fading channel environment because channel state information (CSI) is not necessary (hence, no pilot symbols are necessary). The differential scheme, however, still requires the channels to be static over two space time codeword periods. Thus the differential scheme still suffers from severe performance degradation under the higher speed channel variations as considered in this paper [12].

In an effort to mitigate the effects of time-varying channels, a time domain block-linear filter (TDBLF) was proposed [13]. The block-linear filter design as well as the filtering process, however, is computationally quite expensive. In [12], a sequential decision feedback sequence estimation (SDFSE) with an adaptive threshold (AT), a traditional single carrier equalization technique [2], was applied to the problem and was shown to provide a good tradeoff between the performance and complexity. The main drawback of the sequence estimation, though, is that the complexity is time-variant that is not amenable to hardware implementation. Furthermore, the complexity is dependent on the constellation size.

In [14], the complexity of [13] is reduced by exploiting the inter-carrier interference (ICI) generation mechanism. An optimal linear preprocessing is used to restrict the ICI support in the frequency domain, and an iterative minimum mean-squared error estimation is performed. While the complexity of [13] is prohibitively high when the OFDM symbol is long, the complexity of [14] is linear in the OFDM symbol length. In [15], the band structure of the frequency domain channel matrix is exploited in order to reduce the complexity. Unlike in [13], [14] where the subchannels are equalized separately, all the subchanels are

Manuscript received March 12, 2006; approved for publication by Kyungwhoon Cheun, Division II Editor, August 17, 2006.

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[1]When the spectral bandwidth B is given, the symbol time in single carrier systems is $T_s = 1/B$. In OFDM systems, however, the symbol time becomes $T_s \times N$, where N is the FFT size. In OFDM systems, the channel is assumed to be static over the time $T_s \times N$.

[2]There exists a duality between the inter-carrier interference (ICI) in OFDM and inter-symbol interference (ISI) in single carrier systems. Consequently, the ISI compensation techniques for single carrier systems can also be used to deal with ICI in OFDM.

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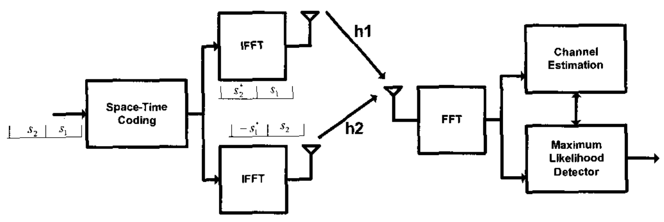

Fig. 1. A block diagram of OFDM systems with transmit diversity.

jointly equalized in [15].

In this paper, we propose a frequency domain block-linear filter (FDBLF) whose length is much shorter than the TDBLF [13], thereby achieving a significant complexity reduction in the filter design as well as in the filtering process. First, we extend the block diagonal approximation the of system matrix of single-input single-output (SISO) system [16] to the multiple-input multiple-output (MIMO) case. Using the extended block diagonal approximation, we divide a large system equation into a set of small system equations in the frequency domain. Then, we design a small FDBLF for each of the small system equations. We show that the FDBLF is computationally much more efficient than the previous TDBLF. Since different signal detection algorithms react differently to CSI estimation error, the CSI estimator described in [13] is adopted for fairness for the comparison in terms of error performance. Monte Carlo simulations suggest that the proposed FDBLF shows a negligible performance degradation while requiring much less computations, compared with the previous TDBLF. We show that the error performance gap between the two schemes becomes even smaller when CSI estimation error is included.

The remainder of the paper is organized as follows. Section II describes the system that is considered in the paper. The two transmit antennas and one receive antenna system is described in detail. Section III briefly reviews the related time domain approach. Section IV proposes our new frequency domain block linear filter. Section V compares the two schemes in terms of computational complexity. The two schemes are compared in terms of error performance via Monte Carlo simulations in Section VI. Finally, Section VII presents some conclusions.

Notation: In this paper, a bold face letter denotes a vector or a matrix as will be clear from the context; $\mathbf{I}_M$ denotes the $M \times M$ identity matrix; $(\cdot)^*$ denotes complex conjugate; $(\cdot)^T$ denotes transpose; $(\cdot)^H$ denotes Hermitian (conjugate transpose); $|\cdot|$ denotes absolute value; in general, a lowercase letter stands for a time domain signal while an upper case letter denotes frequency domain signal; $Diag(\cdot)$ with argument matrices denotes a block diagonal matrix; $[\cdot]_k$ denotes the k-th element of the argument vector; $[\cdot]_{k,m}$ denotes the (k,m)-th entry of the argument matrix.

II. SYSTEM DESCRIPTION

In this section, we illustrate the two transmit antennas and one receive antenna OFDM system under consideration in Fig. 1. When the Alamouti code [4] is used as a space-time code at each subchannel, assuming perfect carrier and timing synchronization, the relationship between the transmitted signal and the received signal can be described as

$$\underbrace{\begin{bmatrix}\mathbf{Y}_1\\ \mathbf{Y}_2^*\end{bmatrix}}_{\triangleq \mathbf{Y}} = \mathbf{Q}^{(Rx)} \underbrace{\begin{bmatrix}H_{11}\mathbf{Q}^H & H_{12}\mathbf{Q}^H\\ H_{22}^*\mathbf{Q} & -H_{21}^*\mathbf{Q}\end{bmatrix}}_{\triangleq \mathbf{H}} \underbrace{\begin{bmatrix}\mathbf{X}_1\\ \mathbf{X}_2\end{bmatrix}}_{\triangleq \mathbf{X}} + \underbrace{\begin{bmatrix}\mathbf{Z}_1\\ \mathbf{Z}_2^*\end{bmatrix}}_{\triangleq \mathbf{Z}} \quad (2)$$

where

$$\mathbf{X}_i \triangleq [X_i(0)\; X_i(1)\; \cdots\; X_i(N-1)]^T, \quad i = 1, 2$$
$$\mathbf{Y}_l \triangleq [Y_l(0)\; Y_l(1)\; \cdots\; Y_l(N-1)]^T, \quad l = 1, 2$$
$$\mathbf{Z}_l \triangleq [Z_l(0)\; Z_l(1)\; \cdots\; Z_l(N-1)]^T, \quad l = 1, 2$$
$$\mathbf{Q}^{(Rx)} \triangleq Diag(\mathbf{Q}, \mathbf{Q}^H)$$

and the H_{li} is defined at the bottom of this page.

The parameter N is the FFT size. The symbol $X_i(k)$ is the complex symbol transmitted over the k-th subchannel from the i-th transmit antenna. The symbol $Y_l(k)$ is the received signal at the k-th subchannel during the l-th symbol period. The noise $Z_l(\cdot)$ is considered as circularly symmetric complex Gaussian random variable with variance of $\sigma_z^2/2$ per dimension. When the cyclic prefix (CP) of length at least $L-1$ is considered, the Toeplitz channel matrix H_{li} describes the channel between the i-th transmit antenna and the receive antenna during the l-th symbol period. The symbol $h_{l,i}(n;m)$, $0 \le n \le N-1$, $0 \le m \le L-1$ is the m-th multipath gain at the sampling time of nT_s. The length of the two channels is assumed to be the same as LT_s, where $T_s = 1/B$ and B is the spectral bandwidth of the system. The matrix $\mathbf{Q}$ is the $N \times N$ FFT matrix whose the (m,n)-th entry is $Q(m,n) = \frac{1}{\sqrt{N}} e^{-j2\pi mn/N}$, $0 \le m, n \le N-1$.

If the channels are quasi-static, the following is true

$$\left[\mathbf{Q}^{(Rx)}\mathbf{H}\right]_{k,m} = 0 \text{ if } (k)_N \neq (m)_N \quad (3)$$

where $(\cdot)_N$ denotes the modulo operation. Consequently, space-time decoding can be performed for each subchannel separately, using the simple Alamouti decoding.

When the channels are fast fading, however, the approximation in (3) does not hold, leading to severe performance degradation of the Alamouti decoding. In the following sections, signal detection methods in fast fading channels are discussed.

Extension to more than one receive antenna system is straightforward using linear combining at the receiver; this is considered in Subsection IV-C.

$$H_{li} \triangleq \begin{bmatrix} h_{l,i}(0;0) & 0 & \cdots & 0 & h_{l,i}(0;L-1) & \cdot\cdot & h_{l,i}(0;1) \\ h_{l,i}(1;1) & h_{l,i}(1;0) & 0 & \cdots & 0 & \cdot\cdot & h_{l,i}(1;2) \\ \vdots & \vdots & \vdots & \ddots & \ddots & \vdots & \vdots \\ 0 & \cdots & 0 & h_{l,i}(N-1;L-1) & \cdots & & h_{l,i}(N-1;0) \end{bmatrix} \quad (1)$$

III. TIME DOMAIN BLOCK-LINEAR FILTER

In this section, we briefly review the previous time domain block-linear filter (TDBLF) that mitigates the channel variations [13]. Assuming two transmit antennas and one receive antenna, a $2N \times 2N$ TDBLF is inserted right before the demodulating FFT, resulting in the following composite matrix

$$\overline{\mathbf{G}}_t \triangleq \mathbf{Q}^{(Rx)}\mathbf{W}_t\mathbf{H} \tag{4}$$

where $\mathbf{W}_t$ is the TDBLF. The off-diagonal entries of $\overline{\mathbf{G}}_t$ are viewed as interference and the following signal to interference plus noise ratio $\mathrm{SINR}_t(k)$ is defined

$$\mathrm{SINR}_t(k) \triangleq \frac{E_x|\mathbf{e}_k^H\overline{\mathbf{G}}_t\mathbf{e}_k|^2}{\frac{\sigma_z^2}{2N}\mathrm{tr}(\mathbf{W}_t\mathbf{W}_t^H) + E_x\sum_{m\neq k}|\mathbf{e}_k^H\overline{\mathbf{G}}_t\mathbf{e}_m|^2} \tag{5}$$

where $\mathrm{tr}(A)$ is the trace of matrix A. If no power control is adopted, the transmit power is equally divided, i.e., $\mathbf{E}[|X_i(k)|^2] = E_x$ for all i and k. The vector $\mathbf{e}_k$ is the k-th column of a $2N \times 2N$ identity matrix, i.e., a length $2N$ with the k-th entry of 1 and zeros elsewhere. It is desired that the TDBLF maximizes $\mathrm{SINR}_t(k),\ k = 0, 1, \cdots, 2N-1$. It was shown [13] that an optimization problem can be formulated as

$$\begin{aligned} &\text{For each } k = 0, 1, \cdots, 2N-1 \\ &\max_{\mathbf{w}_k} \mathbf{w}_k^H\mathbf{h}_k\mathbf{h}_k^H\mathbf{w}_k \qquad (6) \\ &\text{subject to } \mathbf{w}_k^H\left(\frac{\sigma_z^2}{E_x}\mathbf{I}_{2N} + \mathbf{R}_k\right)\mathbf{w}_k = 1 \\ &\text{and } \mathbf{w}_k^H\mathbf{w}_k = 1 \end{aligned}$$

where $\mathbf{w}_k \triangleq \mathbf{W}_t^H\mathbf{q}_k$, $\mathbf{q}_k \triangleq \mathbf{Q}^{(Rx)}\mathbf{e}_k$, $\mathbf{h}_k \triangleq \mathbf{H}\mathbf{e}_k$, and $\mathbf{R}_k \triangleq \frac{\sigma_z^2}{E_s}\mathbf{I}_{2N} + \mathbf{H}\mathbf{H}^H - \mathbf{h}_k\mathbf{h}_k^H$.

This optimization problem can be solved as

$$\begin{aligned} &\text{For each } k = 0, 1, \cdots, 2N-1 \\ &\tilde{\mathbf{w}}_k = \mathbf{R}^{-1}\mathbf{h}_k \qquad (7) \\ &\mathbf{w}_{k,opt} = \tilde{\mathbf{w}}_k/|\tilde{\mathbf{w}}_k| \qquad (8) \end{aligned}$$

where $\mathbf{R} \triangleq \mathbf{H}\mathbf{H}^H + \frac{\sigma_z^2}{E_x}\mathbf{I}_{2N}$.

Note that only one matrix inversion is necessary in the filter design process for all k. From this filter design procedure, the following $2N \times 2N$ TDBLF is obtained

$$\mathbf{W}_t = \mathbf{Q}^{(Rx)H}\left[\mathbf{w}_{0,opt}\ \mathbf{w}_{1,opt}\ \cdots\ \mathbf{w}_{2N-1,opt}\right]^H. \tag{9}$$

Finally, transmitted signals are estimated using

$$\begin{aligned} \hat{X}(k) &= \frac{[\mathbf{Q}^{(Rx)}\mathbf{W}_t\mathbf{Y}]_k}{\overline{G}_t(k,k)} \qquad (10) \\ &= \frac{\mathbf{w}_{k,opt}^H\mathbf{Y}}{\overline{G}_t(k,k)},\quad k = 0, 1, \cdots, 2N-1 \qquad (11) \end{aligned}$$

where $\hat{X}(k)$ is the estimated symbol of $X(k)$ and $\overline{G}_t(k,k)$ is the k-th diagonal entry of $\overline{\mathbf{G}}_t$. Note that the $(k+N)$-th entry of $\mathbf{X}$ is the k-th entry of $\mathbf{X}_2$. From (11), we can see that $\mathbf{W}_t$ does not have to be calculated explicitly.

IV. FREQUENCY DOMAIN BLOCK-LINEAR FILTER

In this section, a frequency domain block-linear filter (FDBLF) is proposed whose size $(4q+2)\times(4q+2)$ is much smaller than the time domain filter in Section III. The parameter q is the bandwidth of a matrix that is described in detail in Subsection IV-A. First, Subsection IV-A extends a previous block diagonal approximation of SISO system to MIMO case. Based on the extended block diagonal approximation, Subsection IV-B proposes a FDBLF. Finally, Subsection IV-C considers receiver combining for more than one receive antenna systems.

A. Block Diagonal Approximation

We investigate the structure of the following matrix

$$\begin{aligned} \mathbf{G} &\triangleq \begin{bmatrix} \mathbf{G}_{11} & \mathbf{G}_{12} \\ \mathbf{G}_{21} & \mathbf{G}_{22} \end{bmatrix} \triangleq \begin{bmatrix} \mathbf{Q}H_{11}\mathbf{Q}^H & \mathbf{Q}H_{12}\mathbf{Q}^H \\ \mathbf{Q}^H H_{22}^*\mathbf{Q} & -\mathbf{Q}^H H_{21}^*\mathbf{Q} \end{bmatrix} \\ &= \mathbf{Q}^{(Rx)}\mathbf{H}. \end{aligned} \tag{12}$$

Note that $\mathbf{G}_{il},\ i,l = 1,2$ denotes the subsystem reflecting the channel between the i-th transmit antenna and the receive antenna during the l-th symbol period. Instead of analyzing all the entries of $\mathbf{G}$, we investigate the submatrix $\mathbf{G}_{11}$. The (k,m)-th entry of $\mathbf{G}_{11}$ is expressed as

$$G_{11}(k,m) = \frac{1}{\sqrt{N}}\sum_{p=0}^{L-1}\mathcal{H}_p(k,m)e^{-j2\pi kp/N} \tag{13}$$

where

$$\mathcal{H}_p(k,m) \triangleq \frac{1}{\sqrt{N}}\sum_{n=0}^{N-1}h_{1,1}(n,p)e^{-j2\pi n(k-m)/N}. \tag{14}$$

From (13), an upper bound of $|G_{11}(k,m)|$ is derived as

$$\begin{aligned} |G_{11}(k,m)| &\le \frac{1}{\sqrt{N}}\sum_{p=0}^{L-1}\left|\mathcal{H}_p(k,m)e^{-j2\pi kp/N}\right| \qquad (15) \\ &= \frac{1}{\sqrt{N}}\sum_{p=0}^{L-1}\left|\mathcal{H}_p(k,m)\right|. \qquad (16) \end{aligned}$$

Note that $\mathcal{H}_p(k,m)$ is the $(k-m)$-th harmonic frequency coefficient of the time-variant p-th resolved multipath. It was shown that when the channel variation is not severe, each path can be assumed to change in a linear fashion [16]. If a multipath changes in a linear fashion, $|\mathcal{H}_p(k,m)|$ decreases dramatically as $|k-m|$ increases. It was also shown that even when the channel variation is rather severe, ICI from far away subchannels may be ignored because of channel estimation error as well as its less significance in magnitude [12]. Therefore, each submatrix of $\mathbf{G}$ can be approximated as a band matrix with both lower and upper bandwidth of q [17], i.e.,

$$|G_{il}(k,m)| \approx 0 \text{ if } |k-m| > q,\quad 1 \le i,l \le 2. \tag{17}$$

Using (17), we perform a block diagonal approximation that is illustrated for $q = 1$ in Fig. 2. Fig. 2 shows that submatrices are approximately band matrices. The dimension of the original

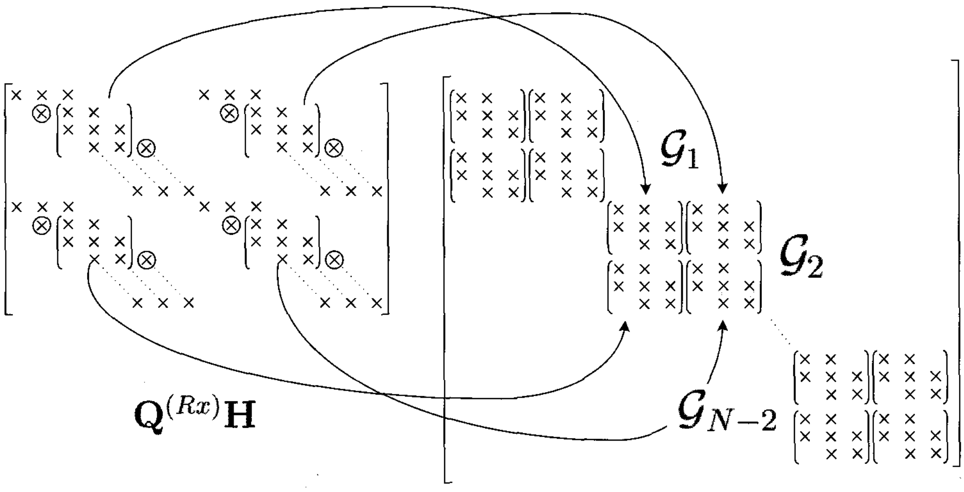

Fig. 2. Illustration of a block diagonal approximation when $q = 1$. The entries that are significant but are approximated as 0 in the block diagonal approximation are denoted as circled crosses.

matrix is $2N \times 2N$ and that of the block diagonal approximated matrix is $(4q+2)(N-2q) \times (4q+2)(N-2q)$. We observe that some significant terms are ignored in the approximation process. A general block diagonal approximation can be expressed as

$$\mathcal{Y} \approx \mathcal{G}\mathcal{X} + \mathcal{Z} \tag{18}$$

where

$$\begin{aligned}
\mathcal{Y} &\triangleq [\mathbf{Y}_{1,q}^T\ \mathbf{Y}_{2,q}^H \cdots \mathbf{Y}_{1,N-1-q}^T\ \mathbf{Y}_{2,N-1-q}^H]^T \\
\mathcal{X} &\triangleq [\mathbf{X}_{1,q}^T\ \mathbf{X}_{2,q}^T \cdots \mathbf{X}_{1,N-1-q}^T\ \mathbf{X}_{2,N-1-q}^T]^T \\
\mathcal{Z} &\triangleq [\mathbf{Z}_{1,q}^T\ \mathbf{Z}_{2,q}^H \cdots \mathbf{Z}_{1,N-1-q}^T\ \mathbf{Z}_{2,N-1-q}^H]^T \\
\mathbf{Y}_{l,k} &\triangleq [Y_l(k-q) \cdots Y_l(k)\ \cdots\ Y_l(k+q)]^T \\
\mathbf{X}_{i,k} &\triangleq [X_l(k-q) \cdots X_l(k)\ \cdots\ X_l(k+q)]^T \\
\mathbf{Z}_{l,k} &\triangleq [Z_l(k-q)\ \cdots\ Z_l(k) \cdots Z_l(k+q)]^T \\
\mathcal{G} &\triangleq Diag(\mathcal{G}_q,\ \mathcal{G}_{q+1}, \cdots, \mathcal{G}_{N-1-q}) \\
\mathcal{G}_k &\triangleq \begin{bmatrix} \mathbf{G}_{11,k} & \mathbf{G}_{21,k} \\ \mathbf{G}_{22,k} & \mathbf{G}_{12,k} \end{bmatrix}.
\end{aligned}$$

The entries of the matrix $\mathcal{G}_k \in \mathbb{C}^{(4q+2)\times(4q+2)}$, i.e., $G_{il,k}(m,n)$, $0 \le m,n \le 2q$, $i,l = 1,2$ are defined as

$$\begin{cases} G_{il}(k-q+m, k-q+n, & \text{if } |k-m| \le q, \\ 0 & , \text{ otherwise.} \end{cases} \tag{19}$$

Once the block diagonal approximation is done, the large equation (18) can be divided into the following small equations for each $k = q, q+1, \cdots, N-q-1$

$$\mathcal{Y}_k \approx \mathcal{G}_k\mathcal{X}_k + \mathcal{Z}_k \tag{20}$$

where $\mathcal{X}_k \triangleq [\mathbf{X}_{1,k}^T\ \ \mathbf{X}_{2,k}^T]^T$, $\mathcal{Y}_k \triangleq [\mathbf{Y}_{1,k}^T\ \ \mathbf{Y}_{2,k}^H]^T$, and $\mathcal{Z}_k \triangleq [\mathbf{Z}_{1,k}^T\ \ \mathbf{Z}_{2,k}^H]^T$. We note that the index k in (20) is from q to $N-q-1$ because q virtual carriers are assumed at both ends of the spectrum.

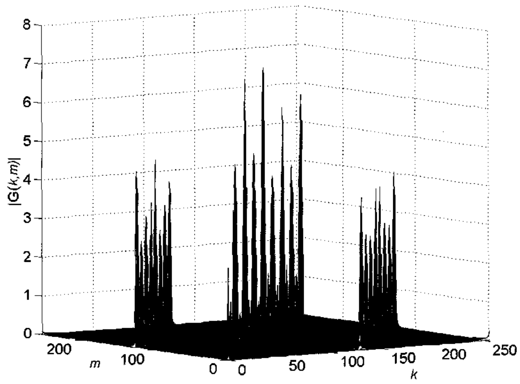

Fig. 3. An Example of $|G(k,m)|$, $0 \le k, m \le 2N-1$.

Numerical Example: To present the feasibility of the block diagonal approximation, a numerical example of the sparse structure of the system matrix is provided. Fig. 3 is an illustration of instantaneous magnitude of entries of the system matrix. The FFT size is 128; a spectral bandwidth of 400 kHz is assumed; the Doppler frequency is 297 Hz which leads to the normalized Doppler frequency of 0.1188 ($f_D(N+\nu)T_s$), where f_D is the Doppler frequency, ν is the CP length, T_s is the sample time (the inverse of the spectral bandwidth). Equal gain two path channels with delays of 0 and $4T_s$ were simulated. As can be observed in Fig. 3, the submatrices in (12) are close to band matrices. The off diagonal terms become smaller as the entry position is further from the diagonal. We can also observe the frequency selectivity of the channels, i.e., for a given i and l, $\mathbf{G}_{il}[k,k]$ changes as k increases from 0 to $N-1$.

B. FDBLF

In this subsection, a frequency domain block linear filter (FDBLF) is proposed that exploits the block diagonal approximation in the previous subsection.

$$\mathbf{G}_{il,k}^{upper} \triangleq \begin{bmatrix} G_{il}(k-q,k-2q) & G_{il}(k-q,k-2q+1) & \cdots & G_{il}(k-q,k-q-1) \\ 0 & G_{il}(k-q+1,k-2q+1) & \cdots & G_{il}(k-q+1,k-q-1) \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & G_{il}(k-1,k-q-1) \end{bmatrix} \tag{21}$$

$$\mathbf{G}_{il,k}^{lower} \triangleq \begin{bmatrix} G_{il}(k+1,k+q+1) & 0 & \cdots & 0 \\ G_{il}(k+2,k+q+1) & G_{il}(k+2,k+q+2) & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ G_{il}(k+q,k+q+1) & \cdots & \cdots & G_{il}(k+q,k+2q) \end{bmatrix} \tag{22}$$

In the process of the block diagonal approximation, not only the negligible terms $G_{il}(k,m)$ with $|k-m|>q$ but also some significant terms $G_{il}(k,m)$ with $|k-m|\le q$ are ignored.

Including the significant terms also, a more exact expression is

$$\mathcal{Y}_k = \mathcal{G}_k\mathcal{X}_k + \mathcal{G}_k^{upper}\mathcal{X}_{k-q} + \mathcal{G}_2^{lower}\mathcal{X}_{k+q} + \mathcal{Z}_k \tag{23}$$

where

$$\mathcal{G}_k^{upper} \triangleq \begin{bmatrix} \mathbf{G}_{11,k}^{upper} & \mathbf{0}_{q\times(q+1)} & \mathbf{G}_{21,k}^{upper} & \mathbf{0}_{q\times(q+1)} \\ \mathbf{0}_{(q+1)\times q} & \mathbf{0}_{(q+1)\times(q+1)} & \mathbf{0}_{(q+1)\times q} & \mathbf{0}_{(q+1)\times(q+1)} \\ \mathbf{G}_{22,k}^{upper} & \mathbf{0}_{q\times(q+1)} & \mathbf{G}_{12,k}^{upper} & \mathbf{0}_{q\times(q+1)} \\ \mathbf{0}_{(q+1)\times q} & \mathbf{0}_{(q+1)\times(q+1)} & \mathbf{0}_{(q+1)\times q} & \mathbf{0}_{(q+1)\times(q+1)} \end{bmatrix}, \tag{24}$$

and

$$\mathcal{G}_k^{lower} \triangleq \begin{bmatrix} \mathbf{0}_{(q+1)\times(q+1)} & \mathbf{0}_{(q+1)\times q} & \mathbf{0}_{(q+1)\times(q+1)} & \mathbf{0}_{(q+1)\times q} \\ \mathbf{0}_{q\times(q+1)} & \mathbf{G}_{11,k}^{lower} & \mathbf{0}_{q\times(q+1)} & \mathbf{G}_{21,k}^{lower} \\ \mathbf{0}_{(q+1)\times(q+1)} & \mathbf{0}_{(q+1)\times q} & \mathbf{0}_{(q+1)\times(q+1)} & \mathbf{0}_{(q+1)\times q} \\ \mathbf{0}_{q\times(q+1)} & \mathbf{G}_{22,k}^{lower} & \mathbf{0}_{q\times(q+1)} & \mathbf{G}_{12,k}^{lower} \end{bmatrix}. \tag{25}$$

The matrices $\mathbf{G}_{il,k}^{upper}$ and $\mathbf{G}_{il,k}^{lower}$ are defined at the top of this page.

The first term on the right side of (23) is regarded as desired component, the second and the third terms are considered as interference. Defining a new noise term, we rewrite (23) as

$$\mathcal{Y}_k = \mathcal{G}_k\mathcal{X}_k + \tilde{\mathcal{Z}}_k \tag{26}$$

where

$$\tilde{\mathcal{Z}}_k \triangleq \mathcal{G}_k^{upper}\mathcal{X}_{k-q}^{upper} + \mathcal{G}_k^{lower}\mathcal{X}_{k+q}^{lower} + \mathcal{Z}_k. \tag{27}$$

If we design a block linear filter of dimension $(4q+2)\times(4q+2)$ and apply to (26), we obtain

$$\mathbf{V}_k\mathcal{Y}_k = \underbrace{\mathbf{V}_k\mathcal{G}_k}_{\triangleq\bar{\mathcal{G}}_k}\mathcal{X}_k + \mathbf{V}_k\tilde{\mathcal{Z}}_k. \tag{28}$$

Similar to the TDBLF case, it is desired that $\bar{\mathcal{G}}_k$ becomes a diagonal matrix. To derive an SINR expression, we evaluate the noise variance after filtering in (28)

$$\mathbf{E}\left\{\left|\tilde{\mathcal{Z}}_k(p)\right|^2\right\} \triangleq \mathbf{E}\left|\mathbf{e}_p^H\mathbf{V}_k\tilde{\mathcal{Z}}_k\right|^2 \tag{29}$$

$$= \mathbf{v}_{k,p}^H\mathbf{E}\left\{\tilde{\mathcal{Z}}_k\tilde{\mathcal{Z}}_k^H\right\}\mathbf{v}_{k,p} \tag{30}$$

where $\mathbf{v}_{k,p} \triangleq \mathbf{V}_k^H\mathbf{e}_p$. The above covariance matrix becomes

$$\begin{aligned}\mathbf{E}\left\{\tilde{\mathcal{Z}}_k\tilde{\mathcal{Z}}_k^H\right\} &= \mathbf{E}\{(\mathcal{G}_k^{upper}\mathcal{X}_{k-q} + \mathcal{G}_k^{lower}\mathcal{X}_{k+q} + \mathcal{Z}_k) \\ &\quad\cdot(\mathcal{G}_k^{upper}\mathcal{X}_{k-q} + \mathcal{G}_k^{lower}\mathcal{X}_{k+q} + \mathcal{Z}_k)^H\} \\ &= \sigma_z^2\mathbf{I}_{4q+2} + E_x\cdot\{\mathcal{G}_k^{lower}\mathcal{G}_k^{lower\ H}\} \\ &\quad + E_x\cdot\left\{\mathcal{G}_k^{upper}\mathcal{G}_k^{upper\ H}\right\}.\end{aligned} \tag{31}$$

Considering the off-diagonal terms of $\bar{\mathcal{G}}_k$ as interference, we design the FDBLF to maximize

$$\mathrm{SINR}_{f,k}(p) \triangleq \frac{E_x|\mathbf{e}_p^H\bar{\mathcal{G}}_k\mathbf{e}_p|^2}{\mathbf{v}_{k,p}^H\mathbf{E}\left\{\tilde{\mathcal{Z}}_k\tilde{\mathcal{Z}}_k^H\right\}\mathbf{v}_{k,p} + E_x\sum_{m\neq k}|\mathbf{e}_p^H\bar{\mathcal{G}}_k\mathbf{e}_m|^2} \tag{32}$$

where the vector $\mathbf{e}_p$ is now a unit norm column vector of length $(4q+2)$ with the p-th element of 1.

The solution to this optimization problem can be formulated in a similar way as (6)

$$\begin{aligned}&\text{For each } k = q, q+1,\cdots, N-q-1 \\ &\text{For } p = q, 3q+1 \\ &\max_{\mathbf{v}_{k,p}} \mathbf{v}_{k,p}^H\mathbf{g}_{k,p}\mathbf{g}_{k,p}^H\mathbf{v}_{k,p} = 1 \\ &\quad \text{subject to } \mathbf{v}_{k,p}^H\left(\frac{\sigma_z^2}{E_x}\mathbf{I}_{4q+2} + \mathcal{R}_{k,p}\right)\mathbf{v}_{k,p} = 1 \\ &\quad \text{and } \mathbf{v}_{k,p}^H\mathbf{v}_{k,p} = 1\end{aligned} \tag{33}$$

where $\mathbf{g}_{k,p} \triangleq \mathcal{G}_k\mathbf{e}_p$ and $\mathcal{R}_{k,p} \triangleq \mathcal{G}_k\mathcal{G}_k^H - \mathbf{g}_{k,p}\mathbf{g}_{k,p}^H + \mathcal{G}_k^{lower}\mathcal{G}_k^{lower\ H} + \mathcal{G}_k^{upper}\mathcal{G}_k^{upper\ H}$.

The solution to this optimization is as follows

For each $k = q, q+1,\cdots, N-q-1$

For each $p = q, 3q+1$

$$\mathcal{R}_k = \frac{\sigma_z^2}{E_x}\mathbf{I}_{4q+2} + \mathcal{G}_k\mathcal{G}_k^H + \mathcal{G}_k^{lower}\mathcal{G}_k^{lower\ H} + \mathcal{G}_k^{upper}\mathcal{G}_k^{upper\ H} \tag{34}$$

$$\mathbf{g}_{k,p} = \mathcal{G}_k\mathbf{e}_p \tag{35}$$

$$\tilde{\mathbf{v}}_p = \mathcal{R}_k^{-1}\mathbf{g}_{k,p} \tag{36}$$

$$\mathbf{v}_{p,opt} = \tilde{\mathbf{v}}_p/|\tilde{\mathbf{v}}_p| \tag{37}$$

$$\mathbf{V}(k;:) = \mathbf{v}_{q,opt}^T \tag{38}$$

Table 1. Complexity comparison of the FDBLF and the TDBLF.

	TDBLF		FDBLF	
Filter Design	$\mathbf{R}$ Construction	$(2N)^3$	$\mathcal{R}_k,\ \forall\, k$ Construction	$[(4q+2)^3+8q^3]N$
	$\mathbf{R}^{-1}$	$(2N)^3$	$\mathcal{R}_k^{-1}, \forall\, k$	$(4q+2)^3 N$
	$\mathbf{R}^{-1}\mathbf{h}_k,\ \forall\, k$	$(2N)^3$	$\mathcal{R}_k^{-1}\mathbf{g}_{k,p}, \forall\, k$	$2(4q+2)^2 N$
Filtering	Eq. (11)	$(2N)^2$	Eqs.(40) and (41)	$2(4q+2)N$
Total	$3\times(2N)^3+(2N)^2$		$[2(4q+2)^3+8q^3+2(4q+2)^2+2(4q+2)]N$	

$$\mathbf{V}(k+N;:)=\mathbf{v}_{3q+1,opt}^T. \tag{39}$$

Note that the dimension of the matrix to be inverted (36) is $(4q+2)\times(4q+2)$ that is much smaller than in time domain filter design process. After obtaining the above frequency domain filter, the transmitted signals are estimated as

For each $k=q,q+1,\cdots,N-q-1$

$$\hat{X}(k)=\frac{\mathbf{V}(k;:)\mathcal{Y}_k}{\mathcal{G}_k(q,q)} \tag{40}$$

$$\hat{X}(k+N)=\frac{\mathbf{V}(k+N;:)\mathcal{Y}_k}{\mathcal{G}_k(3q+1,3q+1)}. \tag{41}$$

C. Receiver Combining

When more than one receive antenna exists, an appropriate combining of the received signals is necessary. Although any combining method such as selection combining and switched combining could be used, the maximal ratio receiver combining (MRRC) is adopted since it shows the best performance among the three. If two receive antennas are assumed, transmitted signals are estimated as

For each $k=q,q+1,\cdots,N-q-1$

$$\hat{X}(k)=\frac{\sum_{j=1}^{2}\overline{\mathcal{G}}_{j,k}^*(q,q)\mathbf{V}_j(k;:)\mathcal{Y}_{j,k}}{\sum_{j=1}^{2}\left|\overline{\mathcal{G}}_{j,k}(q,q)\right|^2} \tag{42}$$

$$\hat{X}(k+N)=\frac{\sum_{j=1}^{2}\overline{\mathcal{G}}_{j,k}^*(3q+1,3q+1)\mathbf{V}_j(k;:)\mathcal{Y}_{j,k}}{\sum_{j=1}^{2}\left|\overline{\mathcal{G}}_{j,k}(3q+1,3q+1)\right|^2} \tag{43}$$

where $j=1,2$ indicates the receive antenna index. Note that the filter design and filtering process for a multiple receive antenna system is the same as the single receive antenna case. Receiver combining for the previous TDBLF can be done in a similar way.

V. COMPLEXITY COMPARISON

This section compares the previous TDBLF and the proposed FDBLF in terms of computational complexity. One receive antenna is assumed because more than one receive antenna does not change the relative complexity of the two schemes. First, we compare the design complexity of the TDBLF (7) and FDBLF (36). As can be seen in (7), TDBLF requires a $2N\times 2N$ matrix inversion ($(2N)^3$ multiplications) and a matrix vector multiplication ($(2N)^2$) for each $k=0,1,\cdots,2N-1$. If we include the cost of the $\mathbf{R}$ matrix construction ($(2N)^3$), the design of a TDBLF costs $3\times(2N)^3$ multiplications in total. The frequency domain filter design procedure requires a $(4q+2)\times(4q+2)$ matrix inversion ($(4q+2)^3$) and a matrix vector multiplication ($(4q+2)^2$) for each $k=q,q+1,\cdots,N-q-1$. If we include the cost of the $\mathcal{R}_k$ matrix construction ($(4q+2)^3+8q^3$), FDBLF design costs $N[3(4q+2)^3+8q^3+2(4q+2)^2]$ multiplications in total.

Now, we compare the filtering process (11), (40), and (41). The time domain filtering costs $(2N)^2$ multiplications and the frequency domain filtering costs $2N(4q+2)$ multiplications. Combining the costs of filter design and filtering procedures, the ratio of the computational complexity of the two schemes can be expressed as

$$\begin{aligned}R(q,N)&=\frac{\text{FDBLF Cost}}{\text{TDBLF Cost}}\\&=\frac{2(4q+2)^3+8q^3+2(4q+2)^2+2(4q+2)}{24N^2+4N}.\end{aligned} \tag{44}$$

The computational complexity is summarized in Table 1. Note that when a Doppler frequency is fixed, the normalized Doppler frequency decreases as the FFT size decreases and that when the normalized Doppler frequency is small, a smaller q can be used because $2q$ is the number of nearest subchannels causing interchannel interference. Therefore, a smaller N indicates that a smaller q is allowed.

VI. SIMULATION RESULTS AND DISCUSSION

In this section, simulation results are provided to compare the error performance versus signal-to-noise ratio (SNR) of the time domain and the frequency domain block-linear filters. A two transmit antenna and one receive antenna system is considered. Assuming multipaths are independent of each other and wide sense stationary, the SNR is defined as

$$\text{SNR}=\frac{E_x\sum_{i=1}^{2}\sum_{p=0}^{L-1}\mathbf{E}\left|h_i(p)\right|^2}{\sigma_z^2} \tag{45}$$

where $h_i(p)$ stands for the p-th resolved multipath between the i-th transmit antenna and the receive antenna.

The FFT size N is 128 and CP length ν is 32. The spectral bandwidth $(1/T_s)$ is 400 kHz. A very high Doppler frequency of

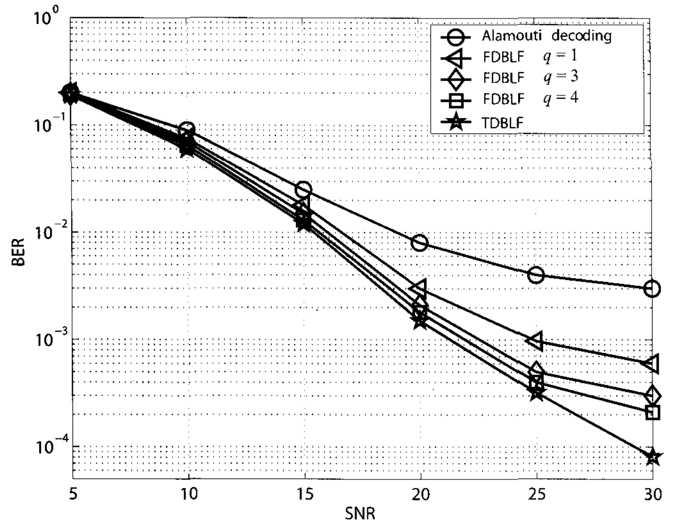

Fig. 4. BER performance of the Alamouti decoding, the previous TDBLF, and the proposed FDBLF when ideal CSI is assumed to be known. Two transmit antennas and one receive antenna OFDM system.

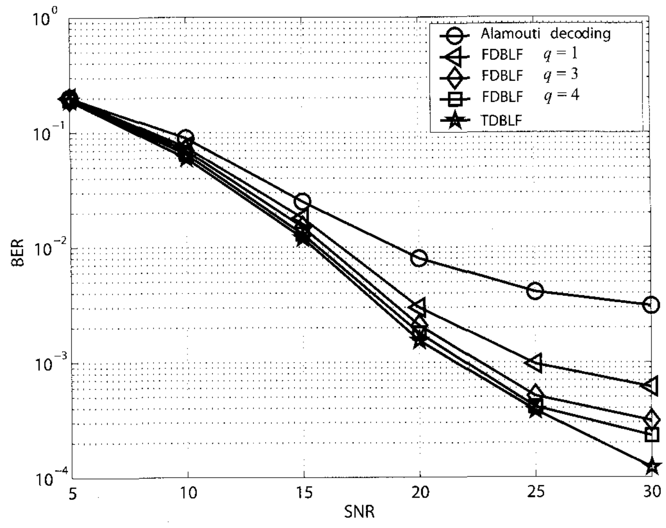

Fig. 5. BER performance of the Alamouti decoding, the previous TDBLF, and the proposed FDBLF when CSI is estimated. Two transmit antennas and one receive antenna OFDM system.

297 Hz is considered, which leads to $f_D(N+\nu)T_s = 0.1188$. Constellation for the symbol mapping is a 16-QAM. We used the reduced typical urban (TU) power delay profile for simulation [18]. We used the Jake's model to generate the time varying channel gains [19]. Since different signal detection algorithms may react differently to CSI estimation error, we consider both cases when exact CSI is available and when CSI is estimated. For CSI estimation, the pilot tone placement and interpolation method in [13] are adopted.

Fig. 4 shows the bit error rate (BER) versus SNR performance of the simple Alamouti decoding, the previous TDBLF, and the proposed FDBLF when ideal CSI is assumed to be known. As can be observed in Fig. 4, the Alamouti decoding shows the worst performance and the TDBLF achieves the best performance. The proposed FDBLF with different q ($q = 1, 3, 4$) falls between the two schemes. As the parameter q increases from 1 to 4, the performance of the FDBLF gets closer to that of the TDBLF. The small performance gap between the time domain and the frequency domain filter approaches seems to be due to the block diagonal approximation in the frequency domain approach.

Fig. 5 shows the BER performance of the three schemes when CSI is estimated. Again, the Alamouti decoding shows the worst performance and the time domain filtering shows the best performance. Unlike in the ideal CSI case, however, the proposed frequency domain filter with $q = 4$ shows almost identical performance with the time domain filter for SNR range from 5 up to 25 dB. This is because the time domain filter approach is more sensitive to CSI estimation error than the proposed frequency domain filter approach. As shown in Fig. 5, the proposed FDBLF shows an error floor at a very high SNR. The error floor is caused by the interference in (27) and (31). As the SNR increases with a fixed signal power, the noise variance decreases, however, the interference does not decrease, causing an error floor. When two receive antennas are used as in Fig. 6, however, the error floor problem seems to be resolved, thanks to the increased diversity order.

Fig. 6 is the BER performance of the three schemes when CSI is estimated. Two transmit antennas and two receive antennas OFDM system with maximal ratio receive combining (MRRC) is considered. Again, the performance of the proposed frequency domain filter falls between the previous TDBLF and the Alamouti decoding. The proposed scheme with $q = 4$ shows almost identical performance as the previous time domain filter over the entire considered SNR range. However, the performance gap of the proposed schemes with different q's is quite different from the case of one receive antenna. The proposed scheme with $q = 1$ shows a very small performance degradation relative to $q = 4$ case. It seems that the increased diversity gain employing two receive antennas is a dominant factor in the performance improvement over one receive antenna system. Therefore, a smaller q for the proposed frequency domain filter is desirable when two receive antennas are used than when one receive antenna is used.

Comparing Figs. 5 and 6, we can observe that the simple Alamouti decoding improves the error performance significantly by employing one more receive antenna at the receiver even in the fast fading channels. The time domain filtering approach requires the SNR of 21.5 dB to achieve BER of 10^{-3} when one receive antenna is used, while the Alamouti decoding requires SNR of only 13.5 dB to achieve BER of 10^{-3} when two receive antennas are used. Therefore, depending on the target BER, the cost of additional receive antenna, and the computational complexity requirement, the simple Alamouti decoding may be a preferred choice even in quite fast fading channels.

Finally, the complexity cost ratio is evaluated using parameters adopted in the simulations. Table 2 shows the complexity ratio of the two schemes for various q and a fixed N. As the parameter q increases, the ratio decreases. Even when the parameter $q = 4$ (the maximum value adopted in the simulations), the proposed FDBLF complexity is approximately $1/31$ of TDBLF. Therefore, it can be concluded that the proposed FDBLF is

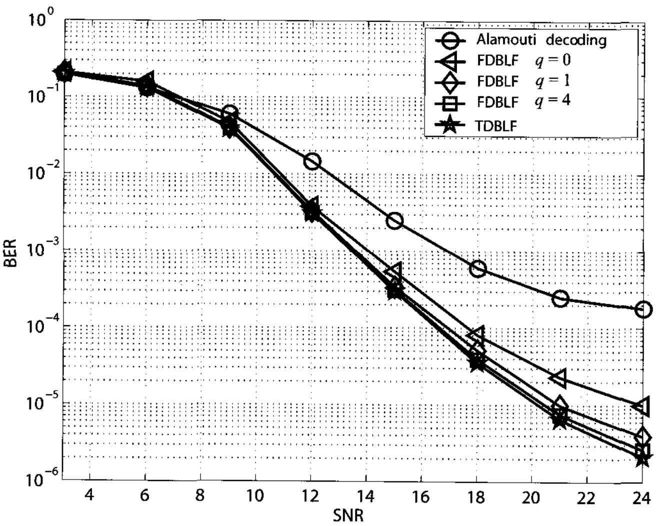

Fig. 6. BER performance of the Alamouti decoding, the previous TDBLF, and the proposed FDBLF when CSI is estimated. Two transmit antennas and two receive antennas OFDM system with maximal ratio receive combining (MRRC).

Table 2. Complexity ratio of the FDBLF to the TDBLF for various q.

$N = 128$	$q = 1$	$q = 2$	$q = 3$	$q = 4$
$R(q, N)$	1/751	1/172	1/64	1/31

computationally much more efficient at the cost of a small performance degradation over a wide SNR range, compared with the previous TDBLF.

VII. CONCLUSION

This paper proposes a frequency domain block-linear filter to mitigate the time-varying channel effects in Alamouti coded MIMO OFDM systems with low computational complexity. The proposed approach exploits the sparse structure of the system matrix in the frequency domain. It was shown that the proposed frequency domain filtering is computationally much more efficient than the previous time domain filtering. Simulation results showed that although performance gap between the two schemes exist when ideal CSI is assumed to be known and when SNR is high, the performance gap becomes smaller when the CSI is estimated, which implies the relative robustness of the proposed scheme to CSI estimation error. Thus frequency domain block-linear filtering should be preferred over the time domain approach in most practical cases.

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ACKNOWLEDGEMENT

This work was supported in part by Yonsei Univ. Research Fund of 2005.

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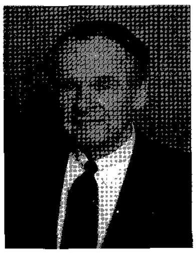

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