Analysis of Equalization Pos-Combining with Channel Estimation and MIMO Joint Equalization-Combining Receivers for Space-Time Block Coding

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Resumo–Neste artigo é investigado o desempenho de duas propostas diferentes de receptor empregando codificação espaço-temporal em bloco para canais seletivos em freqüência. A primeira, designada como receptor com equalização pós-combinação e estimativa de canal, é baseada no método de codificação espaço-temporal com inversão temporal proposto por Lindskog e Paulraj e realiza estimativa do canal, combinação linear e equalização. A segunda, denominada como receptor com equalização e combinação conjunta de múltiplas entradas e múltiplas saídas, é baseada no esquema apresentado por Meshkati e Sousa e realiza equalização, cancelamento de interferência e combinação linear simultaneamente. Um dos principais resultados de simulação obtidos mostra que o receptor com equalização e combinação conjunta de múltiplas entradas e múltiplas saídas oferece um ganho de desempenho em relação ao receptor com equalização pós-combinação e estimativa de canal à medida que a relação sinal-ruído aumenta.

Abstract–In this paper, we investigate the performance of two different receiver proposals using space-time block coding for frequency selective fading channels. The first one, called equalization pos-combining with channel estimation receiver, is based on the time-reverse space-time block coding proposed by Lindskog and Paulraj and it performs channel-estimation, linear combination and equalization. The second one, denoted multiple-input multiple-output joint equalization-combining receiver, is based on the scheme presented by Meshkati and Sousa and it performs channel equalization, interference cancellation and linear combination simultaneously. One of the main simulation results shows that multiple-input multiple-output joint equalization-combining receiver outperforms equalization pos-combining with channel estimation as signal to noise ratio increases.

I. INTRODUCTION

Recently, space-time coding has been proposed to achieve transmit diversity. There are two main schemes in space-time coding: space-time trellis coding (STTC) and space-time block coding (STBC). Although STTC offers coding gain along with diversity [1], STBC has emerged as a promising spatial transmit diversity scheme due to simple decoding complexity at the receiver and for not requiring the channel state information (CSI) at the transmitter. Initially, Alamouti introduced a remarkable transmit diversity scheme in [2] that established the basis for STBC. Later, Tarokh *et al.* generalized STBC to an arbitrary number of transmit antennas [3], [4]. The orthogonal structure of STBC provides decoupling of signals from different antennas and presents a decoding complexity dependent only on constellation size [5].

STBC was designed on the assumption of flat multipath fading channel. However, in high data rate wireless communications, the delay spread is higher than the symbol duration, which gives rise to frequency selective propagation effects. Unlike flat fading channels, optimal design of STBC for frequency selective fading channels is very complex. In order to maintain simple decoding complexity and taking advantage of existing STBC schemes for flat fading channels, most of the works presented in the literature have treated intersymbol interference (ISI) and diversity separately, leading to suboptimal results.

One of these works is the time-reverse space-time block coding (TR-STBC) proposed by Lindskog and Paulraj in [6], that extends space-time block coding from the symbol-level to the block-level to treat frequency selective fading channels. In this approach, a data frame is serial to parallel converted and symbol mapped in two symbol streams. Each transmit symbol frame is divided in two blocks. During the first block, symbol stream 1 is transmitted from antenna 1 and symbol stream 2 is transmitted from antenna 2. In the second block, symbol stream 2 is time reversed, negated and complex conjugated before being transmitted from antenna 1 and symbol stream 1 is time reversed and complex conjugated before being transmitted from antenna 2. At the receiver, channel equalization is performed after linear combining and space-time block decoding to recover the transmit data symbols. Another work to treat frequency selective fading channels is presented in [7]. In this work, Meshkati and Sousa presented a TR-STBC scheme in a multiple-input multiple-output (MIMO) framework and MIMO chip-equalization was used to perform channel equalization, interference cancellation and linear combination.

In this paper, we present an equalization pos-combining with channel estimation receiver (EPCCE) based on a minimum mean square error (MMSE) TR-STBC using recursive least squares channel estimation (RLS-CEs) and recursive least squares channel equalization (RLS-CEq) and we compare its performance against a MIMO joint equalization-combining receiver at symbol-level (MIMO-JEC) based on the receiver presented in [7] and against an ideal equalization pos-combining (EPC), which performs perfect channel estimation.

The paper is organized as follows: the EPCCE is introduced in section II; the MIMO-JEC is presented in section III; simulation results are presented in section IV and finally, conclusion remarks are presented in section V.

II. EQUALIZATION POS-COMBINING WITH CHANNEL ESTIMATION

We consider that the base station employs two transmit antennas placed far enough so that their signals undergo independent fading. A given data frame is firstly serial to parallel converted and then symbol mapped into two symbol streams, $\mathbf{b}_1$ and $\mathbf{b}_2$, respectively. Each transmit symbol frame is divided in two blocks. During the first block, $\mathbf{b}_1$ is transmitted from antenna 1 and $\mathbf{b}_2$ is transmitted from antenna 2. In the second block, $\mathbf{b}_2$ is time reversed, negated and complex conjugated before be transmitted from antenna 1 and $\mathbf{b}_1$ is time reversed and complex conjugated before be transmitted from antenna 2.

Assuming that the frequency selective fading channels from antenna 1 and antenna 2 are stationary during a frame duration, and that its impulse response can be modeled as FIR filters, we can represent the channel vector from transmit antenna i to receive antenna j by:

$$\mathbf{h}_{i,j}(n)=\sum_{k=1}^{L}\beta_{i,j}^{k}\cdot\psi\left(n-\tau_{i,j}^{k}\right) \qquad (1)$$

Where L is the number of resolvable multipath components; $\beta_{i,j}^{k}$ and $\tau_{i,j}^{k}$ are, respectively, the complex gain and time delay of the kth multipath component of the channel from antenna i to receive antenna j; $\psi(n)$ is the filtered waveform, which includes the effects of transmitter and receiver filters.

We can represent the received signal at the jth receive antenna and in the block interval *block* by:

$$\mathbf{r}_{j}^{block}(n)=\sum_{i=1}^{2}\mathbf{h}_{i,j}(n)*\mathbf{s}_{i}^{block}(n)+\mathbf{v}_{j}^{block}(n) \qquad (2)$$

Where $\mathbf{s}_{i}^{block}(n)$ is the transmitted signal from transmit antenna i at time n; $\mathbf{v}_{j}^{block}(n)$ is the complex white Gaussian noise at time n, and $*$ represents convolution operation.

Considering first the case of a receiver employing just one receive antenna ($j=1$) and assuming that the STBC codification matrix is given by [6]:

$$\mathbf{s}(n)=\begin{bmatrix}\mathbf{s}_{1}^{1}(n) & \mathbf{s}_{2}^{1}(n)\\ \mathbf{s}_{1}^{2}(n) & \mathbf{s}_{2}^{2}(n)\end{bmatrix}=\begin{bmatrix}\mathbf{b}_{1}(n) & \mathbf{b}_{2}(n)\\ -\mathbf{b}_{2}(-n)^{*} & \mathbf{b}_{1}(-n)^{*}\end{bmatrix} \qquad (3)$$

Where $\mathbf{b}_{i}(n)$ is the transmitted stream from antenna i at time n given by: $\mathbf{b}_{i}(n)=\begin{bmatrix}b_{i}(n) & b_{i}(n-1) & \cdots & b_{i}(n-N_{s}+1)\end{bmatrix}$ and N_{s} is the number of symbols per block.

We can represent the received signal at the receive antenna in block 1 and 2 in a vector notation by:

$$\begin{bmatrix}\mathbf{r}_{1}^{1}(n)\\ \mathbf{r}_{1}^{2}(n)\end{bmatrix}=\begin{bmatrix}\mathbf{b}_{1}(n) & \mathbf{b}_{2}(n)\\ -\mathbf{b}_{2}(-n)^{*} & \mathbf{b}_{1}(-n)^{*}\end{bmatrix}*\begin{bmatrix}\mathbf{h}_{1,1}(n)\\ \mathbf{h}_{2,1}(n)\end{bmatrix}+\begin{bmatrix}\mathbf{v}_{1}^{1}(n)\\ \mathbf{v}_{1}^{2}(n)\end{bmatrix} \qquad (4)$$

We can recover the transmit symbols by extracting $\mathbf{b}_1(n)$ and $\mathbf{b}_2(n)$, from the received signals at block 1 and block 2 using the scheme presented in [6]. Let $\mathbf{x}_{1}^{1}(n)\hat{=}\mathbf{r}_{1}^{1}(n)$ and $\mathbf{x}_{1}^{2}(n)\hat{=}\mathbf{r}_{1}^{2}(-n)^{*}$. Combining $\mathbf{x}_{1}^{1}(n)$ and $\mathbf{x}_{1}^{2}(n)$ by providing perfect estimation (or trained-aided estimation) of the fading channels, we can extract the streams $\mathbf{y}_1(n)$ and $\mathbf{y}_2(n)$ using simple signal processing.

Specifically, in order to obtain stream $\mathbf{y}_1(n)$, we can combine the modified received streams at block 1 and 2, $\mathbf{x}_{1}^{1}(n)$ and $\mathbf{x}_{1}^{2}(n)$, as follows:

$$\mathbf{y}_{1}(n)=\mathbf{h}_{1,1}(-n)^{*}*\mathbf{x}_{1}^{1}(n)+\mathbf{h}_{2,1}(n)*\mathbf{x}_{1}^{2}(n)^{*} \qquad (5)$$

Similarly, we can obtain stream $\mathbf{y}_2(n)$ by:

$$\mathbf{y}_{2}(n)=\mathbf{h}_{2,1}(-n)^{*}*\mathbf{x}_{1}^{1}(n)-\mathbf{h}_{1,1}(n)*\mathbf{x}_{1}^{2}(n)^{*} \qquad (6)$$

$\mathbf{y}_1(n)$ and $\mathbf{y}_2(n)$ are decoupled versions of symbol streams $\mathbf{b}_1(n)$ and $\mathbf{b}_2(n)$. However, it is still necessary to mitigate ISI for perfect recovery of $\mathbf{b}_1(n)$ and $\mathbf{b}_2(n)$.

We can obtain an optimum weight vector, $\mathbf{w}_i$, $i=1,2$, in the LS sense for appropriate channel equalization by [8]:

$$\mathbf{w}_{i}=\left(\frac{1}{N_{t}}\sum_{n=0}^{N_{t}-1}\mathbf{y}_{i}(n)\cdot\mathbf{y}_{i}^{H}(n)\right)^{-1}\cdot\left(\frac{1}{N_{t}}\sum_{n=0}^{N_{t}-1}b_{i}^{train}(n)\cdot\mathbf{y}_{i}(n)\right) \qquad (7)$$

Where N_t is the number of training symbols available and $b_{i}^{train}(n)$ is the nth training symbol of the ith symbol stream.

Recursive least squares (RLS) can also be used to obtain $\hat{\mathbf{w}}_i$, $i=1$, 2. In the following, we describe briefly the algorithm (for additional information see [8]).

i. Initialize $\mathbf{w}_{1}^{0}=\mathbf{0}$ and $\hat{\mathbf{R}}_{N_{t}}^{-1}=\left(\sigma^{2}\right)^{-1}\cdot\mathbf{I}$ (8)

ii. Compute $\mathbf{K}_{gain}(n)=\dfrac{\lambda^{-1}\cdot\hat{\mathbf{R}}_{N_{t}}^{-1}\cdot\mathbf{y}_{i}(n)^{*}}{1+\lambda^{-1}\cdot\mathbf{y}_{i}(n)^{T}\cdot\hat{\mathbf{R}}_{N_{t}}^{-1}\cdot\mathbf{y}_{i}(n)^{*}}$ (9)

iii. Determine $\varepsilon(n)=b_{i}^{train}(n)-\mathbf{y}_{i}(n)^{T}\cdot\mathbf{w}_{i}^{n-1}$ (10)

iv. Compute $\mathbf{w}_{i}^{n}=\mathbf{w}_{i}^{n-1}+\mathbf{K}_{gain}(n)\cdot\varepsilon(n)$ (11)

v. Update $\hat{\mathbf{R}}_{N_{t}}^{-1}=\lambda^{-1}\cdot\left\{\hat{\mathbf{R}}_{N_{t}}^{-1}-\mathbf{K}_{gain}(n)\cdot\mathbf{y}_{i}(n)^{T}\cdot\hat{\mathbf{R}}_{N_{t}}^{-1}\right\}$ (12)

Where $\hat{\mathbf{R}}_{N_{t}}^{-1}$ is the inverse of the correlation matrix, $\mathbf{I}$ is the identity matrix, σ^{2} is the noise variance at each receive antenna and λ is the forgetting factor.

Using this procedure, we can determine the estimates $\hat{\mathbf{b}}_1(n)$ and $\hat{\mathbf{b}}_2(n)$ of the transmitted symbol streams. We can also use RLS in a similar manner to perform channel estimation before linear combining to obtain $\mathbf{y}_1(n)$ and $\mathbf{y}_2(n)$.

The method presented can be extended to an arbitrary number of received antennas. The encoding and transmission will be identical to the case of one receive antenna and for the two receive antennas case ($j=2$), the received signals in blocks 1 and 2 are given by:

$$\begin{bmatrix} \mathbf{r}_1^1(n) \\ \mathbf{r}_1^2(n) \\ \mathbf{r}_2^1(n) \\ \mathbf{r}_2^2(n) \end{bmatrix} = \begin{bmatrix} \mathbf{b}_1(n) & \mathbf{b}_2(n) & & \\ -\mathbf{b}_2(-n)^* & \mathbf{b}_1(-n)^* & & \mathbf{0} \\ & & \mathbf{b}_1(n) & \mathbf{b}_2(n) \\ \mathbf{0} & & -\mathbf{b}_2(-n)^* & \mathbf{b}_1(-n)^* \end{bmatrix} * \begin{bmatrix} \mathbf{h}_{1,1}(n) \\ \mathbf{h}_{2,1}(n) \\ \mathbf{h}_{1,2}(n) \\ \mathbf{h}_{2,2}(n) \end{bmatrix}$$

$$+ \begin{bmatrix} \mathbf{v}_1^1(n) \\ \mathbf{v}_1^2(n) \\ \mathbf{v}_2^1(n) \\ \mathbf{v}_2^2(n) \end{bmatrix} \quad (13)$$

When the receiver employs two receive antennas, we can obtain stream $\mathbf{y}_1(n)$ by:

$$\mathbf{y}_i(n) = \mathbf{h}_{1,1}(n)^* * \mathbf{x}_1^1(n) + \mathbf{h}_{2,1}(n) * \mathbf{x}_1^2(n) + \mathbf{h}_{1,2}(n)^* * \mathbf{x}_2^1(n)$$
$$+ \mathbf{h}_{2,2}(n) * \mathbf{x}_2^2(n) \quad (14)$$

And we can obtain stream $\mathbf{y}_2(n)$ by:

$$\mathbf{y}_2(n) = \mathbf{h}_{2,1}(n)^* * \mathbf{x}_1^1(n) - \mathbf{h}_{1,1}(n) * \mathbf{x}_1^2(n) + \mathbf{h}_{2,2}(n)^* * \mathbf{x}_2^1(n)$$
$$- \mathbf{h}_{1,2}(n) * \mathbf{x}_2^2(n) \quad (15)$$

Using the same equalization procedure presented for the one single receive antenna case, we can also obtain the estimates $\hat{\mathbf{b}}_1(n)$ and $\hat{\mathbf{b}}_2(n)$ of the transmitted symbol streams.

III. MIMO EQUALIZATION-COMBINING

Streams $\mathbf{y}_1(n)$ and $\mathbf{y}_2(n)$ obtained by (5), (6) and (14), (15) for the one single receive antenna case and for the two receive antennas case, respectively, can be viewed as a MIMO system. It is also possible to include the equalization procedure in this MIMO framework [7]. In Fig.1, we have the representation of MIMO-JEC for the one receive antenna case, where $\mathbf{x}_1^1(n)$ and $\mathbf{x}_1^2(n)$ are the inputs and $\hat{\mathbf{b}}_1(n)$ and $\hat{\mathbf{b}}_2(n)$ are the outputs.
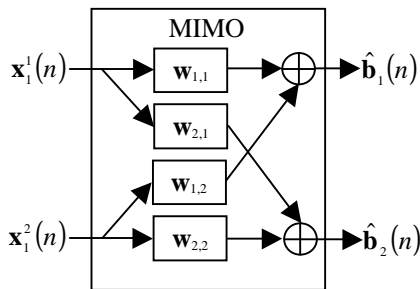

Fig.1. MIMO Equalization Combining (one receive antenna)

Considering the mathematical manipulations presented in [7], and taking into account system differences, we can obtain $\hat{\mathbf{b}}_1(n)$ and $\hat{\mathbf{b}}_2(n)$ for the one receive antenna case ($j=1$), including diversity and equalization, by:

$$\hat{\mathbf{b}}_1(n) = \mathbf{w}_{1,1}(n) * \mathbf{x}_1^1(n) + \mathbf{w}_{1,2}(n) * \mathbf{x}_1^2(n) \quad (16)$$

And,

$$\hat{\mathbf{b}}_2(n) = \mathbf{w}_{2,1}(n) * \mathbf{x}_1^1(n) + \mathbf{w}_{2,2}(n) * \mathbf{x}_1^2(n) \quad (17)$$

In Fig.2, we present the MIMO-JEC for the two receive antennas case, where $\mathbf{x}_1^1(n)$ and $\mathbf{x}_1^2(n)$ are the inputs corresponding to receive antenna 1 and $\mathbf{x}_2^1(n)$ and $\mathbf{x}_2^2(n)$ are the inputs corresponding to receive antenna 2.
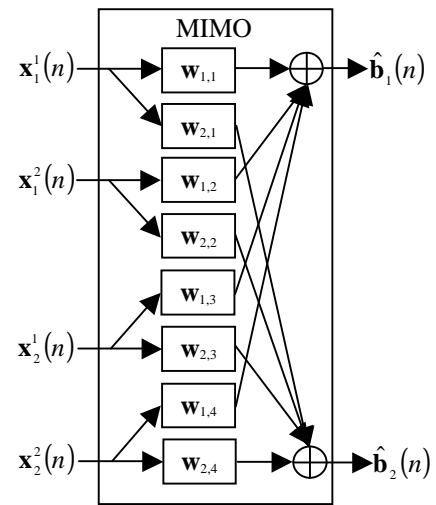

Fig.2. MIMO Equalization Combining (two receive antennas)

For the two receive antennas case ($j=2$), we extend the results presented in [7] to obtain the estimates $\hat{\mathbf{b}}_1(n)$ and $\hat{\mathbf{b}}_2(n)$ by the following procedure:

$$\hat{\mathbf{b}}_1(n) = \mathbf{w}_{1,1}(n) * \mathbf{x}_1^1(n) + \mathbf{w}_{1,2}(n) * \mathbf{x}_1^2(n) + \mathbf{w}_{1,3}(n) * \mathbf{x}_2^1(n)$$
$$+ \mathbf{w}_{1,4}(n) * \mathbf{x}_2^2(n) \quad (18)$$

And,

$$\hat{\mathbf{b}}_2(n) = \mathbf{w}_{2,1}(n) * \mathbf{x}_1^1(n) + \mathbf{w}_{2,2}(n) * \mathbf{x}_1^2(n) + \mathbf{w}_{2,3}(n) * \mathbf{x}_2^1(n)$$
$$+ \mathbf{w}_{2,4}(n) * \mathbf{x}_2^2(n) \quad (19)$$

MIMO-JEC coefficients can be obtained by the joint space-time RLS adaptation procedure presented in [9]. From Fig.1 and Fig.2, we notice that MIMO-JEC presents a more complex structure than EPPCE and it requires training symbols in the two transmit frame blocks (in both transmit streams from antenna 1 and antenna 2) for correctly obtaining $\hat{\mathbf{b}}_1(n)$ and $\hat{\mathbf{b}}_2(n)$.

IV. SIMULATION RESULTS

In this section, we compare the performance of EPCCE and MIMO-JEC receivers for a wireless communication system employing QPSK data modulation in a frequency selective multipath fading channel. The transmitter employs two transmit antennas whose signals present independent fading and the receiver may use one or two receive antennas ($j=1,2$).

For the simulations, EPC receiver is used as a benchmark of performance. EPC assumes perfect channel estimation and equalization is performed by using the RLS algorithm, EPCCE employs RLS for both channel estimation and equalization, and MIMO-JEC employs MIMO-RLS for the joint equalization-combining procedure [9]. In all three schemes, it is considered that $\lambda=1$.

Simulations are performed varying the number of training symbols and signal to noise ratio (SNR). The transmitted frames are composed by two blocks of 250 symbols (N_s=500) and simulation results are obtained computing 4000 frames (N_{fr}=4000). Exception is Fig.9, where the transmitted frames are composed by two blocks of 500 symbols (N_s=1000) to allow investigation of the asymptotic behavior of the analyzed schemes. Training is available in both frame blocks (due to STBC structure) and guard symbols are transmitted to avoid interblock interference.

We assume that the propagation channel coefficients are independent Rayleigh distributed with order 2 ($L=3$) and block time-invariant. For simulations purpose, we consider that the length of the channel estimation filter is of the same length of the channel and that the length of each equalizer filter (or subfilter) is 2 times the channel length plus 1.

In Fig.3 and Fig.4, we compare the performance of EPCCE, EPC and MIMO-JEC varying the number of training symbols per transmit antenna for SNR=10dB, using one and two receive antennas, respectively. The results for the one receive antenna case, show that MIMO-JEC outperforms EPCCE when the number of training symbols is higher than 30. For the two receive antennas case, EPCCE outperforms MIMO-JEC for training symbols up to 50 symbols per transmit antenna. We can also see that perfect channel estimation is an issue for improving system performance.

In Fig.5 and Fig.6, we compare the performance of EPCCE, EPC and MIMO-JEC varying the number of training symbols per transmit antenna for SNR=15dB, using one and two receive antennas, respectively. For the one receive antenna case, MIMO-JEC outperforms EPCCE when the number of training symbols is higher than 12 and for the two receive antennas case, MIMO-JEC outperforms EPCCE when the number of training symbols is higher than 40.

In Fig.7 and Fig.8, we compare the performance of EPCCE, EPC and MIMO-JEC varying SNR using one and two receive antennas, respectively. For the one receive antenna case, we use N_t=25 and for the two receive antennas case, we consider N_t=50. The results show that MIMO-JEC outperforms EPCCE as SNR increases (higher than 10dB for the one receive antenna case and higher than 13 db for the two receive antennas case).

Finally, in Fig.9 is presented an estimate of the asymptotic behavior of the analyzed receivers as a function of the number of training symbols for SNR=10dB. For the one receive antenna case, all receivers present similar asymptotic performance and for the two receive antennas case, MIMO-JEC presents a slight better asymptotic performance.
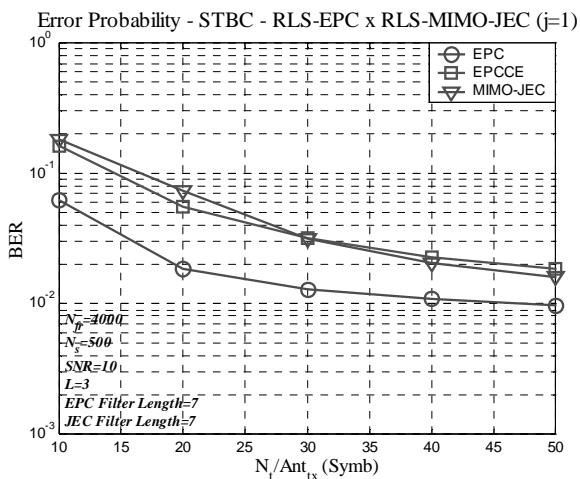

Fig.3. Comparison between EPC, EPCCE and MIMO-JEC varying the number of training symbols for SNR=10dB (one receive antenna)
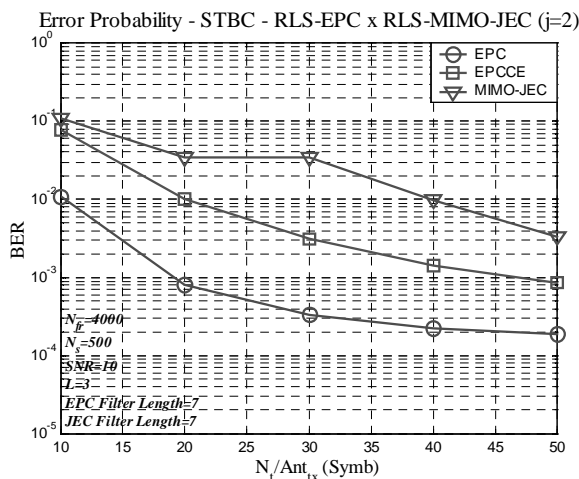

Fig.4. Comparison between EPC, EPCCE and MIMO-JEC varying the number of training symbols for SNR=10dB (two receive antennas)
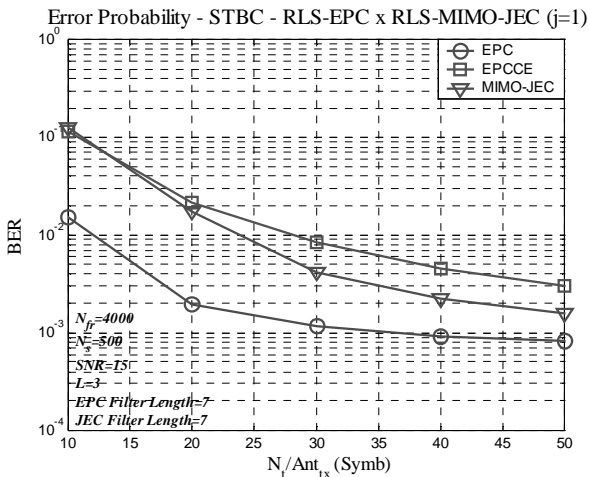
Fig.5. Comparison between EPC, EPCCE and MIMO-JEC varying the number of training symbols for SNR=15dB (one receive antenna)
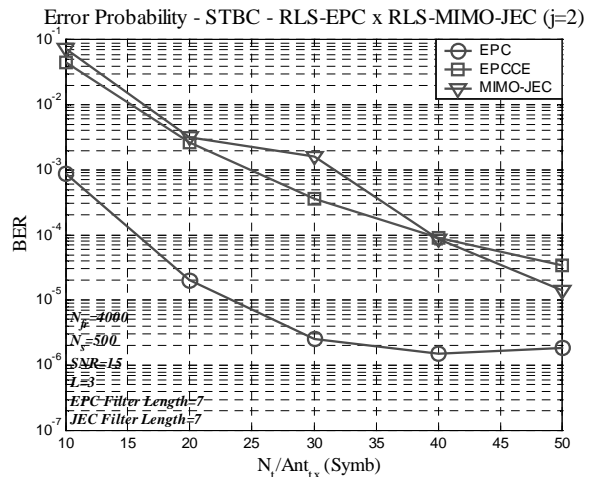
Fig.6. Comparison between EPC, EPCCE and MIMO-JEC varying the number of training symbols for SNR=15dB (two receive antennas)
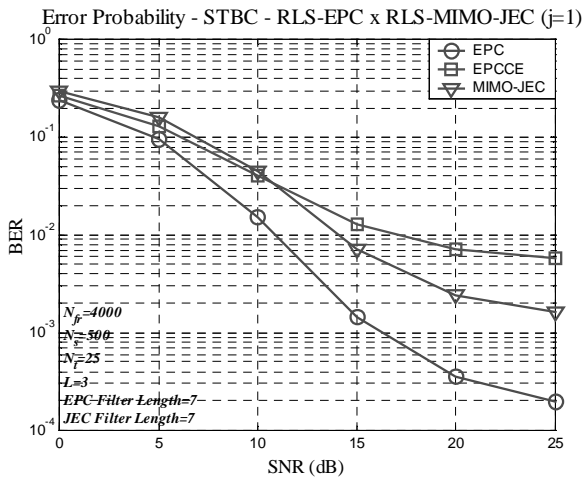
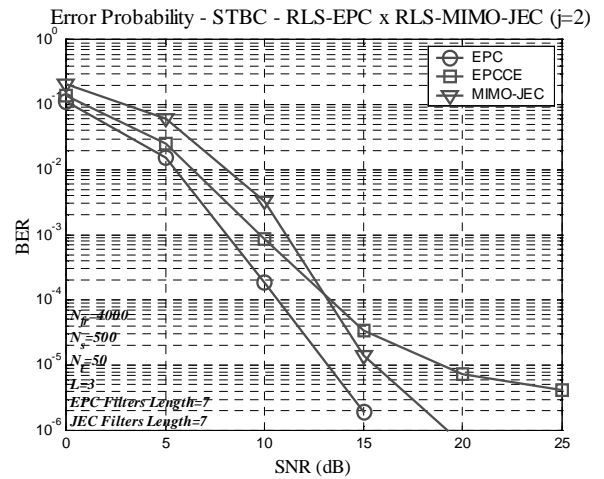Fig.7. Comparison between EPC, EPCCE and MIMO-JEC varying SNR for N_t=25 (one receive antenna)

Fig.8. Comparison between EPC, EPCCE and MIMO-JEC varying SNR for N_t=50 (two receive antennas)
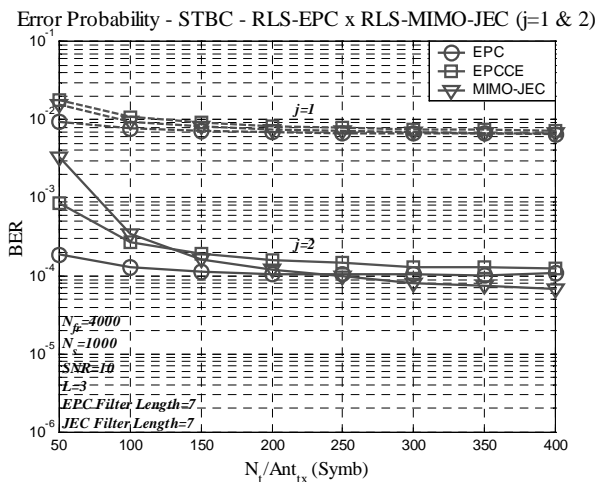
Fig.9. Asymptotic Performance of EPC, EPCCE and MIMO-JEC varying the number of training symbols for SNR=10dB (one and two receive antennas)

V. CONCLUSIONS

In this paper, we presented two modified receivers employing space-time block coding for wireless communication systems in frequency selective multipath fading channels. The first one, denoted by EPCCE receiver, is based on a MMSE implementation of the TR-STBC. The second one, denoted MIMO-JEC receiver, is based on the chip level MIMO equalizer presented in [7].

Performance comparison between EPCCE and MIMO-JEC varying the number of training symbols and SNR were performed by simulation. EPC, which performs perfect channel estimation, was used as a reference for the results.

Simulation results showed that MIMO-JEC outperforms EPCCE as SNR increases. For low and moderate SNR, MIMO-JEC outperforms EPCCE if enough training symbols are available; otherwise EPCCE will present performance improvement against MIMO-JEC.

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