Differential Semi–Blind Spatial-Temporal Beamforming for DS–WCDMA Systems in Frequency Selective Time-Varying Channels

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***Resumo*– Neste artigo, é proposto um novo receptor empregando formatação de feixe espaço-temporal semi-cega diferencial para sistemas assíncronos de múltiplo acesso por divisão de código de banda-larga por seqüência direta (DS-WCDMA). O novo receptor apresenta robustez a variações de fase e redução dos efeitos de distorção não linear.**

***Abstract*–In this paper, we propose a new differential semi-blind spatial-temporal beamforming receiver for asynchronous direct sequence wideband code division multiple access (DS-WCDMA) systems. The new receiver presents improved robustness to phase variations and reduction of the effects of nonlinear distortion.**

I. INTRODUCTION

Although the main problem arising from high data rate DS-WCDMA systems is the severe inter-symbol interference (ISI), in addition to the multiple access interference (MAI), the presence of a time-varying fading channel, even with a relative small Doppler spread (compared to the symbol duration), may also reduce overall system performance. In this situation, it may be convenient to employ differential encoding/decoding schemes. Using this simple approach, phase fluctuations in the received signal do not affect system performance as long as they can be considered negligible within a symbol interval. We note, however, that this robustness to channel variability is achieved at the expense of a higher signal-to-noise ratio (SNR).

As we address the problem of differential detection, the method of data modulation considered is the $\pi/4$ shifted DQPSK, which eliminates the needs for accurate carrier recovery and greatly simplifies the implementation of the receiver. The choice is motivated by the efficiency of the method with respect to traditional QPSK or OQPSK, which require coherent detection [1]. Another advantage is that transitions in the signal constellation do not pass through the origin that reduces the effects of nonlinear distortion caused by the R.F. amplifier. This is obtained by the superposition of two QPSK signal constellations offset by 45 degrees relative to each other, resulting eight phases. The phases are alternatively selected from one of the two QPSK constellations [2].

In [3], a semi-blind spatial-temporal beamforming receiver (BST-SBCMACI) based on the SBCMACI (semi-blind constant modulus algorithm with channel identification) [4] was initially presented. The algorithm uses the constant modulus property of the transmitted signal and performs semi-blind subspace channel identification as a precursor to semi-blind equalization. The resulting receiver allows for coherent combination of the desired signal multipath, cancellation of the interfering users, removal of phase ambiguities present in blind algorithms and significant reduction of the required number of training symbols.

In spite of all the advantages of this receiver, stationary conditions frame by frame are necessary. In order to relieve this requirement, we propose in this paper a new differential BST-SBCMACI receiver (BST-SBCMACI-DIFF) that presents improved robustness to phase variations and reduction of the effects of nonlinear distortion by employing $\pi/4$ shifted DQPSK data modulation and differential detection [2].

The paper is organized as follows: The system model is presented in section II; LS optimization and semi-blind subspace channel identification in section III; SBCMACI in section IV; simulation results in section V and finally, concluding remarks in section VI.

II. SYSTEM MODEL

We consider the reverse link of an asynchronous DS-CDMA system employing complex spreading [5] and $\pi/4$-DQPSK data modulation to reduce peak-average ratio and achieve better robustness to phase variations. There are M users in the system and each user may transmit N_b data symbols per packet over assumed stationary conditions. It is also assumed that the receiver employs an antenna array consisting of A identical elements equally-spaced by $\lambda_{ant}/2$, where λ_{ant} is the carrier frequency wavelength.

Assuming that the inverse signal bandwidth is large compared to the propagation time across the array, the complex envelopes of the signals received by different antenna elements from a given path are identical except for phase and amplitude differences that depend on the path angle of arrival (AOA), array geometry and the element pattern [6].

The AOA of the lth multipath signal from the mth user is θ_m^l and $\mathbf{a}(\theta_m^l)$ is the array response vector (spatial signature vector) to the multipath signal arriving from the direction θ_m^l, with $\mathbf{a}(\theta_m^l) = \left[a_1(\theta_m^l), \cdots, a_A(\theta_m^l)\right]^T$.

We can represent the reverse link baseband complex signal in the following vector form:

$$\mathbf{r}(t) = \sum_{m=1}^{M} \sum_{k=0}^{N_b-1} \sqrt{\gamma_m} \cdot b_m(k) \cdot \mathbf{h}_m(t - kT_s) + \mathbf{v}(t) \tag{1}$$

Where $\mathbf{r}(t) = [r_1(t), \cdots, r_A(t)]^T$; γ_m is the transmitted signal power of the mth user; T_s is the symbol duration; $b_m(k) = (b^I_{m,k} + jb^Q_{m,k})/\sqrt{2}$ is the information symbol of the mth user at time k with $b^I_{m,k}, b^Q_{m,k} \in \{+1, -1\}$. Information symbols are differentially encoded by $b_m(k) = b_m(k-1) \cdot \varphi_k$ and $\varphi_k = \exp(jb^Q_{m,k} \cdot (\pi/2 - b^I_{m,k} \cdot \pi/4))$; $\mathbf{v}(t) = [v^1(t), \cdots, v^A(t)]^T$ is a complex white Gaussian noise vector with variance σ^2 and $\mathbf{h}_m(t) = [h^1_m(t), \cdots, h^A_m(t)]^T$ is the normalized complex signature waveform vector of the mth user, described by:

$$\mathbf{h}_m(t) = \sum_{n=0}^{G-1} c_m(n) \cdot \mathbf{p}_m(t - nT_c), \; 0 \le t \le T_s$$

Where; T_c is the chip duration; G is the processing gain (T_s/T_c); $c_m(n) = (\upsilon^I_{m,n} + j\upsilon^Q_{m,n})/\sqrt{2G}$ is the complex signature sequence (or spreading code) of the mth user at time n with $\upsilon^I_{m,n}, \upsilon^Q_{m,n} \in \{+1, -1\}$ and $\mathbf{p}_m(t) = [p^1_m(t), \cdots, p^A_m(t)]^T$ is the chip waveform vector of the mth user that has been filtered at the transmitter and receiver and distorted by the multipath channel. We can model $\mathbf{p}_m(t)$ as:

$$\mathbf{p}_m(t) = \sum_{l=0}^{L_m-1} \beta^l_m \cdot \mathbf{a}(\theta^l_m) \cdot \psi(t - \tau^l_m)$$

Where $L_m = \lceil T_m/T_c \rceil$ is the number of resolvable multipath components for the mth user and T_m is the delay spread experienced by the mth user; β^l_m and τ^l_m are the complex gain and time delay of the lth path of the mth user and $\psi(t)$ is the filtered chip waveform, which includes the effect of the transmitter and receiver filter.

Finally, sampling the received signal at the chip rate and assuming $T_c = 1$, we obtain the following discrete-time signal:

$$\mathbf{r}(n) = \sum_{m=1}^{M} \sqrt{\gamma_m} \sum_{k=0}^{N_b-1} b_m(k) \cdot \mathbf{h}_m(n - kG) + \mathbf{v}(n) \tag{2}$$

Considering that the receiver is in perfect synchronization with the first multipath component, l_m, of the desired user m ($\tau^{l_m}_m = 0$) and that each of the A stacked impulse responses of $\mathbf{p}_m(n)$ is FIR with order q_m such that $L_m \le q_m \le (L-1) \cdot G$ for some integer L, the discrete-time received signal corresponding to the kth symbol can be written as:

$$\mathbf{r}_s(k) = [\mathbf{H}(L-1), \cdots, \mathbf{H}(0)] \cdot [\mathbf{b}(k-L+1)^T, \cdots, \mathbf{b}(k)^T]^T + \mathbf{v}_s(k) \tag{3}$$

Where

$\mathbf{r}_s(k) = [\mathbf{r}(k \cdot G)^T, \cdots, \mathbf{r}((k+1) \cdot G - 1)^T]^T$;

$$\mathbf{H}(l) = \begin{bmatrix} \mathbf{h}_1(l \cdot G) & \cdots & \mathbf{h}_M(l \cdot G) \\ \vdots & & \vdots \\ \mathbf{h}_1((l+1) \cdot G - 1) & \cdots & \mathbf{h}_M((l+1) \cdot G - 1) \end{bmatrix};$$

$\mathbf{b}(k) = [b_1(k), \cdots, b_M(k)]^T$;

$\mathbf{v}_s(k) = [\mathbf{v}(k \cdot G)^T, \cdots, \mathbf{v}((k+1) \cdot G - 1)^T]^T$

In some cases, depending on the channel length, number of users (M), processing Gain (G) and number of antenna elements (A), it might be necessary to process more than one received vector at a time in order to estimate the kth symbol [4]. Stacking μ consecutive symbols ($\mu \ge 2$), we can define the vector that will be processed, $\mathbf{r}_\mu(k)$, as:

$$\mathbf{r}_\mu(k) = \mathbf{H}_\mu \cdot \mathbf{b}_\mu(k) + \mathbf{v}_\mu(k); \tag{4}$$

Where

$\mathbf{r}_\mu(k) = [\mathbf{r}_s(k)^T, \cdots, \mathbf{r}_s(k+\mu-1)^T]^T$;

$\mathbf{b}_\mu(k) = [\mathbf{b}(k)^T, \cdots, \mathbf{b}(k+\mu-1)^T]^T$;

$\mathbf{v}_\mu(k) = [\mathbf{v}_s(k)^T, \cdots, \mathbf{v}_s(k+\mu-1)^T]^T$;

$$\mathbf{H}_\mu = \begin{bmatrix} \mathbf{H}(L-1) & \cdots & \mathbf{H}(0) & \cdots & 0 \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ 0 & \cdots & \mathbf{H}(L-1) & \cdots & \mathbf{H}(0) \end{bmatrix}$$

III. LS OPTIMIZATION AND SEMI-BLIND SUBSPACE CHANNEL IDENTIFICATION

We can obtain an optimum spatial-temporal weight vector, $\hat{\mathbf{w}}_m$, in the LS sense that provides appropriate beam pattern to the desired user (user m) by [7]:

$$\hat{\mathbf{w}}_m = \underbrace{\left(\frac{1}{N_t} \sum_{k=0}^{N_t-1} \mathbf{r}_\mu(k) \cdot \mathbf{r}^H_\mu(k) \right)^{-1}}_{\hat{\mathbf{R}}_{N_t}} \cdot \underbrace{\left(\frac{1}{N_t} \sum_{k=0}^{N_t-1} b^*_m(k) \cdot \mathbf{r}_\mu(k) \right)}_{\hat{\mathbf{P}}^m_{N_t}} \tag{5}$$

Where $\hat{\mathbf{R}}_{N_t}$ and $\hat{\mathbf{P}}^m_{N_t}$ are the estimated autocorrelation matrix and crosscorrelation matrix using N_t training symbols, respectively. It is also possible to determine $\hat{\mathbf{P}}^m_{N_t}$ by performing channel identification. This procedure allows to work at the chip level, increasing the amount of available training data and the estimation accuracy.

As shown in [8], it is possible to perform channel identification based on eigendecomposition of the estimated autocorrelation matrix using N_b symbols, $\hat{\mathbf{R}}_{N_b}$, as follows:

$$\hat{\mathbf{R}}_{N_b} = \frac{1}{N_b} \sum_{k=0}^{N_b-1} \mathbf{r}_\mu(k) \cdot \mathbf{r}^H_\mu(k) = [\hat{\mathbf{U}}_s \;\; \hat{\mathbf{U}}_n] \begin{bmatrix} \hat{\mathbf{\Lambda}}_s & \\ & \hat{\mathbf{\Lambda}}_n \end{bmatrix} \cdot [\hat{\mathbf{U}}_s \;\; \hat{\mathbf{U}}_n]^H \tag{6}$$

Where $\hat{\mathbf{U}}_s = [\hat{\mathbf{u}}_1, \cdots, \hat{\mathbf{u}}_\xi]$, $\hat{\mathbf{U}}_n = [\hat{\mathbf{u}}_{\xi+1}, \cdots, \hat{\mathbf{u}}_{AG\mu}]$, $\hat{\mathbf{\Lambda}}_s$ and $\hat{\mathbf{\Lambda}}_n$ contain the estimated signal space eigenvectors, the estimated noise space eigenvectors and the corresponding eigenvalues for the signal and noise space vectors, respectively. ξ is the dimensionality of the signal space (rank of $\mathbf{H}_\mu$) which can be estimated using the MDL (minimum description length) criterion [9].

In DS-CDMA systems, we can use the spreading code for the desired user to perform channel classification. In [4], a semi-blind channel identification based on the following semi-blind regularized LS optimization was presented:

$$\hat{\hat{\mathbf{p}}}_m = \arg\min_{\mathbf{p}} \underbrace{\frac{1}{AGN_t}\left\|\mathbf{r}_{N_t} - \mathbf{X}_{N_t}^m \mathbf{p}\right\|^2}_{Training-based} + \alpha \cdot \underbrace{\left(\mathbf{p}^H \mathbf{\Pi}_m \mathbf{p}\right)}_{Blind-based} \tag{7}$$

Where α is some positive constant [4];

$\hat{\hat{\mathbf{p}}}_m = \left[\mathbf{p}_m(0)^T, \cdots, \mathbf{p}_m(q_m)^T\right]^T$;

$\mathbf{r}_{N_t} = \left[\mathbf{r}(0)^T, \cdots, \mathbf{r}((N_t-1)G)^T\right]^T$;

$\mathbf{X}_{N_t}^m = \left[\tilde{\mathbf{X}}_{N_t}^m \otimes I_A\right]_{AGN_t \, x \, A(q_m+1)}$; $\mathbf{I}_{A\,(A\,x\,A)}$ is the identity matrix;

$$\tilde{\mathbf{X}}_{N_t}^m = \begin{bmatrix} x_m(0) & 0 & \cdots & 0 \\ \vdots & x_m(0) & \cdots & \vdots \\ \vdots & \vdots & \ddots & \vdots \\ x_m((N_t-1)G) & \vdots & \cdots & x_m((N_t-1)G - q_m) \end{bmatrix};$$

$$x_m(n) = \sum_{k=0}^{N_b-1} b_m(k) \cdot c_m(n-kG);$$

$$\mathbf{\Pi}_m = \mathbf{C}_m^H \cdot \tilde{\mathbf{\Xi}}^H \cdot \tilde{\mathbf{\Xi}} \cdot \mathbf{C}_m; \; \mathbf{C}_m = \left[\tilde{\mathbf{C}}_m \otimes \mathbf{I}_A\right]_{ALG \, x \, A(q_m+1)};$$

$$\tilde{\mathbf{C}}_m = \begin{bmatrix} c_m(0) & & & 0 \\ \vdots & c_m(0) & & \\ c_m(G-1) & \vdots & \ddots & \\ & c_m(G-1) & & c_m(0) \\ & & \ddots & \vdots \\ & & & c_m(G-1) \\ 0 & \cdots & \cdots & 0 \end{bmatrix}_{LG \, x(q_m+1)};$$

$$\tilde{\mathbf{\Xi}} = \begin{bmatrix} \hat{\mathbf{\Xi}}_\mu & & \\ \vdots & \ddots & \hat{\mathbf{\Xi}}_\mu \\ \hat{\mathbf{\Xi}}_1 & & \vdots \\ & \ddots & \hat{\mathbf{\Xi}}_1 \end{bmatrix}_{(AG\mu-\xi)(\mu+L-1)\,x\,AGL}; \; \hat{\mathbf{U}}_n^H = \left[\hat{\mathbf{\Xi}}_1, \cdots, \hat{\mathbf{\Xi}}_\mu\right]_{(AG\mu-\xi)\,x\,AG\mu};$$

Whose solution is given by [8]:

$$\hat{\hat{\mathbf{p}}}_m = \left(\frac{1}{AGN_t}\mathbf{X}_{N_t}^{mH}\mathbf{X}_{N_t}^m + \alpha\mathbf{\Pi}_m\right)^{-1} \cdot \left(\frac{1}{AGN_t}\mathbf{X}_{N_t}^{mH}\mathbf{r}_{N_t}\right) \tag{8}$$

IV. SEMI–BLIND CONSTANT MODULUS ALGORITHM WITH CHANNEL ESTIMATION

As presented in [4], SBCMACI first computes the subspace space-time beamforming weight vector for the *m*th user by:

$$\hat{\mathbf{w}}_{m,sub} = \hat{\mathbf{U}}_s \cdot \hat{\mathbf{\Lambda}}_s^{-1} \cdot \hat{\mathbf{U}}_s^H \cdot \mathbf{C}_m \cdot \hat{\hat{\mathbf{p}}}_m = \mathbf{\Gamma} \cdot \mathbf{C}_m \cdot \hat{\hat{\mathbf{p}}}_m \tag{9}$$

And then performs the following semi-blind regularized LS iterative procedure:

i. Initialize $\mathbf{w}_m^0 = \hat{\mathbf{w}}_{m,sub}$ (10)

ii. Generate $\tilde{b}_m^{(i)}$, a sequence that contains the N_t training symbols and the $N_b - N_t$ estimated data symbols:

$$\tilde{b}_m^{(i)} = \left\{ b_m(0), \cdots, b_m(N_t-1), \frac{\mathbf{w}_m^{(i)H} \cdot \mathbf{r}_\mu(N_t)}{\left|\mathbf{w}_m^{(i)H} \cdot \mathbf{r}_\mu(N_t)\right|}, \cdots \right.$$
$$\left. \cdots, \frac{\mathbf{w}_m^{(i)H} \cdot \mathbf{r}_\mu(N_b-1)}{\left|\mathbf{w}_m^{(i)H} \cdot \mathbf{r}_\mu(N_b-1)\right|} \right\} \tag{11}$$

iii. Compute

$$\mathbf{w}_m^{(i+1)} = \mathbf{\Gamma} \cdot \underbrace{\left(\frac{1}{N_b}\sum_{k=0}^{N_b-1} \tilde{b}_m^{(i)*}(k) \cdot \mathbf{r}_\mu(k)\right)}_{\tilde{\mathbf{P}}_{N_b}^{m\,(i)}} = \mathbf{\Gamma} \cdot \tilde{\mathbf{P}}_{N_b}^{m\,(i)} \tag{12}$$

iv. Determine $\varepsilon(i) = \left\|\mathbf{w}_m^{(i+1)} - \mathbf{w}_m^{(i)}\right\|^2 / \left\|\mathbf{w}_m^{(i)}\right\|^2$ (13)

v. Repeat ii. to iv until $\varepsilon(i) < \varepsilon_w$

Where ε_w is some small positive constant.

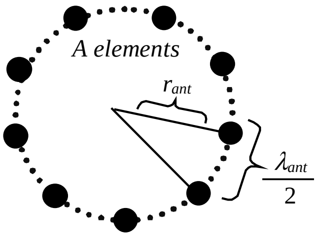
V. SIMULATION RESULTS

In this section, we evaluate the performance of the BST-SBCMACI-DIFF receiver varying the number of training symbols, SNR and number of users and we compare the results with that one obtained by using the BST-SBCMACI receiver presented in [3].

For the simulations, we consider an asynchronous DS-CDMA system with complex spreading operating at 2GHz. There are 8 users (*M*=8) per cell, each one transmitting frames with 200 symbols (N_b=200). The chip rate is 3.84 Mcps, the symbol rate is 256 Ksps and the processing gain is 15 (*G*=15). The spreading sequences are Gold-like and they are normalized to unit energy. The cell ratio is 500 *m*, and the users are randomly positioned around the cell between 50 *m* and 500 *m* ($50 \le d_m \le 500$) and with angles between -180° and 180°.

We assume for all the users that the propagation channel has 4 resolvable paths ($L_m = 4$) with a maximum angle spread of 60^o and a maximum propagation delay of 3.33 μs (~13 chip).

The base station employs a circular array antenna (see Fig.1) with 9 equally spaced elements ($\lambda_{ant}/2$).



$$\mathbf{a}\left(\theta_m^l\right) = \left[\; e^{-j\frac{2\pi}{\lambda_{ant}} r_{ant}\cos\left(\theta_m^l\right)}, \ldots, e^{-j\frac{2\pi}{\lambda_{ant}} r_{ant}\cos\left(\theta_m^l - \frac{2\pi(A-1)}{A}\right)} \;\right]^T$$

Fig.1. Circular Antenna Array

Spreading sequences, multipath delays, angle spread, complex gains and AOA of the MAI are randomly generated at each frame. Spreading sequences are normalized to unit energy. The results are obtained computing 2500 frames (N_{fr}=2500) per evaluated parameter (N_t, SNR and M). For all simulations, we consider μ=1, L=2, α=3 and $\varepsilon_w = 10^{-5}$. The signal space is estimated by using MDL method [9].

Simulations are performed for two different scenarios. The first, considered in Fig.2 to Fig.6, employs a time invariant frequency selective channel whose channel coefficients for each frame are Rayleigh distributed. The second, considered in Fig.7 to Fig.9, employs a time variant frequency selective channel, whose channel coefficients for each sample are obtained by the channel model presented in [10].

In Fig.2, the performance of BST-SBCMACI varying the number of training symbols is compared against the performance of BST-SBCMACI-DIFF for M=8 and SNR=6dB. The results show that the differential scheme requires more training symbols for a given BER in a time invariant frequency selective channel.

In Fig.3, we compare the performance of BST-SBCMACI against BST-SBCMACI-DIFF varying the SNR for M=8 and N_t=10. As we mentioned before, the robustness of differential schemes is obtained at the expense of a higher SNR.

In Fig.4 and Fig.5, we show the constellation diagrams of the first evaluated frame of BST-SBCMACI-DIFF and BST-SBCMACI, respectively, for M=8, SNR=6dB and N_t=10.

In Fig.6, comparison between BST-SBCMACI and BST-SBCMACI-DIFF receivers varying the number of users is presented. We consider SNR=6dB and N_t=10. The results show that the system capacity using BST-SBCMACI-DIFF is lower than that one obtained by using BST-SBCMACI in a time invariant frequency selective channel.

In Fig.7 and Fig.8, we show the performance of the BST-SBCMACI and BST-SBCMACI-DIFF receivers varying the SNR for a time variant frequency selective channel [10] with three different Doppler spreads (Δ_{Dopp}=50, Δ_{Dopp}=100 and Δ_{Dopp}=150), respectively.

Finally, comparison between BST-SBCMACI and BST-SBCMACI-DIFF receivers varying the SNR for a time variant frequency selective channel [10] with two different Doppler spreads (Δ_{Dopp}=50 and Δ_{Dopp}=150) is presented in Fig.9. The results show that BST-SBCMACI-DIFF outperforms BST-SBCMACI as SNR and Doppler spread increases.

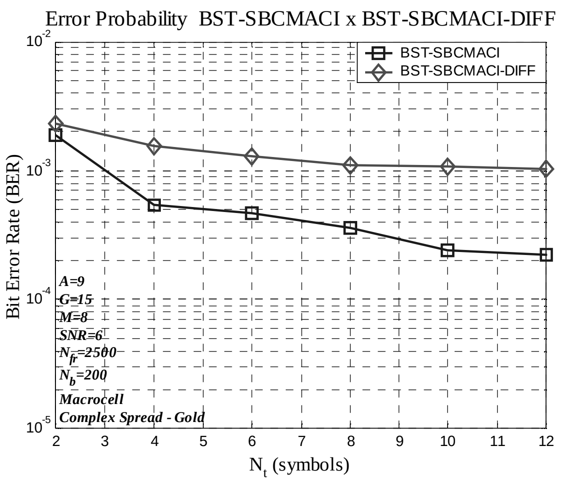


Fig.2. BER of BST-SBCMACI and BST-SBCMACI-DIFF varying the number of training symbols (SNR= 6dB, M=8 and A=9)

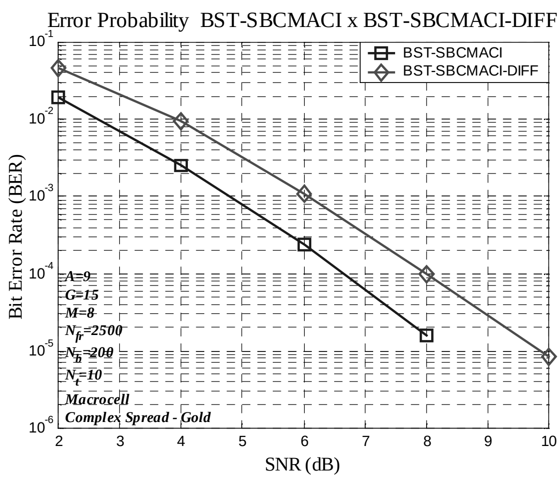


Fig.3. BER of BST-SBCMACI and BST-SBCMACI-DIFF varying SNR (M=8, N_t=10 and A=9)

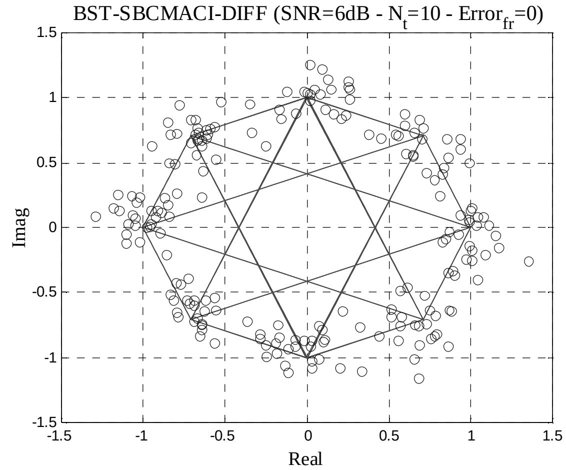


Fig.4. Constellation Diagram for BST-SBCMACI-DIFF (SNR=6dB, M=8, N_t= 10 and A=9)

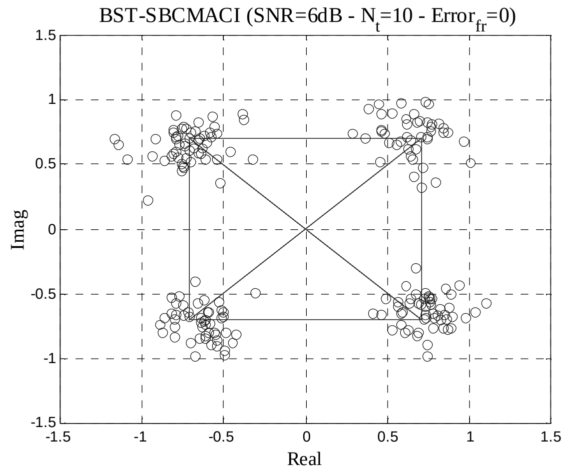


Fig.5. Constellation Diagram for BST-SBCMACI with (SNR=6dB, M=8, N_t= 10 and A=9)

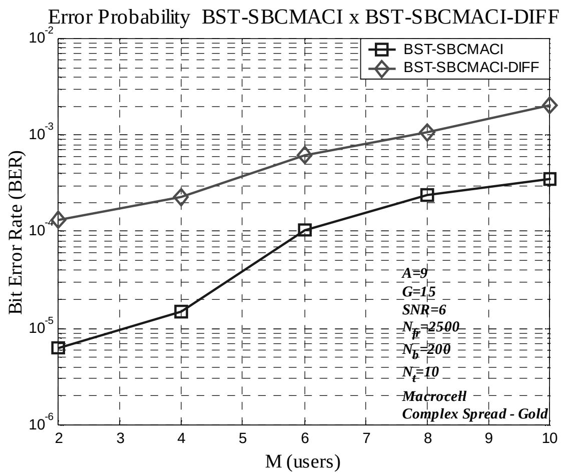


Fig.6. BER of BST-SBCMACI and BST-SBCMACI-DIFF varying the number of users (SNR= 6dB, N_t=10 and A=9)

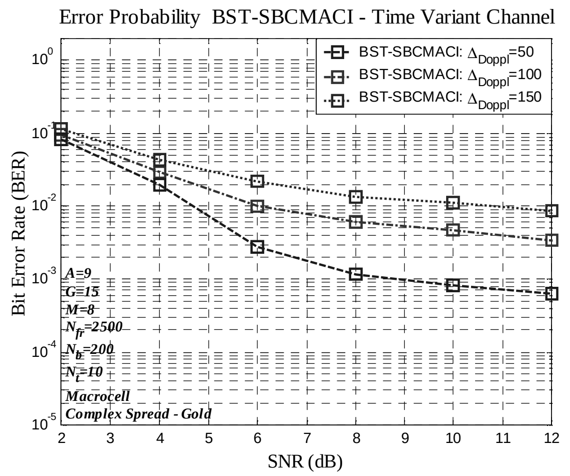


Fig.7. BER of BST-SBCMACI varying SNR for a time variant channel with three different Doppler spreads (M=8, N_t=10 and A=9)

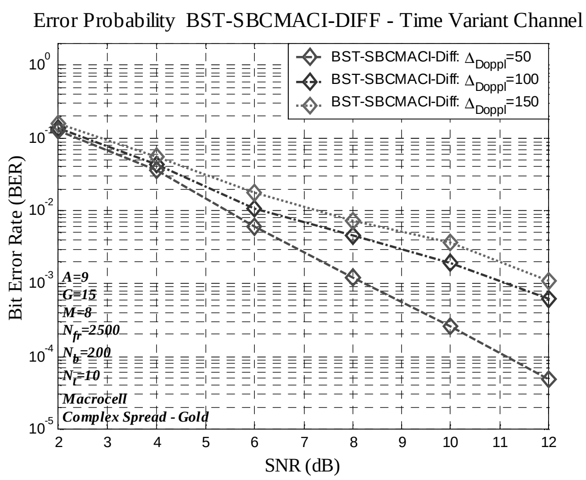


Fig.8. BER of BST-SBCMACI-DIFF varying SNR for a time variant channel with three different Doppler spreads (M=8, N_t=10 and A=9)

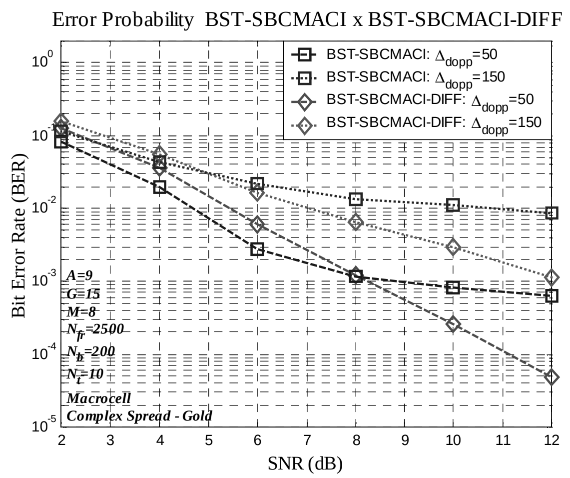


Fig.9. Comparison of BST-SBCMACI and BST-SBCMACI-DIFF varying SNR for a time variant channel with Δ_{Dopp} =50 and Δ_{Dopp} =150 (M=8, N_t=10 and A=9)

VI. CONCLUSIONS

In this paper, we have presented a new differential semi-blind spatial-temporal beamforming receiver for high data rate DS-WCDMA systems. The presented receiver is based on the semi-blind spatial-temporal beamforming receiver proposed in [3] and presents improved robustness to phase variations presented in time varying fading channels and reduction of the effects of nonlinear distortion.

Simulations were performed considering an asynchronous DS-WCDMA system employing complex spreading and a circular antenna array in time invariant and time variant frequency selective multipath fading channels.

The obtained results for time invariant channels show that BST-SBCMACI-DIFF requires higher SNR to reach the same performance figures of BST-SBCMACI. For time variant channels, BST-SBCMACI-DIFF outperform BST-SBCMACI as SNR and Doppler spread increases.

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