New decoder for Space-Time Block Coded Multiuser System over Flat and Time-Selective Fading Channels

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***Abstract*—The conventional zero forcing (ZF) and minimum mean squared error (MMSE) decoders are constructed based on the assumption that the channels are flat fading (FF) and quasi-static fading (QSF). The two decoders exhibit the inter symbol interference (ISI) once the channels are time-selective fading (TSF) when decoding one user's signals. In this paper, we propose a new decoder design scheme applicable to the existing transceiver structure for space-time block coded (STBC) multi-user (MU) communication system over flat fading (FF) and time selective fading (TSF) channels. we analyze the impacts of the channels time variations on the two known ZF and MMSE decoders, then, the analytical bit error rate (BER) expressions of these decoders are presented and the results through computer simulation indicate the significant performance gains of our new decoder compared with the conventional ZF and MMSE decoders especially at high SNR for various values of α.**

I. INTRODUCTION

Space-time block code (STBC) has been shown to enable diversity gain in wireless systems[1] thus leading to performance improvement in fading channels. In multi-user (MU) communications scenario, multi-user interference (MUI) is the major challenge that limits the capacity as well as the data rates of the system. The first works that deal with MUI cancellation for STBC systems are presented in [2]-[3]. A scheme has been outlined for multi-user detection (MUD) in which the Alamouti STBC (ASTBC) is used at the transmitters. The zero-forcing (ZF) and minimum mean-squared error (MMSE) interference suppression techniques that exploit the structure of the STBC are developed in [2]-[3] and the simulation results show that both MUI suppression as well as diversity advantage is achieved for both users.

The important assumption in the known schemes [2]-[3] is that the channels are both flat fading (FF) and quasi-static fading (QSF). Based on the QSF assumption, the known ZF and MMSE decoders will induce to inter symbol interference (ISI) if decoding one user's signals when the channels vary from one symbol duration to another, i.e., time selective fading (TSF). The ZF and MMSE decoders, as in [2]-[3] are just fit for two users scenario. Based on the users' grouping at the transmitter and group filtering idea at the receiver, T. A. Tran *et al.* proposed a novel transceiver for space-time block coded multi-user communication system, as shown in [4]. There, each group is assigned one group signature (GS), all users in that group use the same GS as their signature code. The inter group interference (IGI) is then completely removed by the group filtering (GF) at the receiver. Following [5], the author applys the iterative soft-input soft-output (SISO) MUD and the corresponding channel decoders to detect and decode the transmitted signals of each user in each group. The channels considered in [4] are QSF. Simultaneously, the more iterative operations at the receiver are required to approach the approximate performance of single user system.

In this paper, we exploit the transceiver structure[4] for STBC MU communication system. The users are grouped and the users in one group use the same GS. At the receiver, we apply the GF to cancel IGI and obtain the IGI-free interested group. Instead of using SISO MUD to decode the transmitted signals of each user in the group, we propose a new decoder applicable to FF and TSF channels. Finally, the analytical bit error rate (BER) expressions are presented for these decoders and the results through Monte Carlo simulation validate the robustness of our new decoder opposed to the channels time variations.

The outline of this paper is as follows. In section II, system model is given. Section III describes the IGI cancellation scheme. We propose the new decoder design in section IV. Analytical BER of these decoders is derived and the numerical results are presented in section V. Finally, conclusions are drawn in section VI. In this paper, lowercase bold typeface letters (e.g., $\boldsymbol{x}$) represent vectors, uppercase, bold typeface letters (e.g., $\boldsymbol{X}$) represent matrices, and $\boldsymbol{I}_m$ denotes an $m\times m$ identity matrix. Superscripts $(\bullet)^*$, $(\bullet)^T$, and $(\bullet)^\dagger$ denote conjugate, and transpose, conjugate transpose, respectively. The $diag(x)$ stands for a diagonal matrix with x on its diagonal, $\boldsymbol{0}_{k\times p}$ denotes an all-zero matrix of size $k\times p$; $[\bullet]_p$ denotes the p^{th} entry of a vector;

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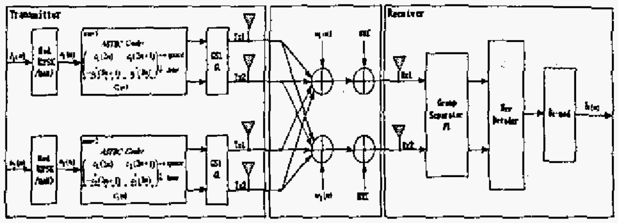

Figure 1. System model.

$[\bullet]_{p,q}$ denotes the $(p,q)^{th}$ entry of a matrix; and, $[\bullet]_{p\times q}$ denotes the top left corner sub-matrix of size $p \times q$.

II. SYSTEM MODEL

Fig. 1 shows the system model, based on the new proposed decoder structure. For the sake of exposition, we consider a four-user STBC system where each user employs 2 transmit antennas and the receiver employs 2 receive antennas. We point out that the idea can be applied to a system having an arbitrary number of users. We divide four users into two groups equally. In Fig. 1, we just describe the first group and there are two users in the group. The information bit stream $b_i(n)$ for user i, $i=1\bullet 4$ is modulated and generates the complex sequence $\{c_i(n)\}$ which has the same energy E_s and symbol duration T_s. The complex sequence $\{c_i(n)\}$ is inputted to the ASTBC coder and the code matrix of STBC of the i^{th} user $\boldsymbol{C}_i(n)$, $i=1\bullet 4$ is given by[1]

$$\boldsymbol{C}_i(n)=\begin{pmatrix} c_i(2n) & c_i(2n+1) \\ -c_i^*(2n+1) & c_i^*(2n) \end{pmatrix}. \tag{1}$$

The output of the ASTBC coder is fed into the GS encoding block before being transmitted over the channel.

The channel model considered in this work is frequency-flat time-selective fading, namely FF and TSF channel. Following [6], we adopt the first order Makovian model for modeling the TSF channel. Nevertheless, the new decoder proposed in this work is valid for various TSF channel models. Yet, we choose this model for the sake of performance evaluation. Let $h_{jl}^i(n)$ denote the complex channel gain from l^{th} transmit antenna of the i^{th} user to j^{th} receive antenna at time instant n, $l,j=1\bullet 2$, $i=1\bullet 4$, where $h_{jl}^i(n)$ are modeled as *i.i.d.* circularly complex Gaussian random variables with zero mean and variance $\sigma_h^2=0.5$ per dimension. We assume that the all users experience the same channels time variations, so, the channel gains vary from one symbol duration to another as

$$h_{jl}^i(n)=\alpha h_{jl}^i(n-1)+v_{jl}^i(n), \quad l,j=1\bullet 2, i=1\bullet 4. \tag{2}$$

In (2), the parameter α accounts for the Doppler shift of the channel. The $v_{jl}^i(n)$ are complex additive white Gaussian noise (AWGN) samples with zero mean and variance σ_v^2 and are independent of $h_{jl}^i(n)$. From (2), we have

$$|\alpha|^2+\sigma_v^2=\sigma_h^2=1. \tag{3}$$

The value of α is dependent on the Doppler shift of the channel.

Let $\boldsymbol{r}^j(n)$ and $\boldsymbol{w}^j(n)$ denote the received signal and noise vectors at the j^{th} receive antenna at time instant n. Let $\boldsymbol{G}_g$ and $\boldsymbol{F}_g$ denote the GS and GF of the g^{th} group, $g=1\bullet 2$. The received signal at each receive antenna is a noisy superposition of the signals transmitted from four users. We assume that the noise is complex AWGN with zero mean and variance $0.5\sigma^2$ per dimension. The received signal vectors at the first receive antenna at time instants n and $n+1$ are then given by

$$\begin{aligned}\boldsymbol{r}^1(n)=&\,\boldsymbol{G}_1\sum_{i=1}^{2}\sum_{l=1}^{2}h_{1l}^i(2n)c_i(2n+l-1)\\ &+\boldsymbol{G}_2\sum_{i=3}^{4}\sum_{l=1}^{2}h_{1l}^i(2n)c_i(2n+l-1)+\boldsymbol{w}^1(n)\end{aligned} \tag{4}$$

$$\begin{aligned}\boldsymbol{r}^1(n+1)=&\,\boldsymbol{G}_1\sum_{i=1}^{2}\sum_{l=1}^{2}(-1)^l h_{1l}^i(2n+1)c_i^*(2n+2-l)\\ &+\boldsymbol{G}_2\sum_{i=3}^{4}\sum_{l=1}^{2}(-1)^l h_{1l}^i(2n+1)c_i^*(2n+2-l)+\boldsymbol{w}^1(n+1).\end{aligned} \tag{5}$$

The received signal vectors at the second receive antenna are similarly given as

$$\begin{aligned}\boldsymbol{r}^2(n)=&\,\boldsymbol{G}_1\sum_{i=1}^{2}\sum_{l=1}^{2}h_{2l}^i(2n)c_i(2n+l-1)\\ &+\boldsymbol{G}_2\sum_{i=3}^{4}\sum_{l=1}^{2}h_{2l}^i(2n)c_i(2n+l-1)+\boldsymbol{w}^2(n)\end{aligned} \tag{6}$$

$$\begin{aligned}\boldsymbol{r}^2(n+1)=&\,\boldsymbol{G}_1\sum_{i=1}^{2}\sum_{l=1}^{2}(-1)^l h_{2l}^i(2n+1)c_i^*(2n+2-l)\\ &+\boldsymbol{G}_2\sum_{i=3}^{4}\sum_{l=1}^{2}(-1)^l h_{2l}^i(2n+1)c_i^*(2n+2-l)+\boldsymbol{w}^2(n+1).\end{aligned} \tag{7}$$

In (4)-(7), the received signal vectors and the noise vectors have the same dimensions as the group signature $\boldsymbol{G}_g$, $g=1\bullet 2$. Relative to the signals belonging to group $\boldsymbol{G}_1$, the signals belonging to group $\boldsymbol{G}_2$ are MUI and vice versa. In order to obtain the signals from the interested group, we introduce the IGI cancellation scheme next.

III. IGI CANCELLATION SCHEME

At the receiver, each received signal vector is filtered by the respective GF to give the signals transmitted by the interested group and deterministically cancel the IGI from other groups, as shown in [4]. Let $r_g^j(n)$ and $r_g^j(n+1)$ be the outputs of the filter of the g^{th} group, $g = 1 \bullet 2$, at the j^{th} receive antenna, $j = 1 \bullet 2$, at time instants n and $n+1$, respectively. Now, we assume that the GSs and GFs satisfy the following condition

$$F_m \cdot G_g = \delta_{gm} \quad g,m = 1 \bullet 2. \tag{8}$$

In (8), δ_{gm} is the Kronecker's delta function given by

$$\begin{cases} \delta_{gm} = 0 & \text{if } g \neq m \\ \delta_{gm} = 1 & \text{if } g = m. \end{cases} \tag{9}$$

The filtered signals $r_g^j(n)$, $r_g^j(n+1)$ and noise signals $w_g^j(n)$, $w_g^j(n+1)$ are defined as

$$\begin{cases} r_g^j(n) = F_g \cdot r^j(n) \\ r_g^j(n+1) = F_g \cdot r^j(n+1) \end{cases} \tag{10}$$

$$\begin{cases} w_g^j(n) = F_g \cdot w^j(n) \\ w_g^j(n+1) = F_g \cdot w^j(n+1). \end{cases} \tag{11}$$

Also, let us define the following vectors

$$\begin{cases} r_1(n) = [r_1^1(n) \;\; r_1^1(n+1)^* \;\; r_1^2(n) \;\; r_1^2(n+1)^*]^T \\ r_2(n) = [r_2^1(n) \;\; r_2^1(n+1)^* \;\; r_2^2(n) \;\; r_2^2(n+1)^*]^T \end{cases} \tag{12}$$

$$\begin{cases} w_1(n) = [w_1^1(n) \;\; w_1^1(n+1)^* \;\; w_1^2(n) \;\; w_1^2(n+1)^*]^T \\ w_2(n) = [w_2^1(n) \;\; w_2^1(n+1)^* \;\; w_2^2(n) \;\; w_2^2(n+1)^*]^T. \end{cases} \tag{13}$$

In (12)-(13), $r_i(n)$ and $w_i(n)$ are the redefined signal and noise vectors respectively, whose elements are defined in (10)-(11). Let $H_{i,j}$ denote the modified channel matrix from the all transmit antennas to the j^{th} receive antenna of user i, and is given by

$$H_{i,j} = \begin{pmatrix} h_{j1}^i(2n) & h_{j2}^i(2n) \\ h_{j2}^{i*}(2n+1) & -h_{j1}^{i*}(2n+1) \end{pmatrix}. \tag{14}$$

Using the received signal structures in (4)-(7), the condition (8), and definitions (10)-(13) we obtain

$$r_1(n) = \underbrace{\begin{pmatrix} H_{1,1} & H_{2,1} \\ H_{1,2} & H_{2,2} \end{pmatrix}}_{H} \underbrace{\begin{pmatrix} c_1(n) \\ c_2(n) \end{pmatrix}}_{c} + w_1(n) \tag{15}$$

$$r_2(n) = \begin{pmatrix} H_{3,1} & H_{4,1} \\ H_{3,2} & H_{4,2} \end{pmatrix} \begin{pmatrix} c_3(n) \\ c_4(n) \end{pmatrix} + w_2(n). \tag{16}$$

The $c_i(n)$ in (15)-(16) is the code vector of the i^{th}-user, $i = 1 \bullet 4$, and it is defined as $c_i(n) = [c_i(2n) \;\; c_i(2n+1)]^T$. Examining (15)-(16) we can easily see that the filtered signal vector of the first group does not contain signals from the second group and vice versa. In order to obtain the IGI-free signals of each group, the designs of GSs and GFs in this paper have been constructed as the same in [4], it is important that the designs of GSs and GFs in [4] do not amplify or alter the noise power after the group filtering.

IV. NEW DECODER DESIGN

From (15)-(16) we have obtained the IGI-free signals of each group. Due to the symmetry in (15)-(16), let us assume that we are interested in decoding signals of the first user from the first group. For the ZF and MMSE decoders, it will induce to ISI over TSF channels. So, we firstly analyze the impacts of the channels time variations on the two decoders, then, propose a new decoder applicable to TSF channels based on the ZF idea. The new decoder removes the ISI completely and the latter simulations validate the robustness of our new decoder.

A. Analysis of the Impacts of the Channel Time Variations on the ZF and MMSE Decoders

For the ZF decoder, a clever way of recovering the transmitted symbol while ensuring space-time diversity gains has been developed in [2]. Let $A = H_{1,1} - H_{2,1} H_{2,2}^{-1} H_{1,2}$, since A are nonorthogonal over TSF channels, so, $A^{\dagger} A$ are nondiagonal. It is obvious that the nondiagonal elements of matrix $A^{\dagger} A$ are the ISI relative to the estimation of $c_1(2n)$ and $c_1(2n+1)$.

For the MMSE decoder, since the coefficients β_i, $i = 1 \bullet 2$ are nonzero over TSF channels, the formula $(\alpha_1^{\dagger} r_1(n) - \hat{c}_1(2n))$ in (24) of [3] not only includes co-channel interference (CCI) and AWGN, but also ISI component $\beta_2^* c_1(2n+1)$ for the estimation of $c_1(2n)$; similarly, the formula $(\alpha_2^{\dagger} r_1(n) - \hat{c}_1(2n+1))$ in (38) of [3] not only includes CCI and AWGN, but also ISI component $\beta_1^* c_1(2n)$ for the estimation of $c_1(2n+1)$.

The performances of the ZF and MMSE decoders degrade severely with the increase of channel time variations. The latter simulations verify that the two decoders exhibit error floors at high SNR values over TSF channels.

B. Design of the New Decoder

Our aim in the new decoder design is to completely remove the ISI. The design scheme is described as follows.

Pass the received signal vector $r_1(n)$ in (15) through a transform $\boldsymbol{\Theta}$ so that

$$y = \boldsymbol{\Theta}\cdot r_1(n) = (\boldsymbol{\Theta H})\cdot c + \boldsymbol{\Theta}\cdot w_1(n) = \boldsymbol{D}c + w_1'(n). \tag{17}$$

In (17), $\boldsymbol{D} = diag(p_1\ \ p_2\ \ p_3\ \ p_4)$. p_i's are the diagonal elements of $\boldsymbol{\Theta H}$. Note that the $\boldsymbol{H}$ is a nonorthogonal matrix over TSF channels, we define the following special transform $\boldsymbol{\Theta}$ as

$$\boldsymbol{\Theta} = \begin{pmatrix} \theta_{11} & \theta_{12} \\ \theta_{21} & \theta_{22} \end{pmatrix}. \tag{18}$$

In (18), θ_{ij} , $i,j = 1 \cdot 2$ are 2×2 complex matrices. Substituting (18) back into (17), after some straightforward manipulations, we have

$$\begin{cases} \boldsymbol{\theta}_{11} = -\boldsymbol{\theta}_{12}\boldsymbol{H}_{2,2}\boldsymbol{H}_{2,1}^{-1}, \boldsymbol{\theta}_{12} = \boldsymbol{D}_1\boldsymbol{X}_1^{-1} \\ \boldsymbol{\theta}_{21} = \boldsymbol{D}_2\boldsymbol{X}_2^{-1}, \boldsymbol{\theta}_{22} = -\boldsymbol{\theta}_{21}\boldsymbol{H}_{1,1}\boldsymbol{H}_{1,2}^{-1}. \end{cases} \tag{19}$$

In (19)

$$\begin{cases} \boldsymbol{X}_1 = \boldsymbol{H}_{1,2} - \boldsymbol{H}_{2,2}\boldsymbol{H}_{2,1}^{-1}\boldsymbol{H}_{1,1} \\ \boldsymbol{X}_2 = \boldsymbol{H}_{2,1} - \boldsymbol{H}_{1,1}\boldsymbol{H}_{1,2}^{-1}\boldsymbol{H}_{2,2} \\ \boldsymbol{D}_1 = diag(p_1\ \ p_2), \boldsymbol{D}_2 = diag(p_3\ \ p_4) \\ \boldsymbol{D} = diag(\boldsymbol{D}_1\ \ \boldsymbol{D}_2). \end{cases} \tag{20}$$

Many choices exist for the values of p_i's. In this paper, however, they have been selected as: p_i=the i^{th} diagonal element of $\boldsymbol{H}^\dagger\boldsymbol{H}$.

Thus, the modified signal vector $\boldsymbol{y}$ is as follows

$$y = \begin{pmatrix} y_1 \\ y_2 \end{pmatrix} = \boldsymbol{\Theta}\cdot r_1(n) = \begin{pmatrix} \boldsymbol{D}_1 & \boldsymbol{0}_{2\times 2} \\ \boldsymbol{0}_{2\times 2} & \boldsymbol{D}_2 \end{pmatrix}\begin{pmatrix} c_1(n) \\ c_2(n) \end{pmatrix} + \begin{pmatrix} w_1'\_1 \\ w_1'\_2 \end{pmatrix}. \tag{21}$$

So, in order to decode the first user's signals, we have

$$y_1 = \boldsymbol{D}_1 c_1(n) + w_1'\_1. \tag{22}$$

From (22), since $\boldsymbol{D}_1$ is a diagonal matrix, it is easy to obtain the estimation $\hat{c}_1(2n)$ and $\hat{c}_1(2n+1)$ of the transmitted symbols of the first user.

V. PERFORMANCE ANALYSIS AND COMPUTER SIMULATION RESULTS

A. *Performance Analysis*

It has been proved in [4] that the group separator separates the received signals into groups and does not degrade performance, in term of probability of error. Therefore, performance of our four-user system is the same as performance of a two-user system.

In this section, we provide theoretical analysis for the new decoder over TSF channels and then execute simulations. We assume QPSK modulation and the receiver perfectly knows the channel state information (CSI).

From (22), as a result of the symmetry for decoding symbol $c_1(2n)$ and $c_1(2n+1)$, the total BER for the first user is equal to the BER of decoding symbol $c_1(2n)$ or symbol $c_1(2n+1)$. So, we just present the SNR after obtaining the estimation of symbol $c_1(2n)$. Based on (22), the effective SNR for the new decoder over TSF channels is denoted by

$$SNR\_{NEW} = \frac{|p_1|^2\cdot E_s}{E[|[w_1'\_1]_1|^2]}. \tag{23}$$

In (23)

$$\begin{cases} p_1 = [\boldsymbol{H}^\dagger\boldsymbol{H}]_{1,1} \\ E[|[w_1'\_1]_1|^2] = \sigma^2\cdot[\boldsymbol{\Theta\Theta}^\dagger]_{1,1}. \end{cases} \tag{24}$$

Approximately, we can give the effective SNR for the ZF and MMSE decoder over TSF channels as follows

$$SNR\_{ZF} = \frac{|a_{11}|^2\cdot E_s}{|a_{12}|^2\cdot E_s + E[|[\tilde{w}\_1]_1|^2]}. \tag{25}$$

In (25)

$$\begin{cases} a_{11} = [A^\dagger A]_{1,1}, a_{12} = [A^\dagger A]_{1,2} \\ E[|[\tilde{w}\_1]_1|^2] = \sigma^2\cdot[A^\dagger[FF^\dagger]_{2\times 2}A]_{1,1}. \end{cases} \tag{26}$$

In (26), $\boldsymbol{F}$ is the block linear filter matrix, as shown in [2].

$$SNR\_{MMSE} = \frac{E_s}{E[\|\boldsymbol{\alpha}_1^\dagger\cdot r_1(n) - c_1(2n)\|^2]}. \tag{27}$$

In (27)

$$\begin{cases} E[\|\boldsymbol{\alpha}_1^\dagger\cdot r_1(n) - c_1(2n)\|^2] = E_s\cdot\{1 + \|\boldsymbol{\alpha}_1^\dagger\boldsymbol{H}\|^2 \\ \qquad\qquad -2\,\mathrm{Re}[\boldsymbol{\alpha}_1^\dagger\boldsymbol{H}v]\} + \sigma^2\cdot\|\boldsymbol{\alpha}_1\|^2 \\ v = [1\ \ 0\ \ 0\ \ 0]^T. \end{cases} \tag{28}$$

Let $P_{[NEW]}$, $P_{[ZF]}$ and $P_{[MMSE]}$ denote the BER of the above three decoders, so, for the QPSK modulation, we have [7, (5.2-57)]

$$\begin{cases} P_{[ZF]} = Q(\sqrt{SNR\_{ZF}}) \\ P_{[MMSE]} = Q(\sqrt{SNR\_{MMSE}}) \\ P_{[NEW]} = Q(\sqrt{SNR\_{NEW}}). \end{cases} \tag{29}$$

B. *Simulation Results*

The averaged BER given in (29) of the new decoder is numerically evaluated through Monte Carlo simulation. Fig. 2 presents the theoretical BER and the true simulated BER curves. From Fig. 2, we can observe that the analytical BER agree well with the true BER curves for various values of α, which is referred to as alpha in the figure. So, for the performance comparison, we substitute numerical evaluation for long true simulation for the following simulations.

Fig. 3 shows the performance comparison between the ZF decoder[2] and the new decoder for different α value. From Fig. 3, we can see that at $\alpha = 0.998$, BER=10^{-4}, the new decoder obtain almost 5dB gains. With the decrease of α, i.e., the increase of channels time variations, the performance gains is more significant especially at high SNR values. Fig. 3 also indicates clearly that the conventional ZF decoder soon exhibits error floors for the larger channels time variations, while the new decoder does not.

Fig. 4 shows the performance comparison between the MMSE decoder[3] and the new decoder for different α value. From Fig. 4, we can see that at $\alpha = 0.998$, the new decoder outperforms the MMSE decoder just at high SNR values, this is since the MMSE decoder considers the CCI and AWGN simultaneously, while the performance of the new decoder is the almost same as that of the ZF decoder for small channels time variations, so, it is reasonable that the MMSE decoder outperforms the new decoder especially at low SNR values. But the MMSE decoder soon exhibits error floors with the increase of channels time variations at high SNR values, while the new decoder excels the MMSE decoder in performance at high SNR values for various α values and does not exhibits error floors.

VI. CONCLUSION

We have proposed a new decoder for STBC MU communication system over TSF channels. The analytical BER expression for the new decoder has been presented and the results through theoretical and true simulation validate the robustness of our new decoder.

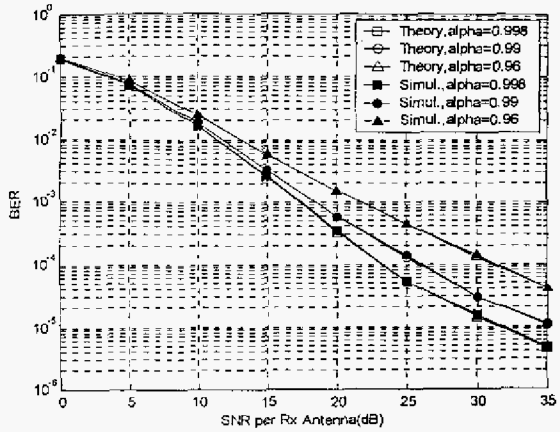

Figure 2. Theoretical BER and simulated BER performances of the new decoder for different α value.

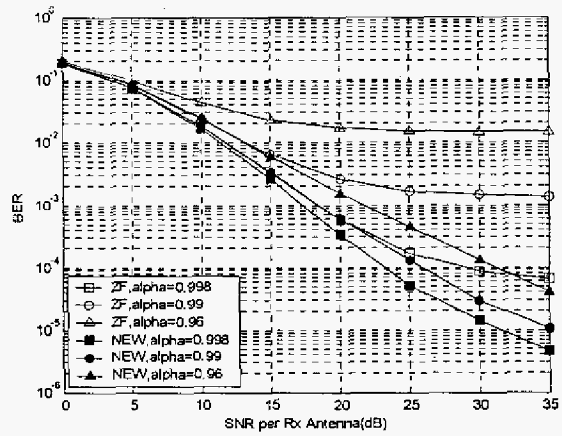

Figure 3. BER performance comparison between the ZF decoder[2] and the new decoder for different α value.

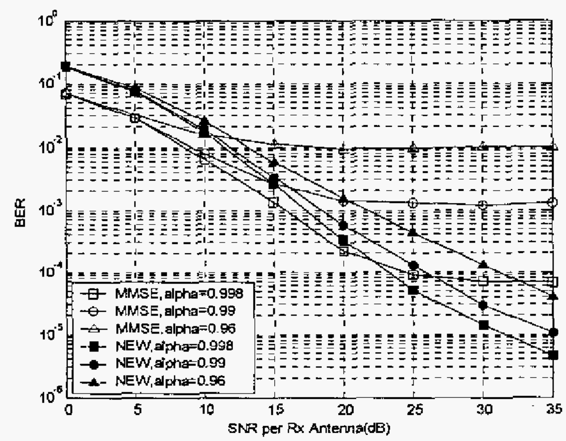

Figure 4. BER performance comparison between the MMSE decoder[3] and the new decoder for different α value.

REFERENCES

[1] S. M. Alamouti, "A simple transmit diversity technique for wireless communications," *IEEE J. Select. Areas Commun.*, vol. 16-8, pp. 1451–1458, Oct. 1998.

[2] A. F. Naguib and N. Seshadri, "Combined interference cancellation and ml decoding of space-time block codes," in *Proc. of the 7th Communications Mini-Conference at GlobeCom '98*, Sydney, 1998.

[3] A. F. Naguib, N. Seshadri and A. R. Calderbank, "Application of space-time block codes and interference suppression for high capacity and high data rate wireless systems)," in *Proc. 32nd Asilomar Conf. Signals, Systems and Computers*, 1998.

[4] T. A. Tran and A. B. Sesay, "A robust and low-complexity transceiver for space-time block coding multiuser wireless communications systems," *IEEE VTC-spring*, vol. 4, pp. 1589–1593, 2002.

[5] X. Wang and H. V. Poor, "Iterative (Turbo) soft interference cancellation and decoding for coded CDMA," *IEEE Trans. Commun.*, Vol. 47-7, pp. 1046-1061, July 1999.

[6] H. S. Wang and P. -C. Chang, "On verifying the first order Makovian assumption for Rayleigh fading channel model," *IEEE Trans. Veh. Technol.*, vol. 45-2, pp. 353–357, May 1996.

[7] J. G. Proakis, *Digital Communications*. 4th ed. New York: McGraw-Hill, 2001.