INTERFERENCE SUPPRESSION OF ASYNCHRONOUS MULTI-USER SPACE-TIME BLOCK CODED TRANSMISSIONS

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ABSTRACT

We present a receiver design for joint interference suppression and equalization of asynchronous multi–user space–time block coded (STBC) systems. We describe the design of front end prefilters to synchronize the received blocks from all users. Simulation results show significant performance improvement for asynchronous transmissions.

1. INTRODUCTION

Channel-estimate-based receivers and adaptive receivers for single–user and multiple–user environments assuming synchronous STBC transmissions over flat and frequency selective fading channels have been studied in literature (e.g.,[1],[2]). In the problem treatment, the received signals from all users are usually assumed to be perfectly synchronized.

However, this is often not the case in real systems where the received signals from different users arrive at different times, hence, they are not necessarily synchronized and the data blocks are not aligned with each other. For single carrier frequency domain equalization STBC (SC-FDE STBC) [3], which is a block transmission scheme with cyclic prefixing, it was shown in [2] that we can suppress the interference from multiple synchronous co–channel users very efficiently by exploiting the circulant channel structure. The DFT of the received blocks is used to diagonalize the circulant channel matrices and decouple the users. However, this step cannot be performed if the users are not synchronized. In other words, the received blocks need to be synchronized first before we can exploit the circulant channel structure and diagonalize all subchannels using the DFT. In this paper, we describe the design of a front end prefilter (FEP) to be applied to the received blocks. The purpose of the FEP is to align the asynchronous transmissions from all users.

The paper is organized as follows. In Section 2, we present the system model and the problem formulation in the case of two asynchronous users. In Section 3, we derive expressions for the FEP based on the criterion of energy maximization inside a window of the channel impulse response (CIR). Simulation results are presented in Section 4 and the paper is concluded in Section 5.

Notation: We shall use the notation $\mathbf{h}\star x(n)$, where $\mathbf{h}$ is a column vector of length L to denote the following convolution:

$$\mathbf{h}\star x(n) = \sum_{k=0}^{L-1} h(k)x(n-k)$$

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2. PROBLEM FORMULATION

For simplicity of presentation, we first deal with the two-user case. The extension to the multi-user case is straightforward. The block diagram of the system is shown in Figure 1. It shows two users, each with two antennas, transmitting data using STBC over frequency selective fading channels. The receiver is equipped with two antennas. The received signals $\{y_1(n), y_2(n)\}$ are passed through front end pre-filters $\{\Theta_{ij}(n)\}$ and the outputs $\{z_1(n), z_2(n)\}$ of the filters are combined to form the synchronized received signals and then transformed to the frequency domain where interference suppression and equalization are performed to recover the transmitted users' data $\{\hat{x}(n), \hat{s}(n)\}$.

The received signals at the first and second antennas are given by the convolution of the input signals and the corresponding impulse response sequences of the subchannels of the system:

$$\begin{aligned} y_1(n) &= h_{11}(n)\star x_1(n) + h_{21}(n)\star x_2(n) \\ &+ g_{11}(n)\star s_1(n) + g_{21}(n)\star s_2(n) + n_1(n) \quad (1) \\ y_2(n) &= h_{12}(n)\star x_1(n) + h_{22}(n)\star x_2(n) \\ &+ g_{12}(n)\star s_1(n) + g_{22}(n)\star s_2(n) + n_2(n) \quad (2) \end{aligned}$$

where $x_i(n)$ is the signal transmitted from the i-th antenna of the first user at time n, $s_i(n)$ is the signal transmitted from the i-th antenna of the second user at time n, $h_{ij}(n)$ is the channel impulse response from the first user's i-th antenna to the j-th receive antenna at time n, $g_{ij}(n)$ is the channel impulse response from the second user's i-th antenna to the j-th receive antenna at time n, and $n_1(n)$ and $n_2(n)$ are the noise at the first and second received antennas, respectively, with covariance matrices given by $\mathbf{R}_n = \sigma_n^2\mathbf{I}$. The sequences $x_i(n)$ and $s_i(n)$ have a STBC structure with cyclic prefixes appended to each block. They are generated according to the following encoding rule [2, 3]:

$$\mathbf{x}_{k+1,1}^{(i)}(n)=-\mathbf{x}_{k,2}^{*(1)}\left((-n)_N\right),\ \mathbf{x}_{k+1,2}^{(i)}(n)=\mathbf{x}_{k,1}^{*(i)}\left((-n)_N\right) \quad (3)$$

where $(\cdot)^*$ and $(\cdot)_N$ denote complex conjugation and modulo–N operations, respectively. In addition, a cyclic prefix (CP) of length $P \geq \nu$, where ν denotes the longest channel memory between the transmit antennas and the receive antennas, is added to each transmitted block to eliminate inter–block interference (IBI) and to make all channel matrices *circulant*. The same encoding rule of (3) is used for $s(n)$. Figure 2 shows the structure of $x_i(n)$ and $s_i(n)$. The delay of the second user's cyclic prefix relative to the beginning of the first user's cyclic prefix is denoted by δ.

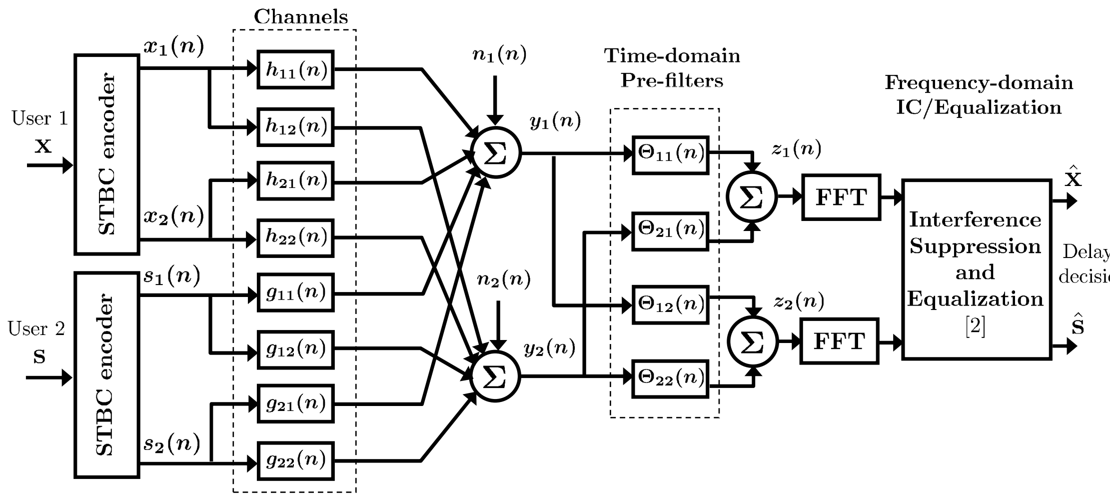

Fig. 1. An overall block diagram for a 2–user system with each user equipped with 2 antennas.

We model this delay as δ zeros at the beginning of the channel impulse response of the second user. This implies that $g_{ij}(n) = 0$ for $n < \delta$.

Four prefilters , $\Theta_{ij}(n)$, $i, j = 1, 2$, are applied to the received signals $y_1(n)$ and $y_2(n)$ to align them. The combined outputs of the filters are given by

$$z_1(n) = y_1(n) \star \Theta_{11}(n) + y_2(n) \star \Theta_{21}(n) \quad (4)$$
$$z_2(n) = y_1(n) \star \Theta_{12}(n) + y_2(n) \star \Theta_{22}(n) \quad (5)$$

All four filters are assumed to have the same length N_Θ. substituting (1) and (2) into (4) and (5) and grouping the terms containing different input signals together, we get the following expression for $z_i(n)$, $i = 1, 2$:

$$\begin{aligned} z_i(n) &= [h_{11}(n) \star \Theta_{1i}(n) + h_{12}(n) \star \Theta_{2i}(n)] \star x_1(n) \\ &+ [h_{21}(n) \star \Theta_{1i}(n) + h_{22}(n) \star \Theta_{2i}(n)] \star x_2(n) \\ &+ [g_{11}(n) \star \Theta_{1i}(n) + g_{12}(n) \star \Theta_{2i}(n)] \star s_1(n) \\ &+ [g_{21}(n) \star \Theta_{1i}(n) + g_{22}(n) \star \Theta_{2i}(n)] \star s_2(n) \\ &+ n_1(n) \star \Theta_{1i}(n) + n_2(n) \star \Theta_{2i}(n) \end{aligned} \quad (6)$$

It is more convenient to express the impulse responses of equivalent channels that result from the convolutions of the channel impulse response sequences $\{h_{ij}(n), g_{ij}(n)\}$ with the impulse response of the pre–filters $\{\Theta_{ij}(n)\}$ in (6) using matrix notation as follows:

$$\begin{aligned} z_i(n) &= \left(\mathbf{H}_{11}\ \mathbf{H}_{12}\right)\begin{pmatrix}\mathbf{\Theta}_{1i}\\ \mathbf{\Theta}_{2i}\end{pmatrix} \star x_1(n) + \left(\mathbf{H}_{21}\ \mathbf{H}_{22}\right)\begin{pmatrix}\mathbf{\Theta}_{1i}\\ \mathbf{\Theta}_{2i}\end{pmatrix} \star x_2(n) \\ &+ \left(\mathbf{G}_{11}\ \mathbf{G}_{12}\right)\begin{pmatrix}\mathbf{\Theta}_{1i}\\ \mathbf{\Theta}_{2i}\end{pmatrix} \star s_1(n) + \left(\mathbf{G}_{21}\ \mathbf{G}_{22}\right)\begin{pmatrix}\mathbf{\Theta}_{1i}\\ \mathbf{\Theta}_{2i}\end{pmatrix} \star s_2(n) \\ &+ n_{z_i}(n) \end{aligned} \quad (7)$$

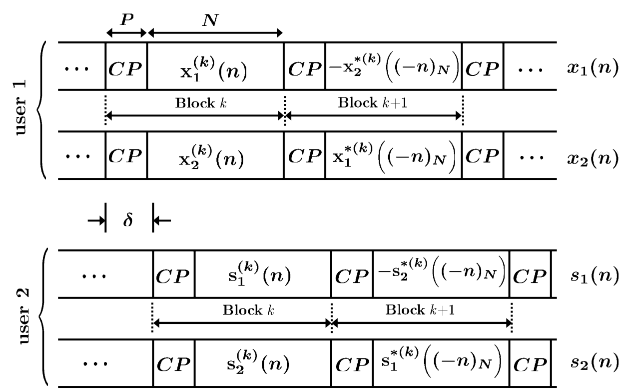

Fig. 2. Data structures for $x_i(n)$ and $s_i(n)$.

where $\mathbf{H}_{ij}$ is the $(N_\Theta + \nu) \times N_\Theta$ convolution matrix given by

$$\mathbf{H}_{ij} = \begin{pmatrix} h_{ij}(0) & & 0 \\ \vdots & \ddots & \\ h_{ij}(\nu) & & h_{ij}(0) \\ & \ddots & \vdots \\ 0 & & h_{ij}(\nu) \end{pmatrix} \quad (8)$$

$\mathbf{G}_{ij}$ is the $(N_\Theta + \nu + \delta) \times N_\Theta$ convolution matrix given by

$$\mathbf{H}_{ij} = \begin{pmatrix} g_{ij}(0) & & 0 \\ \vdots & \ddots & \\ g_{ij}(\nu+\delta) & & g_{ij}(0) \\ & \ddots & \vdots \\ 0 & & g_{ij}(\nu+\delta) \end{pmatrix} \quad (9)$$

$\mathbf{\Theta}_{ij}$ is the $N_\Theta \times 1$ column vector given by

$$\mathbf{\Theta}_{ij} = \left(\Theta_{ij}(0) \quad \ldots \quad \Theta_{ij}(N_\Theta - 1)\right)^T \quad (10)$$

and the noise sequences $n_{z_i}(n)$, $i = 1, 2$, are

$$n_{z_i}(n) \quad = \quad n_1(n) \star \Theta_{1i}(n) + n_2(n) \star \Theta_{2i}(n)$$

Let

$$\mathcal{W}_1 = \begin{pmatrix} \mathbf{\Theta}_{11} \\ \mathbf{\Theta}_{21} \end{pmatrix} \quad , \quad \mathcal{W}_2 = \begin{pmatrix} \mathbf{\Theta}_{12} \\ \mathbf{\Theta}_{22} \end{pmatrix}$$

and

$$\mathcal{H}_i = \begin{pmatrix} \mathbf{H}_{i1} & \mathbf{H}_{i2} \end{pmatrix} \quad , \quad \mathcal{G}_i = \begin{pmatrix} \mathbf{G}_{i1} & \mathbf{G}_{i2} \end{pmatrix}$$

We can then write $z_1(n)$ and $z_2(n)$ alternatively as

$$\begin{aligned} z_1(n) &= \mathcal{H}_1\mathcal{W}_1 \star x_1(n) + \mathcal{H}_2\mathcal{W}_1 \star x_2(n) \\ &\quad + \mathcal{G}_1\mathcal{W}_1 \star s_1(n) + \mathcal{G}_2\mathcal{W}_1 \star s_2(n) + n_{z_1}(n) \\ &\triangleq \tilde{\mathbf{h}}_{11} \star x_1(n) + \tilde{\mathbf{h}}_{21} \star x_2(n) \\ &\quad + \tilde{\mathbf{g}}_{11} \star s_1(n) + \tilde{\mathbf{g}}_{21} \star s_2(n) + n_{z_1}(n) \end{aligned} \tag{11}$$

and

$$\begin{aligned} z_2(n) &= \mathcal{H}_1\mathcal{W}_2 \star x_1(n) + \mathcal{H}_2\mathcal{W}_2 \star x_2(n) \\ &\quad + \mathcal{G}_1\mathcal{W}_2 \star s_1(n) + \mathcal{G}_2\mathcal{W}_2 \star s_2(n) + n_{z_2}(n) \\ &\triangleq \tilde{\mathbf{h}}_{12} \star x_1(n) + \tilde{\mathbf{h}}_{22} \star x_2(n) \\ &\quad + \tilde{\mathbf{g}}_{12} \star s_1(n) + \tilde{\mathbf{g}}_{22} \star s_2(n) + n_{z_2}(n) \end{aligned} \tag{12}$$

where $\tilde{\mathbf{h}}_{ij}$ and $\tilde{\mathbf{g}}_{ij}$ are the equivalent channel impulse response vectors that align the two users. The problem is how to choose $\mathcal{W}_1$ and $\mathcal{W}_2$ so that $s_1(n)$ and $s_2(n)$ are aligned with $x_1(n)$ and $x_2(n)$.

3. DESIGN OF THE FRONT END PRE–FILTERS

Let $\mathbf{e}_i^{(1)}$ denote the i-th unit column vector of size $N_\Theta + \nu$, and let $\mathbf{e}_i^{(2)}$ denote the i-th unit column vector of size $N_\Theta + \nu + \delta$. We are going to select $\tilde{\mathbf{h}}_{ij}$ and $\tilde{\mathbf{g}}_{ij}$ as follows. Let $N_d \geq \delta$ denote some initial delay and let N_s denote some window length. Then we would like the resulting $\{\tilde{\mathbf{h}}_{ij}, \tilde{\mathbf{g}}_{ij}\}$ to have most of their energies concentrated within a window of length N_s following N_d. In order to prevent inter-block-interference at the output of the filters, the filtered channel memory should not exceed the cyclic prefix length, i.e., $N_s \leq P$. In more details, the windows are defined as

$$\tilde{\mathbf{h}}_{ij,win} = \begin{pmatrix} \mathbf{e}_{N_d+1}^{(1)} & \ldots & \mathbf{e}_{N_d+N_s+1}^{(1)} \end{pmatrix}^T \tilde{\mathbf{h}}_{ij} \tag{13}$$

$$\tilde{\mathbf{g}}_{ij,win} = \begin{pmatrix} \mathbf{e}_{N_d+1}^{(2)} & \ldots & \mathbf{e}_{N_d+N_s+1}^{(2)} \end{pmatrix}^T \tilde{\mathbf{g}}_{ij} \tag{14}$$

Let $\tilde{\mathbf{h}}_{ij,wall}$ and $\tilde{g}_{ij,wall}$ denote the remaining samples of $\tilde{\mathbf{h}}_{ij}$ and $\tilde{\mathbf{g}}_{ij}$, respectively. Then

$$\tilde{\mathbf{h}}_{ij,wall} = \begin{pmatrix} \mathbf{e}_1^{(1)} \ldots \mathbf{e}_{N_d}^{(1)} & \mathbf{e}_{N_d+N_s+2}^{(1)} \ldots \mathbf{e}_{N_\Theta+\nu}^{(1)} \end{pmatrix}^T \tilde{\mathbf{h}}_{ij} \tag{15}$$

$$\tilde{\mathbf{g}}_{ij,wall} = \begin{pmatrix} \mathbf{e}_1^{(2)} \ldots \mathbf{e}_{N_d}^{(2)} & \mathbf{e}_{N_d+N_s+2}^{(2)} \ldots \mathbf{e}_{N_\Theta+\nu+\delta}^{(2)} \end{pmatrix}^T \tilde{\mathbf{g}}_{ij} \tag{16}$$

The coefficients of the FEP are computed by maximizing the ratio of the channel energies in $\tilde{\mathbf{h}}_{ij,win}$ and $\tilde{\mathbf{g}}_{ij,win}$ to the sum of the channel energies in $\tilde{\mathbf{h}}_{ij,wall}(n)$ and $\tilde{\mathbf{g}}_{ij,wall}(n)$ plus the noise energy at the filter outputs. Specifically, we solve the following optimization problem (similar to [4, 5]):

$$\max_{\mathcal{W}_j} \left(\sum_{i=1}^{2} \|\tilde{\mathbf{h}}_{ij,win}\|^2 + \|\tilde{\mathbf{g}}_{ij,win}\|^2 \right) \tag{17}$$

subject to the condition

$$\mathcal{W}_j^* \mathbf{R}_n \mathcal{W}_j + \sum_{i=1}^{2} \left(\|\tilde{\mathbf{h}}_{ij,wall}\|^2 + \|\tilde{\mathbf{g}}_{ij,wall}\|^2 \right) = 1$$

for $j = 1, 2$. Using (11)–(16), we rewrite (17) as

$$\max_{\mathcal{W}_j} \mathcal{W}_j^* \mathbf{B}_j \mathcal{W}_j \tag{18}$$

subject to the condition

$$\mathcal{W}_j^* \mathbf{A}_j \mathcal{W}_j = 1$$

where

$$\mathbf{B}_j = \sum_{i=1}^{2} \left(\mathcal{H}_{i,win}^* \mathcal{H}_{i,win} + \mathcal{G}_{i,win}^* \mathcal{G}_{i,win} \right) \tag{19}$$

$$\mathbf{A}_j = \mathbf{R}_n + \sum_{i=1}^{2} \left(\mathcal{H}_{i,wall}^* \mathcal{H}_{i,wall} + \mathcal{G}_{i,wall}^* \mathcal{G}_{i,wall} \right) \tag{20}$$

and

$$\begin{aligned} \mathcal{H}_{i,win} &= \begin{pmatrix} \mathbf{e}_{N_d+1}^{(1)} & \ldots & \mathbf{e}_{N_d+N_s+1}^{(1)} \end{pmatrix}^T \mathcal{H}_i \\ \mathcal{G}_{i,win} &= \begin{pmatrix} \mathbf{e}_{N_d+1}^{(2)} & \ldots & \mathbf{e}_{N_d+N_s+1}^{(2)} \end{pmatrix}^T \mathcal{G}_i \\ \mathcal{H}_{i,wall} &= \begin{pmatrix} \mathbf{e}_1^{(1)} \ldots \mathbf{e}_{N_d}^{(1)} & \mathbf{e}_{N_d+N_s+2}^{(1)} \ldots \mathbf{e}_{N_\Theta+\nu}^{(1)} \end{pmatrix}^T \mathcal{H}_i \\ \mathcal{G}_{i,wall} &= \begin{pmatrix} \mathbf{e}_1^{(2)} \ldots \mathbf{e}_{N_d}^{(2)} & \mathbf{e}_{N_d+N_s+2}^{(2)} \ldots \mathbf{e}_{N_\Theta+\nu}^{(2)} \end{pmatrix}^T \mathcal{G}_i \end{aligned}$$

Introduce the Cholesky factorization $\mathbf{A}_j = \mathbf{L}_{\mathbf{A}_j} \mathbf{L}_{\mathbf{A}_j}^*$, where $\mathbf{L}_{\mathbf{A}_j}$ is a lower–triangular matrix. Then, it is known that the optimum FEP coefficients are given by [6, 7]:

$$\mathcal{W}_{j,opt} = \left(\mathbf{L}_{\mathbf{A}_j}^* \right)^{-1} \mathbf{u}_{j,max} \tag{21}$$

where $\mathbf{u}_{j,max}$ is the orthonormal eigenvector of the matrix $\left(\mathbf{L}_{\mathbf{A}_j} \right)^{-1} \mathbf{B}_j \left(\mathbf{L}_{\mathbf{A}_j}^* \right)^{-1}$ corresponding to the largest eigenvalue $\lambda_{j,max}$. The Shortening Signal to Noise ratio (SSNR) is defined as follows

$$\begin{aligned} \text{SSNR}_{j,opt} &\triangleq 10 \log \left(\frac{\mathcal{W}_{j,opt}^* \mathbf{B}_j \mathcal{W}_{j,opt}}{\mathcal{W}_{j,opt}^* \mathbf{A}_j \mathcal{W}_{j,opt}} \right) \\ &= 10 \log \left(\lambda_{j,max} \right) \end{aligned} \tag{22}$$

The equivalent system is shown in Figure 3. The impulse response sequences of $\tilde{\mathbf{h}}_{ij,win}$ and $\tilde{\mathbf{g}}_{ij,win}$ are now aligned and the equivalent system consists of two synchronous users. Therefore, the output of the prefilter can be transformed to frequency-domain using DFT and the joint interference suppression and equalization technique described in [2] can now be applied.

4. SIMULATION RESULTS

We simulate a system with two users, each equipped with two transmit antennas. The number of receive antennas is equal to the number of users. The channels from each transmit antenna to each receive antenna are assumed to be independent. A Typical

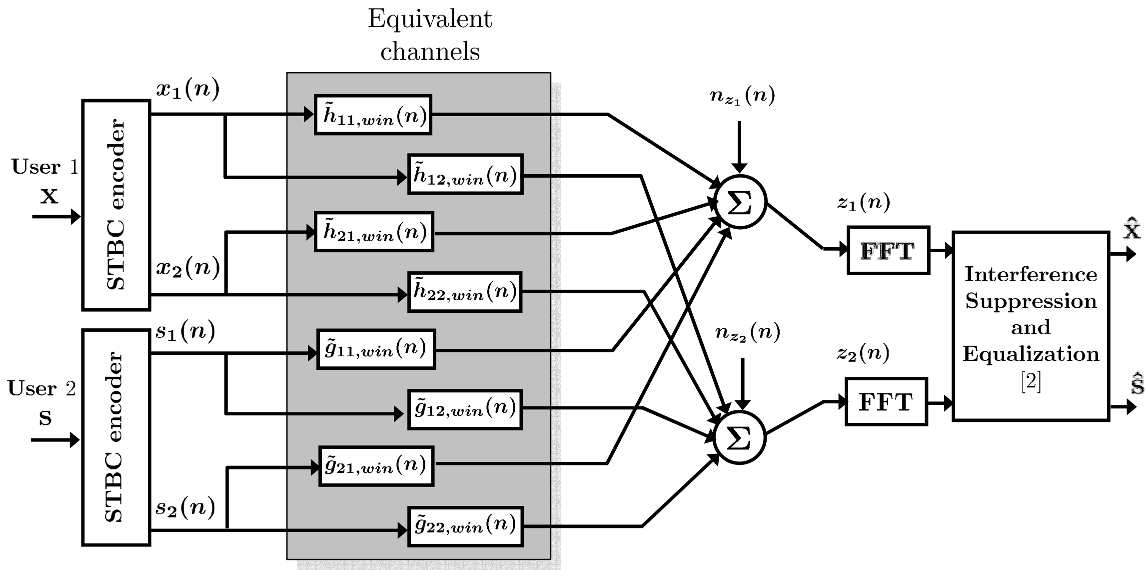

Fig. 3. Block diagram of the equivalent system.

Urban (TU) channel model with overall channel impulse response memory ν equal to 3 is considered for all channels. Moreover, a linearized GMSK transmit pulse shape is used. The data bits of each user are mapped into an 8-PSK signal constellation and they are grouped into blocks of 32 symbols. A cyclic prefix is added to each block. The processed blocks are transmitted at a symbol rate equal to 271 KSymbols/sec. The signal to noise ratio of the two users at the receiver are assumed to be equal. To simulate the performance of the receiver, we assume perfect knowledge of the channel impulse response at the receiver. We calculate the FEP coefficients using (21) and use them to obtain the equivalent channel impulse response. We then use the equivalent channel coefficients to compute the MMSE equalizer coefficients according to [2]. Figure 4 shows the overall system Bit Error Rate for different values of δ. It is obvious that the FEP improves the performance of the asynchronous case significantly.

5. CONCLUSIONS

In this paper, we presented a pre–filtering technique for the synchronization of asynchronous multi–user space–time coded transmissions. We showed how to choose the pre–filter coefficients based on energy maximization inside a certain window of the equivalent channels' impulse response sequences. Simulation results show that the system performance is close to the performance in the synchronous case.

6. REFERENCES

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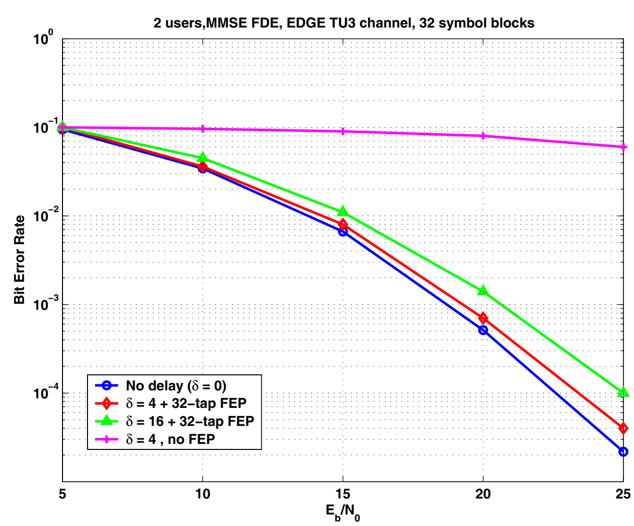

Fig. 4. BER performance of the receiver.