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Research Paper

ICI MITIGATION IN SFBC-OFDM SYSTEMS

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Use of Space-Frequency Block Coded (SFBC) OFDM signals is advantageous in high-mobility broadband wireless access, where the channel is highly time—as well as frequency-selective because of which the receiver experiences both Inter-Symbol Interference (ISI) as well as Inter-Carrier Interference (ICI). ISI occurs due to frequency and/or time selectivity of the channel caused due to the violation of the quasi-static fading. In addition, ICI which results in loss of orthogonality among the subcarriers is due to time selectivity of the channel. In this paper, we are concerned with the detection of SFBC-OFDM signals on time- and frequency-selective MIMO channels. Specifically, we investigate the interference cancellation techniques and evaluate the performance of an receiver for SFBC-OFDM to mitigate the effects of ICI induced by the time selectivity of the channel.

Keywords: SFBC, OFDM signals, ICI migigation

INTRODUCTION

The space time coding and OFDM advantages are combined which makes the wireless systems designs more attractive. This involves Space Frequency Coding (SFC), which is coding across space and frequency. The space frequency coding can be done by taking the space-time codes (e.g., Alamouti code 1998), and apply them in the frequency dimension instead of time dimension (Bolcskei and Paulraj, 2000). That is, the space-time coded symbols are mounted on multiple OFDM subcarriers instead of mounting on multiple time slots. SFC-OFDM is more attractive due to the use of orthogonal space-time block codes (OSTBCs) in the frequency dimension and because of their low complexity decoding (i.e., single symbol decodability) and suitability for fast fading channels (Yang, 2005). For high mobility broadband wireless access, Space-Frequency Block Coded (SFBC) OFDM scheme which uses Alamouti code in the frequency dimension is defined.

The ‘Quasi-Static’ (QS) assumption (i.e., fade remains constant over one block time, which is valid only in slow fading channels) is

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essential for the time dimension OSTBCs to be single symbol decodable. The violation of this QS assumption results in an error-floor. Such a violation is due to rapid time-variations in the fading process. On the other hand, even if time variations is very slow in fading channels the QS assumption of SFBC-OFDM gets violated in frequency dimension in highly frequency selective channels (i.e., different subcarriers, and hence symbols belonging to the same SFBC block mounted on different subcarriers, see different channel gains). The factors effecting this are channel length L, power delay profile of the channel, and the SFBC block size. This QS assumption violation becomes a source of significant Inter-Symbol Interference (ISI) in the frequency dimension in SFBC OFDM in highly frequency-selective channels (i.e., large L). If left uncared for ISI, results in error floors [5]. Further, in any OFDM system, the loss of orthogonality among subcarriers is due to the channel changes within an OFDM symbol duration, which results in Inter-Carrier Interference (ICI) (Robertson and Kaiser, 1999). Thus, in addition to the issue of ISI caused due to frequency selectivity of the channel, SFBC OFDM experiences ICI caused due to time-selectivity of the channel (i.e., channel varying within one OFDM symbol duration) (Stamoulis *et al.*, 2002). Like ISI, ICI, if uncared for, also will result in error floors.

Attempts have been made in the literature to cancel ICI in MIMO OFDM systems. Stamoulis *et al.*, in (2002), proposed ICI-mitigating block linear filters for STBCOFDM. However, in large delay spreads they do not consider the violation of QS assumption. If both ISI (due to violation of QS assumption) as well as ICI (due to time-selectivity) can be estimated and cancelled, then only there is a possibility of improvement in its performance. In this paper, we propose a successive interference cancellation approach to mitigate the effects of ICI in SFBC-OFDM. In addition to ICI, ISI also gets cancelled The proposed detector estimates (using soft output values) and cancels the ISI in the first step, and then estimates and cancels the ICI in the second step. This two step procedure is carried out in multiple cancellation stages. We evaluate the performance of the proposed detector for different codes including i) rate-1 Alamouti code (1998), and ii) rate-2/3 G5 code, for varying degrees of time selectivity (different speeds) and frequency selectivity (different channel lengths, L). We show that the proposed detector effectively cancels the ICI alongwith ISI in high mobility, large delay spread channels.

SYSTEM MODEL

We consider a MIMO OFDM system with N_c subcarriers, N_t transmit antennas, and N_r receive antennas. Let $X_k^{(i)}$ denote the complex data symbol transmitted on the k^{th} subcarrier of an OFDM symbol from the i^{th} transmit antenna. That is, the symbols $\{X^{(i)}, k = 1, \cdots, N_c, i = 1, \cdots, N_t\}$ are transmitted in parallel on N_c subcarriers by N_t transmit antennas. After IDFT processing and insertion of guard interval of ng samples at the transmitter, the discrete-time sequence at the i^{th} transmit antenna is given by

$$X_n^{(i)} = \frac{1}{N_c} \sum_{k=0}^{N_c-1} X_{k+1}^{(i)} e^{j\frac{2®nk}{N_c}} \quad -n_g \leq n \leq N_c - 1 \qquad ...(1)$$

n_g is assumed to be longer than the maximum channel delay spread, L. Assuming perfect carrier synchronizaynchronization, timing, and sampling at the receiver, the discrete-time received sequence at the j^{th} receive antenna can be written as:

$$y_n^{(j)} = \sum_{i=1}^{N_t} \sum_{\substack{l=0 \\ (i:j)}}^{L-1} h^{(i,j)}(n;l) x_{(n-l)}^{i} + w_n^{(j)},\ j = 1, \ldots, N_r;$$

$$-n_g \le n \le N_c - 1 \qquad ...(2)$$

where $h^{(i;j)}(n;l)$ represents the discrete-time, time varying (i.e., time selective) L-length (i.e., frequency selective) channel impulse response between the ith transmit and j^{th} receive antennas, and $w(j)$ is the additive noise on the jth receive antenna, assumed to be complex Gaussian with zero mean and variance N_o.

After guard interval removal and DFT operation, the received signal on the k^{th} subcarrier on the j^{th} receive antenna, $Y_k^{(j)}$, can be written as:

$$Y_k^{(j)} = \sum_{i=1}^{N_t} G_{k,k}^{(i,j)} X_k^{(i)} + \underbrace{\sum_{i=1}^{N_t} \sum_{\substack{m \ne k \\ m=0}}^{N_c-1} G_{k,m}^{(i,j)} X_m^{(i)}}_{ICI} + \underbrace{W_k^{(j)}}_{Noise} \qquad ...(3)$$

where the coefficients $G_{k,m}^{(i,j)}$ are given by,

$$G_{k,m}^{(i,j)} = \frac{1}{N_c} \sum_{n=0}^{N_c-1} \sum_{l=0}^{L-1} h^{(i,j)}(n;l) e^{\frac{j2®n(m-k)}{N_c}} e^{\frac{-j2®ml}{N_c}} \qquad ...(4)$$

Note that $G^{(i,j)}$, $k = m$ denotes the amount of carrier leak from mth subcarrier to the k^{th} subcarrier, which essentially contributes to the ICI term in (3). It is easy to see that if the channel is not time-selective (i.e., time-flat) between all transmit/receive antenna pairs, i.e., if

$$h^{(i,j)}(n_1;l) = h^{(i,j)}(n_2;l)\ \forall i,h,l\ -n_g \le n_1, n_2 \le N_c - 1 \qquad ...(5)$$

then $G_{k,m}^{(i,j)} = 0$, for $k = m$, and (3) reduces to (resulting in no ICI term)

$$Y_k^{(j)} = \sum_{i=1}^{N_t} \tilde{G}_{k,k}^{(i,j)} X_k^{(i)} + W_k^{(j)} \qquad ...(6)$$

where $\tilde{G}_{k,k}^{(i,j)}$ is $G_{(k,k)}^{(i,j)}$ for time-flat channels, given by for time-at channels, given by,

$$\tilde{G}_{k,k}^{(i,j)} = \frac{1}{N_c} \sum_{l=0}^{L-1} h^{(i,j)}(n_1,l) e^{-j\frac{2®kl}{N_c}} \qquad ...(7)$$

Space-Frequency Block Coded OFDM

Here, by adopting the above system model, the case of space-frequency coded symbols transmission on different transmit antecnnas (space) and different subcarriers (frequency) is carried on. That is, $X_k^{(i)}$s are the symbols obtained from the space frequency coding scheme. Specifically, we consider the use of OSTBCs as the space-frequency codes. Let K denote the length of one space-frequency code block. We divide N_c subcarriers into N_g groups each having K subcarriers so that $N_c = N_g K$ + © where each group is called as one SFC block. For example, K = 2 for Alamouti code. If N_c is not a multiple of K, then there will not be any transmission on © subcarriers, or, alternatively, these © subcarriers can be used for pilot transmission.

The SFBC OFDM frame thus obtained can be written as an $N_c \times N_t$ matrix

$$X = \begin{bmatrix} X_1^{(1)} & \ldots & X_1^{(N_t)} \\ X_2^{(1)} & \ldots & X_2^{(N_t)} \\ \vdots & \vdots & \vdots \\ X_{N_c}^{(1)} & \ldots & X_{N_c}^{(N_t)} \end{bmatrix} = \begin{bmatrix} X(1) \\ \vdots \\ X(N_g) \\ 0^{k \times N_t} \end{bmatrix} \qquad ...(8)$$

where $X(q)$ is a $K \times N_t$ matrix, $q = 1, 2, \cdots, N_g$. $X(q)$ is from a OSTBC in P complex information symbols $[v1(q), v2(q), \cdots, vP(q)]$ and rate-P/K. The i^{th} column of X is transmitted on the i^{th} transmit antenna after IDFT processing. For example, for SFBC OFDM with $N_c = 4$ using Alamouti code, $N_t = 2$ and the matrix X can be written as

$$X_{Alamouti} = \begin{bmatrix} v_1(1) & v_2(1) \\ -v_2(1)^* & v_1(1)^* \\ v_1(2) & v_2(2) \\ -v_2(2)^* & v_1(2)^* \end{bmatrix} \quad ...(9)$$

By stacking all the K rows of $X(q)$ into one $KN_t \times 1$ vector, $\bar{x}(q)$. This $\bar{x}(q)$ vector can be written as,

$$\bar{x}(q) = Av(q) \quad ...(10)$$

where $V(q)$ is a $2P \times 1$ vector, given by

$$v(q) = [v_{1I}(q), \ldots, v_{PI}(q), v_{1Q}(q), \ldots, v_{PQ}(q)] \quad ...(11)$$

where $vpI(q)$ and $vpQ(q)$, respectively, are the real and imaginary parts of the pth complex information symbol in the qth group, $p = 1, 2, ..., P$, $q = 1, 2, ..., N_g$. The matrix A in (6.10) is a $KN_t \times 2P$ complex matrix which performs the coding on $v(q)$. For the SFBC OFDM with Alamouti code, A is given by,

$$A_{Alamouti} = \begin{bmatrix} 1 & 0 & j & 0 \\ 0 & 1 & 0 & j \\ 0 & -1 & 0 & -j \\ 1 & 0 & -j & 0 \end{bmatrix} \quad ...(12)$$

Using (12) in (10), we get

$$\bar{x}_{Alamouti}(q) = [v_1(q), v_2(q), -v_2(q)^*, v_1(q)^*]^T \quad ...(13)$$

And

$$X_{Alamouti}(q) = \begin{bmatrix} v_1(q) & v_2(q) \\ -v_2(q)^* & v_1(q)^* \end{bmatrix} \quad ...(14)$$

Received Signal Model for SFBC OFDM

At the receiver, the DFT outputs, $Y_k^{(j)}$s, of (3) are stacked to form a K Nr × 1 vector for each group, as,

$$y(q) = \left[Y_{r+1}^{(1)}, \ldots, Y_{r+1}^{(N_r)}, \ldots, Y_{r+k}^{(1)}, \ldots, Y_{r+k}^{(N_r)}\right]^T \quad ...(15)$$

where $r = (q - 1)K$. Now, $Y(q)$ in (15) can be written in the form,

$$y(q) = H(q)\bar{x}(q) + s(q) + w(q) \quad ...(16)$$

where $H(q)$ is a KN_r X KN_t block diagonal matrix, given by

$$\bar{H}(q) = \begin{bmatrix} H(q,1) & \ldots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \ldots & H(q,k) \end{bmatrix} \quad ...(17)$$

where $H(q, s)$ is the $N_r \times N_t$ channel matrix, $q = 1, 2, \cdots, N_g$, $s = 1, 2, \cdots, K$, given by

$$H(q,s) = \begin{bmatrix} G_{u,u}^{(1,1)} & \ldots & G_{u,u}^{(1,N_t)} \\ \vdots & \ddots & \vdots \\ G_{u,u}^{(N_r,1)} & \ldots & G_{u,u}^{(N_r,N_t)} \end{bmatrix} \quad ...(18)$$

where $u = (q - 1)K + s$. The noise vector $w(q)$ in (16) is given by,

$$w(q) = \left[W_{r+1}^{(1)}, \ldots, W_{r+1}^{(N_r)}, \ldots, W_{r+K}^{(1)}, \ldots, W_{r+K}^{(N_r)}\right]^T \quad ...(19)$$

where $s = (q - 1)K$. The interference vector $I(q)$ in (16) can be written in the form,

$$S(q) = \bar{R}(q)\bar{x}(q) + \sum_{b=1,\, b\neq q}^{N_g} \bar{Q}(q,b)\bar{x}(b) \quad ...(20)$$

where the block matrices $R(q)$ and $Q(q, b)$ are given by,

$$\overline{R}(q)=\begin{bmatrix} 0 & R_{q'+1,\,q'+2} & \cdots & R_{q'+1,\,q'+K} \\ R_{q'+2,\,q'+1} & 0 & \cdots & R_{q'+2,\,q'+K} \\ \vdots & \vdots & \ddots & \vdots \\ R_{q'+K,\,q'+1} & R_{q'+K,\,q'+2} & \cdots & 0 \end{bmatrix} \quad ...(21)$$

$$\overline{Q}(q,b)=\begin{bmatrix} R_{q'+1,\,b'+1} & \cdots & R_{q'+1,\,b'+K} \\ \vdots & \ddots & \vdots \\ R_{q'+K,\,b'+1} & \cdots & R_{q'+K,\,b'+K} \end{bmatrix} \quad ...(22)$$

where $q'=(q-1)K$, $b'=(b-1)K$, and

$$R_{x,y}=\begin{bmatrix} G^{(1,1)}_{x+1,y+1} & \cdots & G^{(1,N_t)}_{x+1,y+K} \\ \vdots & \ddots & \vdots \\ G^{(N_r,1)}_{x+K,y+1} & \cdots & G^{(N_r,N_t)}_{x+K,y+K} \end{bmatrix} \quad ...(23)$$

Detection of SFBC-OFDM

For the case of time-flat and frequency-flat conditions (i.e., no ISI and no ICI), the detector for OSTBCs presented in for conventional space time codes can be used for detecting SFBC-OFDM as follows.

For time flat case, can be written as

$$y(q)=H_{eq}(q)\,v(q)+w(q) \quad ...(24)$$

where $H_{eq}(q)$ is a $KN_r \times 2P$ equivalent channel matrix,is given by

$$H_{eq}(q)=\overline{H}(q)A \quad ...(25)$$

For the frequency-flat case, the quasi-static assumption holds, i.e., in (17),

$$H(q,1)=H(q,2)=\ldots=H(q,K),\forall q \quad ...(26)$$

Now, from the optimal detector for SFBC OFDM under the above conditions can be shown to be of the form,

$$\hat{y}(q)=\Re\left(H^{*}_{eq}(q)y(q)\right) \quad ...(27)$$

where $\hat{y}(q)$ is a $2K \times 1$ vector containing the estimates of the real and imaginary parts of the complex information symbols in a stacked up fashion, which can be shown to be

$$\hat{y}(q)=\Re\left[H^{*}_{eq}(q)H_{eq}(q)\right]V(q)+\Re\left[H^{*}_{eq}(q)w(q)\right]$$
$$=\Lambda(q)V(q)+\hat{w}(q) \quad ...(28)$$

where $\Lambda(q)=\Re\left[H^{*}_{eq}(q)H_{eq}(q)\right]$ is a diagonal matrix, and hence there will not be any inter-symbol interference. Furthermore, it can also be shown that under these conditions $\hat{w}(q)=\Re\left[H^{*}_{eq}(q)w(q)\right]$ is white Gaussian. Hence, the Euclidean distance based symbol-by-symbol detection on $\hat{y}(q)$ is optimal.

On the other hand, for the case when the quasi-static assumption is violated (in this case due to frequencyselectivity of the channel), then $\Lambda(q)$ is not diagonal. Hence, the detector in (27) results in an error floor. The optimum detector for this system would be a Maximum Likelihood (ML) detector in *P* variables which results in exponential receiver complexity. A time-selective channel in this case results in inter-carrier interference (6.16). We illustrate the effect of interferences due to time- and frequency-selectivity of the channel, where we plot the output SIR of SFBC OFDM with Alamouti code (N_t = 2), as a function of user velocity and channel delay spread (L equal-power Rayleigh fading paths) for N_c = 128 subcarriers, Δ_f = 0.5 KHz subcarrier spacing, f_c = 2.5 GHz carrier frequency, N_r = 1 receive antenna, and no noise. The SIR degrades for increasing user velocity and channel delay spread even when the user is static (i.e., velocity = 0 Km/h and hence time-flat fading), the SIR degrades significantly for increasing *L* (e.g., about 30 dB of SIR degradation from *L* = 2 to *L* = 16). Also, for a given L, increasing velocity degrades SIR (e.g., about 8 dB degradation from 0 to 60 Km/h for L = 8). We further observe that cancellation techniques can be employed to recover the performance loss due to time- and frequency-selectivity induced interferences, which is our focus on the following section.

PROPOSED IC RECEIVER FOR SFBCOFDM

In this section, we propose a novel two-step PIC detector that cancels ISI and ICI in SFBC OFDM. The proposed detector estimates and cancels the ISI (caused due to the violation of the quasi-static assumption) in the first step, and then estimates and cancels the ICI (caused due to loss of subcarrier orthogonality) in the second step. This two step procedure is then carried out in multiple stages. The proposed detector is presented in the following.

We consider perfect channel knowledge at the receiver. So, in the notation, we will not differentiate between the actual channel and the channel estimate available at the receiver. The detector, however, can work with imperfect channel estimates.

First, we model the ISI caused by the violation of the quasi-static assumption. To do that, we split the block diagonal matrix $\overline{H}(q)$ in (16) into two parts; i) a quasi-static part $\overline{H}_{qs}(q)$ and ii) a non-quasi-static part $\overline{H}_{nqs}(q)$, such that

$$\overline{H}(q) = \overline{H}_{qs}(q) + \overline{H}_{nqs}(q) \quad ...(29)$$

where,

$$\overline{H}_{qs}(q) = \begin{bmatrix} H(q,1) & 0 & \dots & 0 \\ 0 & H(q,1) & \dots & 0 \\ \vdots & \vdots & \ddots & \dots \\ 0 & 0 & \dots & H(q,1) \end{bmatrix} \quad ...(30)$$

And

$$\overline{H}_{nqs}(q) = \begin{bmatrix} 0 & 0 & \dots & 0 \\ 0 & \Delta H(q,2) & \dots & 0 \\ \vdots & \vdots & \ddots & \dots \\ 0 & 0 & \dots & \Delta H(q,K) \end{bmatrix} \quad ...(31)$$

where $\Delta H(q, m) = H(q, m) - H(q, 1)$

Similarly we can split the channel equivalent matrix $H_{eq}(q)$, as

$$H_{eq}(q) = H_{eq-qs}(q) + H_{eq-nqs}(q) \quad ...(32)$$

where,

$$H_{eq-qs}(q) = \overline{H}_{qs}(q)A, \text{ and } H_{eq-nqs}(q) = \overline{H}_{nqs}(q)A$$

Based on the above above formulations and (20), we can write (16) as,

$$y(q) = \overline{H}_{qs}(q)\overline{x}(q) + \underbrace{\overline{H}_{nqs}(q)\overline{x}(q)}_{QS\ violation}$$

$$+ \underbrace{\overline{R}(q)\overline{x}(q) + \sum_{b=1,\, b\neq q}^{N_g} \overline{Q}(q,b)\overline{x}(b) + w(q)}_{Loss\ of\ orthogonality} \quad ...(33)$$

As in,(27), an estimate of $y(q)$ can be obtained as

$$\hat{y}(q) = \Re\left(H^*_{eq-qs}(q)y(q)\right)$$

$$= \underbrace{\Re\left[H^*_{eq-qs}(q)H_{eq-qs}(q)\right]v(q)}_{Desired\ signal} + \underbrace{\Re\left[H^*_{eq-qs}(q)H_{eq-nqs}(q)\right]v(q)}_{ISI}$$

$$+ \underbrace{\Re\left[H^*_{eq-qs}(q)\overline{R}(q)\overline{X}(q)\right] + \Re\left[H^*_{eq-qs}(q)\sum_{b=1,\, b\neq q}^{N_g} \overline{Q}(q,b)\overline{x}(b)\right]}_{ICI}$$

$$+ \underbrace{\Re\left[H^*_{eq-qs}(q)w(q)\right]}_{Noise} \quad ...(34)$$

As can be seen, (34) the estimate $\hat{y}(q)$ contains the desired signal, ISI, ICI, and noise components. Based on this received signal model in (6.34) and the knowledge of the matrices $H_{eq-qs}(q), H_{eq-nqs}(q), \overline{R}(q), \overline{Q}(q,b) \forall q,b$ we formulate the proposed interference estimation and cancellation procedure as follows

- For each SF code block q, estimate the information symbols $\hat{v}(q)$ from (34), ignoring ISI and ICI.
- For each SF code block q, obtain an estimate of the ISI (i.e., an estimate of the 2nd term in (6.34)) from the estimated symbols $\hat{v}(q)$ in the previous step.
- Cancel the estimated ISI from $\hat{y}(q)$.
- Using $\hat{v}(q)$ from step 1, regenerate $\hat{\hat{x}}(q)$ using (10). Then, using $\hat{\hat{x}}(q)$, obtain an estimate of the ICI (i.e., an estimate of the sum of 3rd and 4th terms in (34)).
- Cancel the estimated ICI from the ISI cancelled output in step 3.
- Take the ISI and ICI cancelled output from step 5 as the input, back to step 1 (for the next stage of cancellation).

Based on the above, and defining

$\Lambda(q) = \Re\left[H^*_{eq-qs}(q) H_{eq-qs}(q)\right]$, the cancellation algorithm for the mth stage can be summarized as follows.

Initialization: Set $m = 1$.

Evaluate

$$\hat{y}^{(m)}(q) = \Re\left(H^*_{eq-qs} y(q)\right), 1 \le q \le N_g \quad \text{...(35)}$$

Loop

Estimate

$$\hat{v}^{(m)}(q) = \hat{y}^{(m)}(q) \Lambda^{-1}(q), 1 \le q \le N_g \quad \text{...(36)}$$

Cancel ISI

$$\hat{y}^{(m+1)}(q) = \hat{y}^{(1)}(q) - \Re\left(H^*_{eq-qs}(q) H_{eq-nqs}(q)\right) \hat{v}^{(m)}(q),$$

$$1 \le q \le N_g \quad \text{...(37)}$$

Form $\hat{\hat{x}}^{(m)}(q)$ from,

$$\hat{\hat{x}}^{(m)}(q) = A \hat{v}^{(m)}(q), 1 \le q \le N_g \quad \text{...(38)}$$

Cancel ICI

$$\hat{y}^{(m+1)}(q) = \hat{y}^{(m+1)}(q) - \Re\left(H^*_{eq-qs}(q) \overline{R}(q) \hat{\hat{x}}^{(m)}(q)\right)$$

$$- \Re\left(H^*_{eq-qs}(q) \sum_{b=1, b \ne q}^{N_g} \overline{Q}(q,b) \hat{\hat{x}}^{(m)}(b)\right)$$

$$1 \le q \le N_g \quad \text{...(39)}$$

$m = m + 1$

goto Loop

It is noted that the above cancellation algorithm has polynomial complexity. Also, since $\Lambda(q) = \Re\left[H^*_{eq-qs} H_{eq-qs}\right]$ is a diagonal matrix, its inversion is simple. In practice, accurate estimation of the channel coefficients is essential, which can be achieved, for example, using the algorithm proposed in Stamoulis *et al.* (2002).

SIMULATION RESULTS

We evaluated the BER performance of the proposed PIC detector for SFBC OFDM in frequency and time selective Rayleigh fading through simulations. In all our performance results in this section, we have used a 5.4 GHz SFBC OFDM system having N_c = 64 subcarriers with a subcarrier spacing of 312.5 KHz, and N_r = 1 one receive antenna. A frequency selective tapped-delay line channel model with $L = 2$ and $L = 4$ equal power paths is used. The time-variations of the Rayleigh fading process on each path is simulated using Jakes model for different Doppler bandwidths (i.e., different mobile speeds). The codes considered include i) rate-1 Alamouti code (G_2) (Alamouti, 1998), and ii) rate-2/3, 5 transmit antenna G_5 code.

In Figure 1, we plot the BER performance of the proposed PIC detector in timeflat, frequency selective Rayleigh fading with $L = 2$ and $L = 4$ equal power paths for SFBC OFDM using Alamouti code and QPSK. So there is no time selectivity induced ICI here. However, there is frequency selectivity induced ISI. In Figure 1, Stage-1 performance corresponds to the case of no cancellation, whereas Stage-3 performance is after two stages of proposed cancellation. For comparison purposes, we have also plotted the performance for time-flat and frequency-flat fading (i.e., no ISI and no ICI), which provides the best possible performance. From Figure 1, it can be observed that for $L = 2$ (i.e., channel delay spread is small), the ISI induced is small, and hence there is no major performance improvement due to the cancellation. However, for $L = 4$ (i.e., delay spread of the channel is large), the ISI induced is high, and, in this case, the proposed cancellation results in significant performance gain (e.g., about 2 dB gain at 2×10^{-2} BER).

In Figure 2, we plot the BER performance of SFBC OFDM on time-selective and frequencyselective fading for rate-2/3 G_5 OSTBC (5 Tx antennas) using 16-QAM. The mobile speed is 50 Km/h and $L = 2$. The BER plot for the case of timeflat and frequency-flat fading (i.e., the case of no ISI and no ICI) is also plotted for comparison. From Figure 2, it can be seen that due to ISI and ICI the performance without cancellation (i.e., Stage-1) is severely affected compared to the case of time-flat and frequency-flat fading. However, the performance is significantly improved by the proposed PIC detector (Stage-2 and Stage-3) because of the effective mitigation

Figure 1: BER Performance of the Proposed PIC Detector in Time Flat Frequency Selective Rayleigh Fading Channel (QPSK)

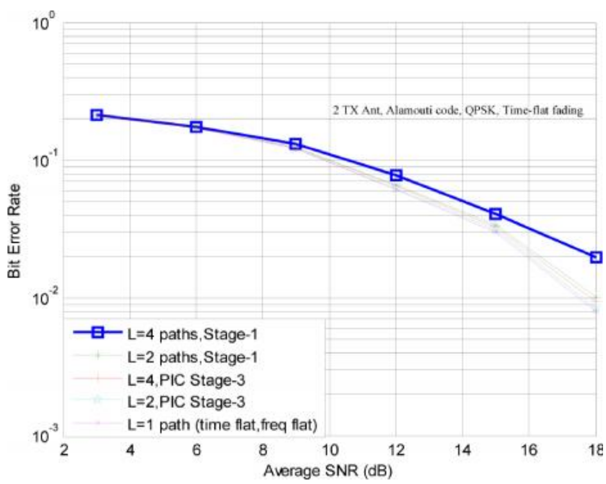
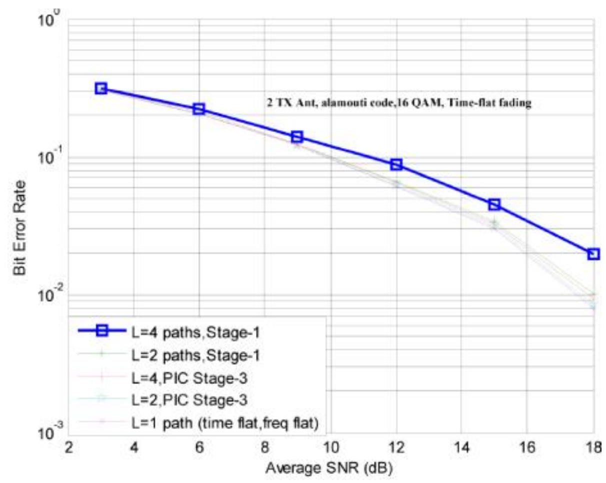
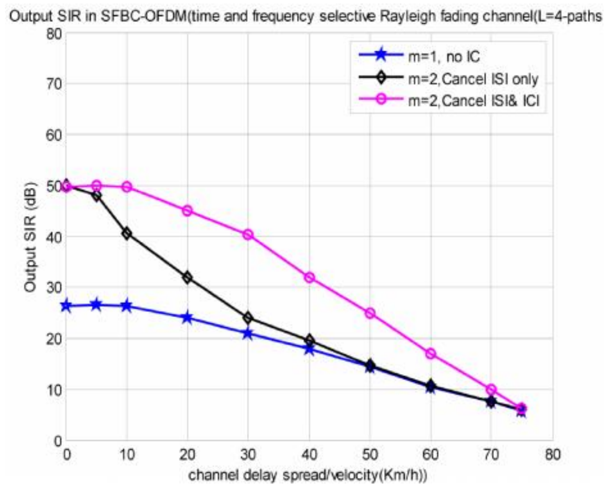


Figure 2: BER Performance of Proposed PIC Detector in Time Flat Frequency Selective Rayleigh Fading Channel (16 QAM)



of ISI and ICI. For example, at a BER of 5 × 10^{-2}, the proposed canceller results in about 6 dB improvement in performance compared to the case of no cancellation at the receiver.

Figure 3: Output SIR Performance with the Proposed IC Detector



In Figure 3 we plot the output SIR performance with the proposed detector with time and frequency selective fading with $L = 4$.

REFERENCES

1. Alamouti S M (1998), "A Simple Transmit Diversity Technique for Wireless Communications", *IEEE Jl. Sel. Areas Commun.*, Vol. 16, October, pp. 1451-1458.
2. Bolcskei H and Paulraj A (2000), "Space-Frequency Coded Broadband OFDM Systems", Proc. IEEE WCNC'2000, September, pp. 1-6.
3. Gong Y and Letaief K B (2003), "An Efficient Space Frequency Coded Wideband OFDM System for Wireless Communications", *IEEE Trans. on Commun.*, Vol. 51, No. 12, pp. 2019-2029.
4. Lee K and Williams D (2000), "A Space-Frequency Transmitter Diversity Technique for OFDM Systems", Proc. IEEE GLOBECOM'2000.
5. Liu Z, Xin Y and Giannakis G B (2002), "Spacetime-Frequency Coded OFDM Over Frequency Selective Fading Channels", *IEEE Trans. on Signal Processing*, Vol. 50, No. 10, pp. 2465-2476.
6. Robertson P and Kaiser S (1999), "Analysis of the Loss of Orthogonality Through Doppler Spread in OFDM Systems", Proc. IEEE GLOBECOM'99, pp. 701-706.
7. Shao L and Roy S (2005), "Rate-One Space Frequency Block Codes with Maximum Diversity for MIMOOFDM", *IEEE Trans. on Wireless Commun.*, Vol. 4, No. 4, pp. 1674-1687.
8. Stamoulis A, Diggavi S N and Al-Dhahir N (2002), "Inter-Carrier Interference in MIMO-OFDM", *IEEE Trans. on Signal Processing*, Vol. 50, No. 10, pp. 2451-2464.
9. Yang H (2005), "A Road to Future Broadband Wireless Access: MIMO-OFDM-Based Air Interface", *IEEE Commun. Mag.*, January, pp. 53-60.