ICI-ISI MITIGATION IN COOPERATIVE SFBC-OFDM WITH CARRIER FREQUENCY OFFSET

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ABSTRACT

Synchronization issues pose a big challenge in cooperative communications. The benefits of cooperative diversity could be easily undone by improper synchronization. The problem arises because it would be difficult, from a complexity perspective, for multiple transmitting nodes to synchronize to a single receiver. For OFDM based systems, loss of performance due to imperfect carrier synchronization is severe, since it results in inter-carrier interference (ICI). The use of space-time/space-frequency codes from orthogonal designs are attractive for cooperative encoding. But orthogonal designs suffer from inter-symbol interference (ISI) due to the violation of quasi-static assumption, which can arise due to frequency- or time-selectivity of the channel. In this paper, we are concerned with combating the effects of $i)$ ICI induced by carrier frequency offsets (CFO), and $ii)$ ISI induced by frequency selectivity of the channel, in a cooperative communication scheme using space-frequency block coded (SFBC) OFDM. Specifically, we present an iterative interference cancellation (IC) algorithm to combat the ISI and ICI effects. The proposed algorithm could be applied to any orthogonal or quasi-orthogonal designs in cooperative SFBC OFDM schemes.

I. INTRODUCTION

The problem of fading and the ways to combat it through spatial diversity techniques have been an active area of research. Multiple input multiple output (MIMO) techniques have become popular in realizing spatial diversity and high data rates through the use of multiple transmit antennas. Recent research has shown that the advantages of spatial diversity could be realized in single-antenna user nodes through user cooperation [1],[2]. One major issue in systems that employ user cooperation lies in achieving proper synchronization in time as well as frequency. For example, in OFDM based cooperative schemes, carrier frequency offsets (CFO) at the destination receiver can be destructive. This is because the orthogonality among various subcarriers gets lost due to CFO, which results in inter-carrier interference (ICI). From a complexity perspective, it is difficult for the multiple transmitting nodes to accurately synchronize to a single receiver. One solution to this problem could be to feed the time and frequency error values back to the transmitting nodes, so that the transmitting nodes can adjust their transmit timing/frequency accordingly. But this approach relies on extra signaling, and hence wastage of precious bandwidth. An alternate and interesting approach would be to use interference cancelling receivers at the destination receiver. We consider the latter approach in this paper.

The use of orthogonal designs [3] (OD) for cooperative communications has been studied in [4]. But ODs suffer from inter-symbol interference (ISI), when the quasi-static (QS) assumption is violated. In OFDM based systems that use ODs, such a violation can occur due to time- and/or frequency-selectivity of the channel, depending on whether space-time block coding (STBC) or space-frequency block coding (SFBC) is used. For example, in a SFBC based cooperative scheme, the QS assumption gets violated because of the frequency selectivity of the channel, which results in inter-symbol interference (ISI). In this paper, we are concerned with combating the ISI and ICI effects in SFBC-OFDM based cooperative schemes. Our specific contribution in this paper is a joint ISI-ICI cancellation algorithm which could be applied to any orthogonal design or quasi-orthogonal design. Our cancellation algorithm operates on soft output values, and hence can be easily applied to a coded system.

The rest of the paper is organized as follows. In Sec. II., we present the cooperative SFBC OFDM system model. In Sec. III., we present our cancellation algorithm. Results and discussions are presented in Sec. IV. Conclusions are given in Sec. V.

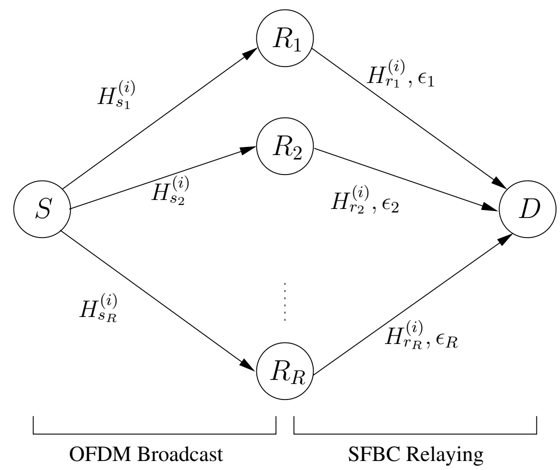

Figure 1: A wireless network consisting of one source, one destination and R relays.

II. SYSTEM MODEL

Consider a wireless network as depicted in Fig. 1 with $R+2$ nodes consisting of a source, a destination and R relays. All nodes are half duplex common nodes. i.e, a node can either transmit or receive at a time. All the nodes communicate using a N subcarrier OFDM scheme.

The transmission protocol is as follows :

- In the first time duration, the source transmits information symbols $X^{(i)}, 1 \leq i \leq N$ on all N subcarriers. All the R relays and the destination receive these N samples. This phase is called the broadcast phase.
- In the second phase, R relays[1] transmit a space-frequency encoded, scaled version of the received symbol. The destination receives this transmission. This phase is called SFBC relay phase.

It is assumed that the channels of all the links are frequency selective, and, without loss of generality we also assume that all the channels have L taps. We assume that all the channel state information is available at the respective receivers. The received signal, $Y_{r_j}^{(i)}$, on the ith carrier at the jth relay can be written as [2]

$$Y_{r_j}^{(i)} = H_{s_j}^{(i)} X^{(i)} + Z_j^{(i)}, \tag{1}$$

where $H_{s_j}^{(i)}$ is the frequency response on the ith carrier of the channel from source to jth relay, and $Z_{r_j}^{(i)}$ is additive white Gaussian noise with zero mean and variance σ^2.

A. Space-Frequency Block Coding at the Relay

We choose a orthogonal design of length K, P information symbols and R transmit antennas. We group the Nsubcarriers into N_g groups such that $N = N_g K + \kappa$. If N is not a multiple of K then there will not be any transmission on κ subcarriers, and accordingly the source will transmit only $N_g P$ information symbols. [3] We consider the Decode-and-Forward (DF) protocol. Let E be the total power per symbol consumed over the entire network. Then, it is shown in [5] that the optimum allocation that maximizes the receive SNR is when half the power is spent on the broadcast phase and the remaining half in the relay phase, i.e., $E\left[|X^{(i)}|^2\right] = E/2$. The received signal in (1) is decoded and scaled such that $E\left[|\widehat{X}_{r_j}^{(i)}|^2\right] = \frac{E}{2R}$, where $\widehat{X}_{r_j}^{(i)}$ is the scaled decoded symbol on the ith carrier at the jth relay. Now we form N_g groups out of the N entries of $\widehat{X}_{r_j}^{(i)}$ for all j. For each group q, we form the $2P \times 1$ vector $\widehat{\mathbf{V}}_j^{(q)}$ given by

$$\widehat{\mathbf{V}}_j^{(q)} = \left[\widehat{v}_{(j,1)I}^{(q)}, \cdots, \widehat{v}_{(j,P)I}^{(q)}, \widehat{v}_{(j,1)Q}^{(q)}, \cdots, \widehat{v}_{(j,P)Q}^{(q)}\right], \tag{2}$$

where $\widehat{v}_{(j,p)I}^{(q)}$ and $\widehat{v}_{(j,p)Q}^{(q)}$, respectively, are the real and imaginary parts of the pth complex information symbol in the qth group of the jth relay, $p = 1, 2, \cdots, P$, $q = 1, 2, \cdots, N_g$. That is, $\widehat{v}_{(j,p)I}^{(q)} = \Re\left[\widehat{X}_{r_j}^{((q-1)P+p)}\right]$ and $\widehat{v}_{(j,p)Q}^{(q)} = \Im\left[\widehat{X}_{r_j}^{((q-1)P+p)}\right]$. The space-frequency coded symbols for the qth group of the jth relay can be obtained as

$$\overline{\mathbf{X}}_{r_j}^{(q)} = \mathbf{A}_j \widehat{\mathbf{V}}_j^{(q)}, \tag{3}$$

where the $K \times 2P$ matrix $\mathbf{A}_j$ performs the generalized orthogonal space-frequency coding at the relay j. For example for the 2-relay case (i.e.,$R = 2$) using Alamouti code

$$\mathbf{A}_1 = \begin{bmatrix} 1 & 0 & \mathbf{j} & 0 \\ 0 & -1 & 0 & \mathbf{j} \end{bmatrix}, \mathbf{A}_2 = \begin{bmatrix} 0 & 1 & 0 & \mathbf{j} \\ 1 & 0 & -\mathbf{j} & 0 \end{bmatrix}. \tag{4}$$

The transmitted space-frequency coded signal from the jth relay can be written as

$$\overline{\mathbf{X}}_{r_j} = \begin{bmatrix} \overline{\mathbf{X}}_{r_j}^{(1)} \\ \vdots \\ \overline{\mathbf{X}}_{r_j}^{(N_g)} \\ \mathbf{0}^{\kappa \times 1} \end{bmatrix}. \tag{5}$$

Let $\overline{X}_j^{(i)}$ denote the ith element of $\overline{\mathbf{X}}_{r_j}$. Then the discrete-time transmit sequence at the jth relay is given by $x_j^{(n)} = IDFT_N(\overline{X}_{r_j}^{(i)})$.

B. Received Signal Model for the Relayed Signal

The received baseband signal at the destination after coarse carrier frequency synchronization and guard time removal is given by

$$y_d^{(n)} = \sum_{j=1}^{R} (x_j^{(n)} \star h_{r_j}^{(n)})\, e^{\frac{\mathbf{j}2\pi\epsilon_j n}{N}} + z_d^{(n)}, \quad 0 \leq n \leq N-1, \tag{6}$$

where $\star$ denotes linear convolution, $h_{r_j}^n$ is the channel impulse response from the jth relay. It is assumed that h_j^n is non-zero only for $n = 0, \cdots, L-1$, where L is the maximum channel delay spread. It is also assumed that the added cyclic prefix is greater than L. $\epsilon_j, j = 1, \cdots, R$ denotes residual carrier frequency offset (CFO) from the j'th relay normalized by the subcarrier spacing, and $z_d^{(n)}$ is the AWGN with zero mean and variance σ^2. We assume that all the nodes are time synchronized and that $\epsilon_j, j = 1, \cdots, R$ are known at the destination. At the destination, $y_d^{(n)}$ is first fed to the DFT block. The DFT output vector, $\mathbf{Y}_d$, can be written in the form

$$\mathbf{Y}_d = \sum_{j=1}^{R} \mathbf{C}_j \mathbf{H}_{r_j} \overline{\mathbf{X}}_{r_j} + \mathbf{Z}_d, \tag{7}$$

where $\mathbf{C}_j$ is a $N \times N$ circulant matrix given by

$$\mathbf{C}_j = \begin{bmatrix} C_j^{(0)} & C_j^{(1)} & \cdots & C_j^{(N-1)} \\ C_j^{(N-1)} & C_j^{(0)} & \cdots & C_j^{(N-2)} \\ \vdots & \vdots & \vdots & \vdots \\ C_j^{(1)} & C_j^{(2)} & \cdots & C_j^{(0)} \end{bmatrix}, \tag{8}$$

[1] We assume that all the relays participate in the cooperative transmission. It is also possible that some relays do not participate in the transmission based on whether the channel state is in outage or not.

[2] We use the following notation : Bold letter fonts are used for matrices and vectors. $\Re(.)$ denote real value of a complex argument and $\Im(.)$ denote imaginary value. $()^*$ denote matrix conjugation and transpose. $diag\{a_1, a_2, \cdots, a_N\}$ is a diagonal matrix having diagonal entries $a_1, a_2, \cdots, a_N$ and $\mathbf{j} = \sqrt{-1}$.

[3] Note that $N_g P \leq N$ since $P/K \leq 1$ for all orthogonal designs.

where $C_j^{(k)} = \text{DFT}_N\left(e^{\frac{\mathbf{j}2\pi n\epsilon_j}{N}}\right)$. $\mathbf{H}_{r_j}$ is the $N \times N$ diagonal channel matrix given by $\mathbf{H}_{r_j} = diag\left[H_{r_j}^{(1)}, H_{r_j}^{(2)}, \cdots, H_{r_j}^{(N)}\right]$, and the channel coefficient in frequency domain $H_{r_j}^{(i)}$ is given by $H_j^{(i)} = \text{DFT}_N\left(h_{r_j}^{(n)}\right)$. Similarly, $\mathbf{Z}_d = \left[Z_d^{(1)}, Z_d^{(2)}, \cdots, Z_d^{(N)}\right]$, where $Z_d^{(i)} = \text{DFT}_N\left(z_d^{(n)}\right)$. Equation (7) can be re-written as

$$\mathbf{Y}_d = \sum_{j=1}^{R} C_j^{(0)} \mathbf{H}_{r_j} \overline{\mathbf{X}}_{r_j} + \underbrace{\sum_{j=1}^{R} \left(\mathbf{C}_j - C_j^{(0)} \mathbf{I}\right) \mathbf{H}_{r_j} \overline{\mathbf{X}}_{r_j}}_{ICI} + \mathbf{Z}_d. \tag{9}$$

If we collect the K entries of $\mathbf{Y}_d$ corresponding to the qth SFBC block and form the $K \times 1$ vector $\mathbf{Y}_d^{(q)}$, then we can write

$$\mathbf{Y}_d^{(q)} = \sum_{j=1}^{R} C_j^{(0)} \mathbf{H}_{r_j}^{(q)} \overline{\mathbf{X}}_{r_j}^{(q)} + \sum_{j=1}^{R} \left(\mathbf{C}_j - C_j^{(0)} \mathbf{I}\right)^{[q]} \mathbf{H}_{r_j} \overline{\mathbf{X}}_{r_j} + \mathbf{Z}_d^{(q)}, \tag{10}$$

$\mathbf{H}_j^{(q)} = diag\left[H_j^{((q-1)K+1)}, \cdots, H_j^{(qK)}\right]$, $\mathbf{Z}_d^{(q)} = diag\left[Z_d^{((q-1)K+1)}, \cdots, Z_d^{(qK)}\right]$ and $(.)^{[q]}$ denotes picking the K rows of a matrix starting from $(q-1)K$.

C. *Detection in the Absence of Interference*

For a frequency flat channel with no carrier frequency offset (10) reduces to

$$\mathbf{Y}_d^{(q)} = \sum_{j=1}^{R} \mathbf{H}_{r_j}^{(q)} \mathbf{A}_j \widehat{\mathbf{V}}_j^{(q)} + \mathbf{Z}_d^{(q)}. \tag{11}$$

Using the property of orthogonal codes [6] that $\Re\{\mathbf{A}_j^* \mathbf{A}_i + \mathbf{A}_i^* \mathbf{A}_j\} = 2\delta_{ij}\mathbf{I}$, where δ_{ij} denotes the Kronecker delta, the detection of $\widehat{V}_j^{(q)}$ can be achieved by the operation

$$\begin{aligned} \widehat{\mathbf{Y}}_d^{(q)} &= \sum_{j=1}^{R} \Re\left\{\mathbf{H}_{\text{eq}_j}^{*(q)} \mathbf{Y}_d^{(q)}\right\} \\ &= \mathbf{\Lambda}^{(q)} \widehat{\mathbf{V}}^{(q)} + \widehat{\mathbf{Z}}^{(q)}, \end{aligned} \tag{12}$$

where $\mathbf{H}_{\text{eq}_j}^{(q)} = \mathbf{H}_{r_j}^{(q)} \mathbf{A}_j$, $\mathbf{\Lambda}^{(q)} = \sum_{j=1}^{R} \Re\{\mathbf{H}_{\text{eq}_j}^{*(q)} \mathbf{H}_{\text{eq}_j}^{(q)}\}$ is a diagonal matrix for flat fading. Here the j index of $\widehat{\mathbf{V}}^{(q)}$ is omitted. In the formulation, it is assumed that the symbol estimates does not change across relay nodes. In reality there could be a difference in these estimates. It can be shown that the total noise $\widehat{\mathbf{Z}}^{(q)}$ is white Gaussian and hence Euclidean distance based symbol-by-symbol detection is optimal. On the other hand, for the case when quasi-static assumption is violated (in this case due to frequency selectivity of the channel), then $\mathbf{\Lambda}^{(q)}$ is not diagonal. The optimum detector in this case would be a maximum likelihood detector in P variables, which will have exponential receiver complexity. The complexity is compounded several folds if inter-carrier interference is present due to residual CFO. The ISI and ICI effects are illustrated in Fig. 2. Here, the the absolute power of ISI and ICI are plotted against the number of sub-carriers. As can be seen, the ISI power decreases with increase in number of subcarriers where as the ICI power increases marginally with the number of sub-carriers. The problem of ISI can be circumvented at a system design level by decreasing the sub-carrier spacing but that would mean increase in complexity due to larger number of sub-carriers and also higher ICI due to CFO. At lower number of sub-carriers, both ISI and ICI exist at a substantial level.

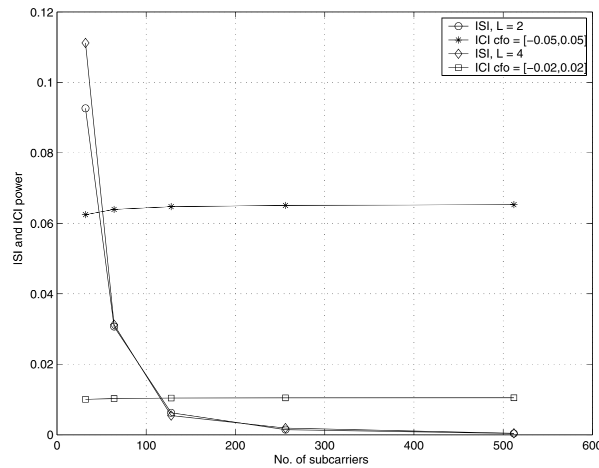

Figure 2: Plot showing ISI and ICI power after combining as a function of number of subcarriers. Frequency-selective fading, ($L = 2$), 2 cooperating users ($R = 2$, Alamouti code), 16-QAM

III. Proposed ISI-ICI cancelling Detector

In this section, we propose a two-step parallel interference canceling (PIC) receiver that cancels the frequency selectivity (FS) induced ISI, and the CFO induced ICI. 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 because of CFO) 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. Hence, 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 in the relay phase. To do that, we split the matrix $\mathbf{H}_{r_j}^{(q)}$ in (7) into two parts; $i)$ a quasi-static part $\mathbf{H}_{\text{qs},r_j}^{(q)}$, and $ii)$ a non-quasi-static part $\mathbf{H}_{\text{nqs},r_j}^{(q)}$, such that

$$\begin{aligned} \mathbf{H}_{r_j}^{(q)} &= \mathbf{H}_{\text{qs},r_j}^{(q)} + \mathbf{H}_{\text{nqs},r_j}^{(q)}, \quad \text{where} \\ \mathbf{H}_{\text{qs},r_j}^{(q)} &= diag\left[H_{r_j}^{((q-1)K+1)}, \cdots, H_{r_j}^{((q-1)K+1)}\right], \\ \mathbf{H}_{\text{nqs},r_j}^{(q)} &= diag\left[0, \Delta H_{r_j}^{(q,2)}, \cdots, \Delta H_{r_j}^{(q,K)}\right], \quad \text{and} \\ \Delta H_{r_j}^{(q,m)} &= H_{r_j}^{((q-1)K+m)} - H_{r_j}^{((q-1)K+1)} \end{aligned} \tag{13}$$

Based on the above formulations, we write (10) as

$$\mathbf{Y}_d^{(q)} = \sum_{j=1}^{R} C_j^{(0)} \mathbf{H}_{\text{qs},r_j}^{(q)} \mathbf{A}_j \widehat{\mathbf{V}}^{(q)} + \underbrace{\sum_{j=1}^{R} C_j^{(0)} \mathbf{H}_{\text{nqs},r_j}^{(q)} \mathbf{A}_j \widehat{\mathbf{V}}^{(q)}}_{\text{QS violation due to FS}} + \underbrace{\sum_{j=1}^{R} (\mathbf{C}_j - C_j^{(0)}\mathbf{I})^{[q]} \mathbf{H}_{r_j} \overline{\mathbf{X}}_{r_j}}_{\text{loss of orthogonality due to CFO}} + \mathbf{Z}_d^{(q)}, \quad (14)$$

Setting $\mathbf{H}_{\text{eq}_j}^{(q)} = \mathbf{H}_{\text{qs},r_j}^{(q)} \mathbf{A}_j$, we can do the combining as in (12) to obtain

$$\begin{aligned}
\widehat{\mathbf{Y}}_d^{(q)} &= \sum_{j'=1}^{R} \Re\left(\mathbf{H}_{\text{eq}_{j'}}^{*(q)} Y_d^{(q)}\right) \\
&= \underbrace{\left(\sum_{j=1}^{R} \Re\left[C_j^{(0)} \mathbf{H}_{\text{eq}_j}^{*(q)} \mathbf{H}_{\text{eq}_j}^{(q)}\right]\right) \widehat{\mathbf{V}}^{(q)}}_{\text{desired signal}} \\
&\quad + \underbrace{\left(\sum_{j'=1}^{R}\sum_{j=1}^{R} \Re\left[C_j^{(0)} \mathbf{H}_{\text{eq}_{j'}}^{*(q)} \mathbf{H}_{\text{nqs},r_j}^{(q)} \mathbf{A}_j\right]\right) \widehat{\mathbf{V}}^{(q)}}_{\text{ISI}} \\
&\quad + \underbrace{\sum_{j'=1}^{R}\sum_{j=1}^{R} \Re\left[\mathbf{H}_{\text{eq}_{j'}}^{*(q)} (\mathbf{C}_j - C_j^{(0)}\mathbf{I})^{[q]} \mathbf{H}_{r_j} \overline{\mathbf{X}}_{r_j}\right]}_{\text{ICI}} + \underbrace{\sum_{j=1}^{R} \Re\left[\mathbf{H}_{\text{eq}_j}^{*(q)} \mathbf{Z}_d^{(q)}\right]}_{\text{noise}}.
\end{aligned} \quad (15)$$

As can be seen, (15) identifies the desired signal, ISI, ICI, and noise components present in the output $\widehat{\mathbf{Y}}_d^{(q)}$. Based on this received signal model in (15) and the knowledge of the matrices $\mathbf{H}_{\text{qs},r_j}^{(q)}$, $\mathbf{H}_{\text{nqs},r_j}^{(q)}$, $\forall\, q, j$ we formulate the proposed interference estimation and cancellation procedure as follows.

1. For each SF code block q, estimate the information symbols $\widehat{\mathbf{V}}^{(q)}$ from (15), ignoring ISI and ICI.
2. For each SF code block q, obtain an estimate of the ISI (i.e., an estimate of the ISI term in (15)) from the estimated symbols $\widehat{\mathbf{V}}^{(q)}$ in the previous step.
3. Cancel the estimated ISI from $\widehat{\mathbf{Y}}_d^{(q)}$.
4. Using $\widehat{\mathbf{V}}^{(q)}$ from step 1, regenerate $\widehat{\overline{\mathbf{X}}}^{(q)}$ using (3). Then, using $\widehat{\overline{\mathbf{X}}}^{(q)}$, obtain an estimate of the ICI (i.e., an estimate of the ICI term in (15)).
5. Cancel the estimated ICI from the ISI canceled output in step 3.
6. Take the ISI and ICI canceled output from step 5 as the input back to step 1 (for the next stage of cancellation).

Based on the above, and $\mathbf{\Lambda}^{(q)} = \sum_{j=1}^{R} \Re\left[C_j^{(0)} \mathbf{H}_{\text{eq}_j}^{*(q)} \mathbf{H}_{\text{eq}_j}^{(q)}\right]$, the cancellation algorithm for the mth stage can be summarized as follows.

Initialization : Set $m = 1$.
Evaluate

$$\widehat{\mathbf{Y}}_d^{(q,m)} = \sum_{j=1}^{R} \Re(\mathbf{H}_{\text{eq}_j}^{*(q)} \mathbf{Y}_d^{(q)}), \quad 1 \le q \le N_g. \quad (16)$$

Loop
Estimate

$$\widehat{\mathbf{V}}^{(q,m)} = (\mathbf{\Lambda}^{(q)})^{-1} \widehat{\mathbf{Y}}_d^{(q,m)}, \quad 1 \le q \le N_g. \quad (17)$$

Cancel ISI

$$\widehat{\mathbf{Y}}^{(q,m+1)} = \widehat{\mathbf{Y}}^{(q,1)} - \sum_{j'=1}^{R}\sum_{j=1}^{R} \Re\left[C_j^{(0)} \mathbf{H}_{\text{eq}_{j'}}^{*(q)} \mathbf{H}_{\text{nqs},r_j}^{(q)} \mathbf{A}_j\right] \widehat{\mathbf{V}}^{(q,m)}, \quad 1 \le q \le N_g. \quad (18)$$

Form $\widehat{\overline{\mathbf{X}}}^{(q,m)}$ *from*

$$\widehat{\overline{\mathbf{X}}}_j^{(q,m)} = \mathbf{A}_j \widehat{\mathbf{V}}^{(q,m)}, \quad 1 \le q \le N_g,\ 1 \le j \le R. \quad (19)$$

Stack $\widehat{\overline{\mathbf{X}}}_j^{(q,m)}$ and form $\widehat{\overline{\mathbf{X}}}_j^{(m)}$
Cancel ICI

$$\widehat{\mathbf{Y}}_d^{(q,m+1)} = \widehat{\mathbf{Y}}_d^{(q,m+1)} - \sum_{j'=1}^{R}\sum_{j=1}^{R} \Re\left[\mathbf{H}_{\text{eq}_{j'}}^{*(q)} (\mathbf{C}_j - C_j^{0}\mathbf{I})^{[q]} \mathbf{H}_{r_j} \widehat{\overline{\mathbf{X}}}_{r_j}^{(m)}\right] \quad (20)$$
$$1 \le q \le N_g.$$

$m = m + 1$
goto Loop.

It is noted that the above cancellation algorithm has polynomial complexity. Also, since $\mathbf{\Lambda}^{(q)}$ is a diagonal matrix, its inversion in (17) is simple.

IV. Simulation Results

We evaluate the BER performance of the proposed interference canceling receiver through simulations. In Fig. (3) we plot the BER performance of two user SFBC OFDM cooperation scheme employing Alamouti code across subcarriers. The total transmit power per symbol is equally divided between broadcast phase and relay phase. The SNR at broadcast phase is kept at 35 dB. Two ray equal power channel is used. The relay phase, nodes have a CFO of [0.1, -0.8] with respect to the destination. 64 subcarriers a re used and 16 QAM modulated symbol is used on each of the subcarriers. The BER performance of flat fading which consume same energy per transmitted symbol and that of non-distributive Alamouti code are given for reference. As can be seen, without interference cancellation (stage 1) the performance is worse than that of an uncoded flat fading. The performance improves significantly with 2 and 3 stages of cancellation. In Fig. (4), three user co-operation system employing rate 3/4 G_3 code and a 128 subcarrier system is considered. The CFO of the nodes are [0.1 -0.08, 0.06]. As can be seen, the performance after interference cancellation converges to a parallel curve within 3 dB of the non-distributive performance of the respective codes. This 3 dB difference is due to half of the power spent on broadcast. In Fig. (5), we plot the BER at a receive SNR of 18 dB (at the destination) against the number of cooperating nodes in the system. The

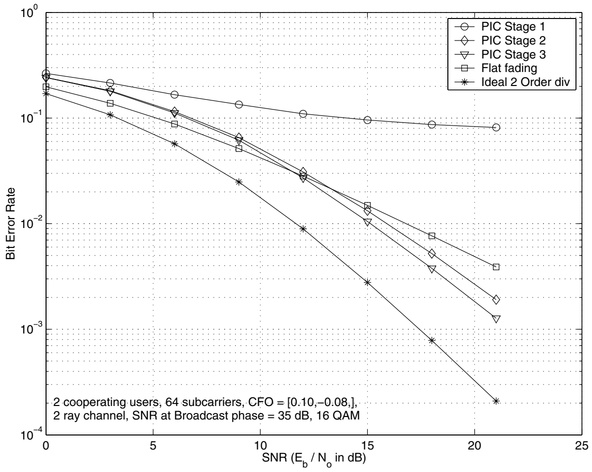

Figure 3: BER performance as a function of SNR for cooperative SFBC OFDM on frequency-selective fading ($L = 2$). $N = 64$, 2 cooperating users ($R = 2$, Alamouti code), CFO = [0.1,-0.08], 16-QAM, SNR on broadcast link = 35 dB.

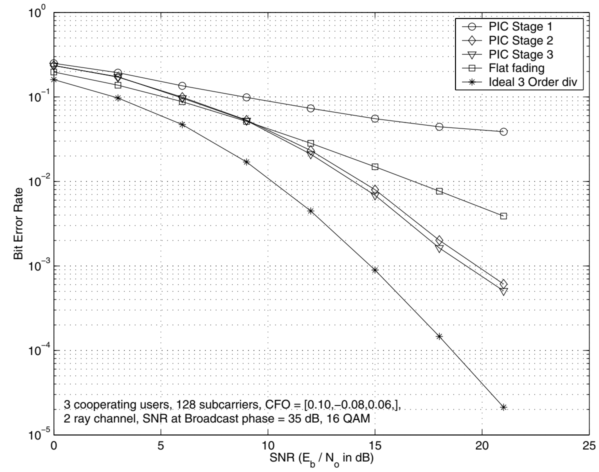

Figure 4: BER performance as a function of SNR for cooperative SFBC OFDM on frequency-selective fading ($L = 2$). $N = 128$, 3 cooperating users ($R = 3$, G_3 code), CFO = [0.1, -0.08, 0.06], 16-QAM, SNR on broadcast link = 35 dB.

broadcast phase SNR is kept at 45 dB. The CFO of the nodes are [0.1, -0.08, 0.06, 0.12, -0.04, -0.07]. The BER of interference free case (flat channel with no CFO) and no cooperation (flat fading) are also plotted for reference. Orthogonal designs G_2, G_3, G_4, G_5 of rates $1, 3/4, 3/4, 2/3$ respectively are used.

V. Conclusions

We propose an interference cancellation algorithm for canceling frequency selectivity induced ISI and carrier frequency induced ICI in a cooperative SFBC OFDM system. In the first step of the proposed algorithm, an estimate of the ISI is obtained and canceled, and in the second step, an estimate of the ICI is obtained and canceled. This two-step procedure is repeated in multiple stages. Our simulation results show that our proposed cancellation algorithm effectively cancels ISI and ICI. The proposed detector can be easily extend to coded cooperative space-frequency systems as well.

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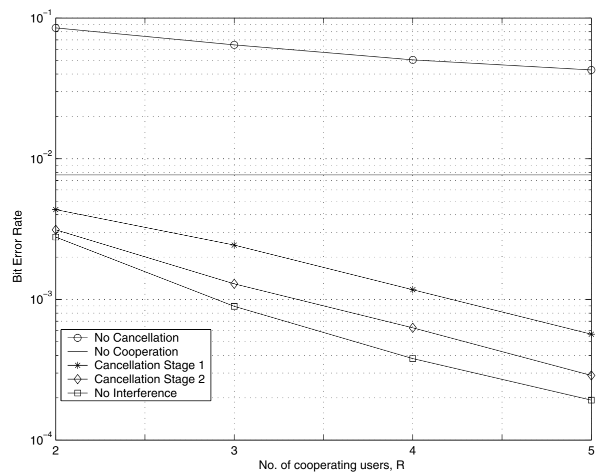

Figure 5: BER performance as a function of number of cooperating users at a receive SNR of 18 dB. CFO = [0.1 -0.08, 0.06, 0.12, -0.04, -0.07], $N = 64$, SNR on broadcast link = 45dB

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