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Description	

Space-Time Coded Turbo Equalization and Multiuser Detection - Asymptotic Performance Analysis in the Presence of Unknown Interference

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***Abstract*— Turbo MIMO equalization in multiuser space-time-trellis-coded (STTrC) system is considered in this paper. The low complexity MMSE receiver with soft cancellation (SC) is proposed for joint inter-symbol-interference (ISI), known co-channel interference (CCI) and unknown co-channel interference (UCCI) suppression.**

It is shown that in the multiuser scenario without UCCI the proposed iterative receiver achieves maximum-likelihood (ML) bound on performance of the single user. The number of receive antennas is thereby equal to the number of users and not to the number of transmit antennas. Furthermore, the upper bound on pairwise error probability (PEP) is derived in the presence of UCCI in the asymptotic case of ideal SC. The result shows that the effect of the interference on the diversity and coding gain is very similar to that of the channel correlation at the receiver side. Namely the diversity and coding gain depend on the rank of the inverse of the matrix R of the UCCI.

I. Introduction

Communications signal transmission and reception using multiple transmit antennas and receive antennas over an multiple-input-multiple-output (MIMO) channel is one of the most promising approaches to increase the link capacity and achievable data rates [1]. Two key approaches have been developed to make effective use of the benefits of the MIMO channels. The first one is Bell-Labs-Layered-Space-Time-Architecture (BLAST) [2] which aims at approaching the channel outage capacity. Another one that combines the benefits of transmit diversity and channel coding is space-time-trellis-coding (STTrC) [3]. Some recent developments combine the benefits of the above two approaches [4].

To fully exploit the benefits of the broadband, frequency selective channels using single carrier communications, the cost efficient implementation of the equalization part of the receiver is a key issue. Furthermore, to fully exploit the capacity of the multipath channel turbo processing has been proposed [5], which turns the multipath channel into a set of parallel diversity channels. Recently, the MMSE-based turbo-equalization has attracted considerable attention due to the possibilities for adaptive implementation [6] and even further complexity reductions [7].

Iterative equalization with STTrC-codes has been introduced in [8], where the optimal MAP equalizer has been used. In this paper, we extend the MMSE-based turbo equalization of [6], [10], [9] to detect STTrC-coded signals. Asymptotic performance analysis in the presence of UCCI is performed in case of perfect SC.

The rest of the paper is organized as follows. Section II describes system model. Section III describes the performance evaluation in the asymptotic case of perfect SC. Section IV contains numerical results. The paper is concluded in Section V.

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II. System and Received Signal Model

Figure 1 describes the system model. Each of K users encodes bit information sequence $c_k(i)$, $k = 1, ..., K$, $i = 1, ..., Bk_0$ using a rate k_0/N_T STTrC code, where N_T and B are the number of transmit antennas and frame length in symbols, respectively. The encoded sequences $b_k(i) \in \mathcal{Q}$, $i = 1, ..., BN_T$ are first grouped in B blocks of N_T symbols, where $\mathcal{Q} = \{\alpha_1, \ldots, \alpha_{2^{k_0}}\}$ denotes the modulation alphabet assumed to be M-phase-shift-keying (M-PSK). However, it is straightforward to extend the receiver derivations to the quadrature-amplitude-modulations (QAM). The coded sequence is then interleaved so that the positions within blocks of length N_T remain unchanged but the positions of the blocks themselves are permuted within frame according to the user-specific interleaver pattern. Thereby the rank properties of the STTrC codes are preserved [11]. The interleaved sequences are then headed by the user-specific training sequences consisting of TN_T symbols. The entire frame is serial-to-parallel converted, resulting in the sequences $b_k^{(n)}(i)$, $n = 1, .., N_T$, $i = 1, ..., B + T$ and transmitted with N_T transmit antennas trough the frequency selective channel.

After coherent demodulation in the receiver, the signals from each of N_R receive antennas are sampled in time domain to capture the multipath components. Observing the signals from different transmit antennas of different users as the virtual users and arranging them in the vector form similarly as in [10], [9] we form the space-time representation of the received signal at time instant i given by

$$\mathbf{y}(i)=\underbrace{\mathbf{H}\mathbf{u}(i)}_{\text{desired}}+\underbrace{\mathbf{H}_I\mathbf{u}_I(i)}_{\text{UCCI}}+\underbrace{\mathbf{n}(i)}_{\text{noise}},\ i=1,...,T+B, \quad (1)$$

where $\mathbf{y}(i)\in\mathbb{C}^{LN_R\times 1}$ is space-time sampled received signal vector, given by

$$\mathbf{y}(i)=[\mathbf{r}^T(i+L-1),\ldots,\mathbf{r}^T(i)]^T, \quad (2)$$

with $\mathbf{r}(i)\in\mathbb{C}^{N_R\times 1}$ being

$$\mathbf{r}(i)=[r_1(i),\ldots,r_{N_R(i)}]^T. \quad (3)$$

L is the number of paths of the frequency selective channel and $r_m(i)$ denotes the signal sample obtained after matched filtering at the mth receive antenna. $\mathbf{H}\in\mathbb{C}^{LN_R\times KN_T(2L-1)}$ is channel matrix with the the form of

$$\mathbf{H}=\begin{bmatrix}\mathbf{H}(0) & \ldots & \mathbf{H}(L-1) & \ldots & \mathbf{0}\\ \vdots & \ddots & & \ddots & \vdots\\ \mathbf{0} & \ldots & \mathbf{H}(0) & \ldots & \mathbf{H}(L-1)\end{bmatrix},$$

and

$$\mathbf{H}(l)=\begin{bmatrix}h_{1,1}^{(1)}(l) & \ldots & h_{1,1}^{(N_T)}(l) & \ldots & h_{K,1}^{(N_T)}(l)\\ \vdots & \ddots & \vdots & \ddots & \vdots\\ h_{1,N_R}^{(1)}(l) & \ldots & h_{1,N_R}^{(N_T)}(l) & \ldots & h_{K,N_R}^{(N_T)}(l)\end{bmatrix},$$

where $h_{k,m}^{(n)}(l)$ denotes the l-th path complex gain between kth user's nth transmit antenna and mth receive antenna. The vector $\mathbf{u}(i)\in\mathbb{Q}^{KN_T(2L-1)\times 1}$ denotes desired users' sequence, and it is defined as

$$\mathbf{u}(i)=[\mathbf{b}^T(i+L-1),\ldots,\mathbf{b}^T(i),\ldots,\mathbf{b}^T(i-L+1)]^T, \quad (4)$$

with

$$\mathbf{b}(i)=[b_1^{(1)}(i),\ldots,b_1^{(N_T)}(i),\ldots,b_K^{(1)}(i),\ldots,b_K^{(N_T)}(i)]^T. \quad (5)$$

Vectors $\mathbf{H}\in\mathbb{C}^{LN_R\times K_IN_T(2L-1)}$ and $\mathbf{u}(i)\in\mathbb{Q}^{K_IN_T(2L-1)\times 1}$ corresponding to the UCCI are similarly defined. Vector $\mathbf{n}(i)\in\mathbb{C}^{LM\times 1}$ contains additive white Gaussian noise (AWGN) with covariance $E\{\mathbf{n}(i)\mathbf{n}^H(i)\}=\sigma^2\mathbf{I}$.

III. Equalizer derivation

Fig. 1 shows the receiver block diagram. Let the kth user be the user of interest. Let us further denote

$$\hat{\mathbf{u}}_k(i)=\tilde{\mathbf{u}}(i)-\tilde{\mathbf{u}}(i)\odot\mathbf{e}_k, \quad (6)$$

where

$$\mathbf{e}_k=[\underbrace{0,...,0}_{[(L-1)K+k-1]N_T},\underbrace{1,...,1}_{N_T},\underbrace{0,...,0}_{(LK-k)N_T}]^T, \quad (7)$$

and $\odot$ denotes elementwise vector product and let

$$\beta_k(i)=[b_k^{(1)}(i),...,b_k^{(N_T)}(i)]^T. \quad (8)$$

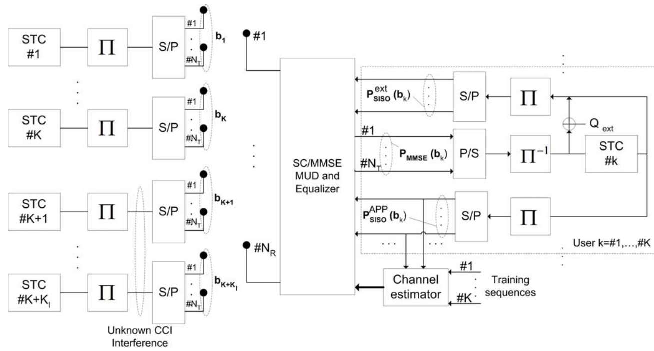

Fig. 1. System model

The vectors $\tilde{\mathbf{u}}(i)$ are obtained by replacing the elements of $\mathbf{u}(i)$ by their soft estimates. Soft estimates of $b_k^{(n)}(i)$, $i=1,...,N_T$ are obtained groupwise as follows

$$\tilde{\beta}_k(i)=\sum_{\delta_q\in\mathcal{Q}^{N_T}}\delta_qP_{SISO}^{ext}(\beta_k(i)=\delta_q), \quad (9)$$

where P_{SISO}^{ext} denotes the extrinsic information obtained after SISO decoding (to be defined in (24)). The signals $b_k^{(n)}(i), n=1,...,N_T$, are jointly detected by filtering the signal

$$\hat{\mathbf{y}}_k(i)=\mathbf{y}(i)-\mathbf{H}\hat{\mathbf{u}}_k(i), i=T+1,...,B+T, \quad (10)$$

using a linear MMSE filter whose weighting matrix $\mathbf{W}_k(i)\in\mathbb{C}^{LN_R\times LN_R}$ satisfies the following criterion

$$\mathbf{W}_k(i)=\arg\min_{\mathbf{W}}||\mathbf{W}^H\hat{\mathbf{y}}_k(i)-\mathbf{A}_k\beta_k(i)||^2. \quad (11)$$

The matrix $\mathbf{A}_k(i)\in\mathbb{C}^{LN_R\times N_T}$ is defined as

$$\mathbf{A}_k=[\mathbf{h}_k^{(1)}...\mathbf{h}_k^{(N_T)}], \quad (12)$$

with $\mathbf{h}_k^{(n)}$ being the $[(L-1)KN_T+kN_T+n]$-th column of the matrix $\mathbf{H}$. The matrix $\mathbf{W}_k(i)$ can be obtained as

$$\mathbf{W}_k(i)=\mathbf{M}_k(i)^{-1}\mathbf{A}_k\mathbf{A}_k^H, \quad (13)$$

where

$$\begin{aligned}\mathbf{M}_k(i) &= \mathbf{H}\mathbf{\Lambda}_k(i)\mathbf{H}^H+\mathbf{H}_I{\mathbf{H}_I}^H+\sigma^2\mathbf{I} \quad (14)\\ &= \mathbf{H}\mathbf{\Lambda}_k(i)\mathbf{H}^H+\mathbf{R}+\sigma^2\mathbf{I}.\end{aligned}$$

The block diagonal matrix $\mathbf{\Lambda}_k(i)$ is defined as

$$\begin{aligned}\mathbf{\Lambda}_k(i) = \ & \text{diag}\{E\{\beta_1(i+L-1)\beta_1^H(i+L-1)\} \quad (15)\\ & -\tilde{\beta}_1(i+L-1)\tilde{\beta}_1^H(i+L-1),...,\\ & E\{\beta_K(i+L-1)\beta_K^H(i+L-1)\}\\ & -\tilde{\beta}_K(i+L-1)\tilde{\beta}_K^H(i+L-1),...,\\ & E\{\beta_{k-1}(i)\beta_{k-1}^H(i)\}-\tilde{\beta}_{k-1}(i)\tilde{\beta}_{k-1}^H(i),\\ & E\{\beta_k(i)\beta_k^H(i)\},\\ & E\{\beta_{k+1}(i)\beta_{k+1}^H(i)\}-\tilde{\beta}_{k+1}(i)\tilde{\beta}_{k+1}^H(i),...,\\ & E\{\beta_K(i-L+1)\beta_K^H(i-L+1)\}\\ & -\tilde{\beta}_K(i-L+1)\tilde{\beta}_K^H(i-L+1)\},\end{aligned}$$

where

$$E\{\beta_j(i-l)\beta_j^H(i-l)\} = \sum_{\delta_q \in \mathcal{Q}^{N_T}} \delta_q \delta_q^H P_{SISO}^{ext}(\beta_j(i) = \delta_q), \quad (16)$$

for all $j = 1, ..., K, l = 0, ..., L-1$, except for $j = k, l = 0$ when

$$E\{\beta_k(i)\beta_k^H(i)\} = \sum_{\delta_q \in \mathcal{Q}^{N_T}} \delta_q \delta_q^H. \quad (17)$$

Assuming that the MMSE filter output $\mathbf{z}_k(i) \in \mathbb{C}^{LN_R \times 1}$ can be viewed as the output of the equivalent Gaussian channel we can write

$$\begin{aligned} \mathbf{z}_k(i) &= \mathbf{W}_k^H(i)\hat{\mathbf{y}}_k(i) \quad (18) \\ &= \mathbf{H}_{e,k}(i)\beta_k(i) + \mathbf{\Psi}_{e,k}(i), \end{aligned}$$

where matrix $\mathbf{H}_{e,k}(i) \in \mathbb{C}^{LN_R \times N_T}$ contains the gains of the equivalent channel defined as

$$\mathbf{H}_{e,k}(i) = E\{\mathbf{z}_k(i)\beta_k^H(i)\} = \mathbf{W}_k^H(i)\mathbf{A}_k. \quad (19)$$

The vector $\mathbf{\Psi}_{e,k}(i) \in \mathbb{C}^{LN_R \times 1}$ is the equivalent additive Gaussian noise with covariance matrix

$$\begin{aligned} \mathbf{R}_{e,k}(i) &= E\{\mathbf{\Psi}_{e,k}(i)\mathbf{\Psi}_{e,k}^H(i)\} \quad (20) \\ &= \mathbf{W}_k^H(i)\mathbf{M}_k(i)\mathbf{W}_k(i) - \mathbf{H}_{e,k}(i)\mathbf{H}_{e,k}(i). \end{aligned}$$

The output of the equivalent channel $\mathbf{z}_k(i)$ and its parameters $\mathbf{H}_{e,k}(i)$ and $\mathbf{R}_{e,k}(i)$ directed to the SISO decoder.

A. SISO Decoding

The SISO channel decoding algorithm used in this paper is a symbol-level *maximum-a-posteriori* (MAP) algorithm from [13]. For the sake of simplicity we omit the full derivation of the MAP algorithm and we refer to [13] and [10]. It should be note that the input required by the decoder is the probability $P(S_i, S_{i+1})$ associated with the transition between two trellis states S_i and S_{i+1} of the STTrC code. The transition probability can be calculated as

$$\begin{aligned} P(S_i, S_{i+1}) &= P(\beta_k(i) = \mathbf{d}^{i,i+1}) \quad (21) \\ &= P_{MMSE}^{ext}(\beta_k(i) = \mathbf{d}^{i,i+1}) \end{aligned}$$

where $\mathbf{d}^{i,i+1} \in \mathbb{C}^{N_T \times 1}$ is the vector of encoder outputs that are associated with the transition (S_i, S_{i+1}). $P_{MMSE}^{ext}(\beta_k(i) = \mathbf{d}^{i,i+1})$ is extrinsic probability obtained by the MMSE detection, which is calculated as

$$P_{MMSE}^{ext}(\beta_k(i) = \mathbf{d}^{i,i+1}) = \quad (22)$$
$$e^{-(\mathbf{z}_k(i) - \mathbf{H}_{k,e}(i)\mathbf{d}_i^{i+1})^H \mathbf{R}_{e,k}^{-1}(i)(\mathbf{z}_k(i) - \mathbf{H}_{k,e}(i)\mathbf{d}_i^{i+1})}.$$

Based on the transition probabilities $P(S_i, S_{i+1})$ the SISO channel decoder calculates the *a posteriori* probabilities for the symbols $\beta_k(i)$, defined as

$$P_{MMSE}^{ext}(\beta_k(i) = \delta_q) = P(\beta_k(i) = \delta_q | \mathbf{z}_k(i), \quad (23)$$
$$\mathbf{H}_{k,e}(i), \mathbf{R}_{e,k}(i), i = T+1, ..., T+B), \delta_q \in \mathcal{Q}^{N_T}.$$

The decoder extrinsic probability is then calculated as

$$P_{SISO}^{ext}(\beta_k(i) = \delta_q) = \frac{P_{SISO}^{app}(\beta_k(i) = \delta_q)}{[P_{MMSE}^{ext}(\beta_k(i) = \delta_q)]^{Q_{ext}}}. \quad (24)$$

The parameter Q_{ext} is an ad-hoc parameter that was introduced in [10]. It is shown in [10] that if the value of Q_{ext} is appropriately chosen so as to be between 0 and 1, the receiver performance can be significantly improved. The same procedure is repeated for all users. The receiver complexity is dominated by the MMSE part which requires inversion of the matrix $\mathbf{M}_k(i)$. The overall complexity is therefore $O\{L^3 N_R^3\}$.

IV. Asymptotic Performance Analysis

In the asymptotic case of ideal feedback (10) becomes

$$\hat{\mathbf{y}}_k(i) = \mathbf{A}_k\beta_k(i) + \mathbf{H}_I\mathbf{u}_I(i) + \mathbf{n}(i), i = T+1, ..., B+T, \quad (25)$$

since all the ISI and known CCI is removed by the soft-cancellation. The MMSE filter weighting matrix becomes

$$\mathbf{W}_k(i) \approx (\mathbf{A}_k\mathbf{A}_k^H + \mathbf{R})^{-1}\mathbf{A}_k. \quad (26)$$

for large SNR. Let $\mathbf{C}$ and $\mathbf{E}$ be two different codeword matrices of size $N_T \times B$ and let us assume that the $\mathbf{C}$ is transmitted. By assuming the equivalent Gaussian channel model (18) it can easily be shown that for the given channel realization $\mathbf{A}_k$ the probability of erroneously deciding in favor of $\mathbf{E}$ can be upper bounded by

$$P(\mathbf{C} \mapsto \mathbf{E}|\mathbf{A}_k) \le e^{-\frac{E_s}{4\sigma^2}d^2(\mathbf{C},\mathbf{E}|\mathbf{A}_k)}, \quad (27)$$

where

$$d^2(\mathbf{C},\mathbf{E}|\mathbf{A}_k) = \sum_{i=T+1}^{T+B} (\mathbf{c}_i - \mathbf{e}_i)^H \mathbf{H}_{e,k}^H \mathbf{R}_{e,k}^{-1} \mathbf{H}_{e,k}(\mathbf{c}_i - \mathbf{e}_i), \quad (28)$$

where $\mathbf{c}_i$ and $\mathbf{e}_i$ are ith columns of $\mathbf{C}$ and $\mathbf{E}$, respectively. Furthermore, it can be shown based on (19), (20) and (26) that

$$\mathbf{H}_{e,k}^H \mathbf{R}_{e,k}^{-1} \mathbf{H}_{e,k} = \mathbf{H}_k^H(\frac{\mathbf{R}}{\sigma^2} + \mathbf{I})^{-1}\mathbf{H}_k. \quad (29)$$

By putting (29) into (28) and by adopting similar approach as in [12] we define

$$\mathbf{y}_k = (\frac{\mathbf{R}}{\sigma^2} + \mathbf{I})^{-\frac{1}{2}}\mathbf{A}_k(\mathbf{c}_i - \mathbf{e}_i). \quad (30)$$

Since it is assumed that the entries of matrix $\mathbf{H}$ (and thereby of $\mathbf{A}_k$ as well) are i.i.d. it can be concluded from (30) that the matrix $(\frac{\mathbf{R}}{\sigma^2} + \mathbf{I})^{-1}$ has a similar effect as a matrix defining a correlation at the receiver side, which was considered in [12]. Therefore, using the result of [12] and by performing similar analysis, it can be concluded that the upper bound of the pairwise error probability equals

$$P(\mathbf{C} \mapsto \mathbf{E}|\mathbf{A}_k) \le (\frac{E_s}{4\sigma^2})^{-sr((\frac{\mathbf{R}}{\sigma^2}+\mathbf{I})^{-1})} \quad (31)$$
$$\prod_{i=0}^{s-1} \lambda_i^{-r((\frac{\mathbf{R}}{\sigma^2}+\mathbf{I})^{-1})}(\mathbf{C},\mathbf{E}),$$

where s is the transmit diversity order of STTrC, λ_i is the eigenvalue of the error matrix $\mathbf{C}-\mathbf{E}$ and $r(\cdot)$ denotes rank of a matrix. Let us now determine the rank of the matrix $(\frac{\mathbf{R}}{\sigma^2}+\mathbf{I})^{-1}$. If $\lambda_{\mathbf{R},i}$ denotes the ith eigenvalue of the matrix $\mathbf{R}$ then the ith eigenvalue of the matrix $(\frac{\mathbf{R}}{\sigma^2}+\mathbf{I})^{-1}$ is equal to $\frac{\sigma^2}{\sigma^2-\lambda_{\mathbf{R},i}}$. In the asymptotic case of large SNR[1] these eigenvalues can be either 0 (for $\lambda_{\mathbf{R},i} \neq 0$) or 1 (for $\lambda_{\mathbf{R},i} = 0$). This, in turn, means that

$$r((\frac{\mathbf{R}}{\sigma^2}+\mathbf{I})^{-1}) = LN_R - r(\mathbf{R}), \quad (32)$$

which finally yields

$$P(\mathbf{C} \mapsto \mathbf{E}|\mathbf{A}_k) \leq (\frac{E_s}{4\sigma^2})^{-s(LN_R - r(\mathbf{R}))} \prod_{i=0}^{s-1} \lambda_i^{-(LN_R - r(\mathbf{R}))}(\mathbf{C},\mathbf{E}). \quad (33)$$

It can be concluded from the analysis above that the maximum achievable diversity order in the presence of UCCI is equal to $s(LN_R - r(\mathbf{R}))$. This diversity order is achievable if the feedback is perfect.

V. Numerical Examples

Performance of the proposed receiver was evaluated through computer simulations. The 4-state QPSK code with $N_T = 2$ presented in [3] was used to encode signals of all MIMO users. The Log-MAP space-time trellis decoder shown in [13] and [10] was used. The user specific random interleavers were assumed. Channel realizations between different antennas and mltipath components are assumed to be mutually independent. Perfect knowledge about the channel and covariance matrix of the UCCI is assumed in the receiver.

In Fig. 2 the SER and FER performances of the proposed receiver are presented for $L = 2$ and different values of K and N_R. The performance is compared to the corresponding ML lower single-user bound. It can be seen that with increasing K and N_R the performance is closer to the corresponding single-user ML bound, due to the increased total diversity order.

In Fig. 3 the SER and FER performances of the proposed receiver are presented for $L = 2$ and $(K, K_I, N_R) = (4, 1, 5)$. ML bound with ideal feedback for $(K, K_I, N_R) = (1, 1, 5)$ case is presented as well. It can be seen that the proposed receiver gradually approaches the case with perfect feedback. For given parameters the rank of the matrix $\mathbf{R}$ is $r(\mathbf{R}) = N_T(2L-1) = 6$. Therefore, diversity loss of order 6 is to be expected comparing to the $(K, K_I, N_R) = (1, 0, 5)$ case. The diversity loss can indeed be observed from comparison with the ML lower single user bound for the $(K, K_I, N_R) = (1, 0, 5)$ case. This was predicted by (33), thereby justifying the analysis presented above.

[1] Note that in high SNR region the assumption of perfect SC is also more likely to be valid than in the low SNR region.

VI. Conclusions

A new iterative receiver scheme for the STTrC coded multiuser system in frequency selective channels is derived. Asymptotic performance analysis is performed in the case of ideal SC cancellation. Simulation results show that the receiver can achieve the corresponding single user ML bound in the absence of UCCI. The presence of UCCI has been shown to have similar effect on the diversity and coding gains as the spatial correlation at the receiver side.

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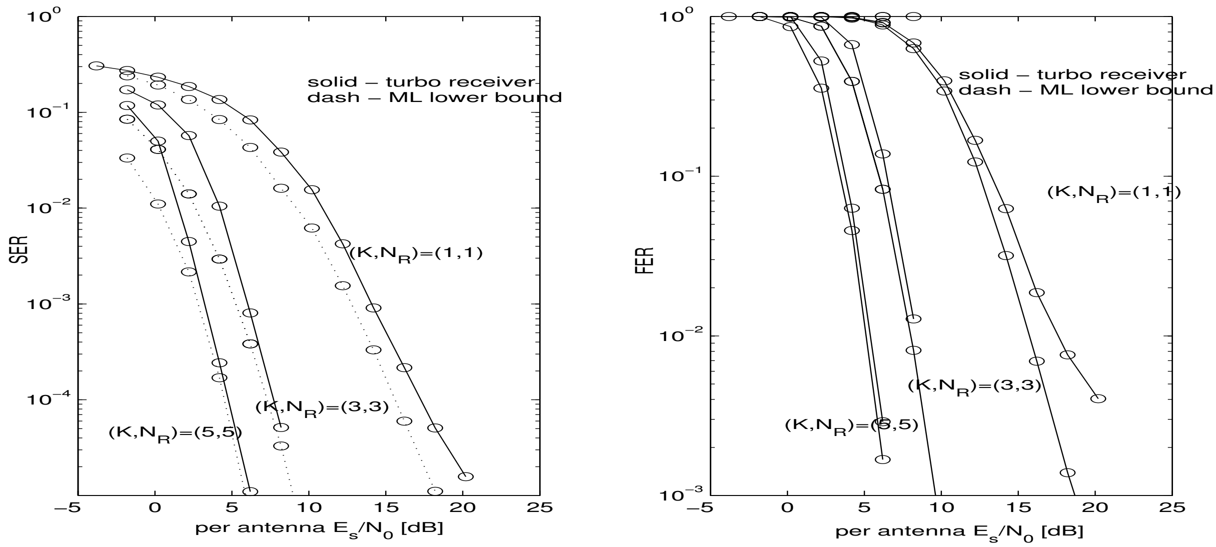

Fig. 2. (a)SER performance, (b) FER performance, $L = 2$, $B = 150$, ideal knowledge of covariance matrix $\mathbf{R}$

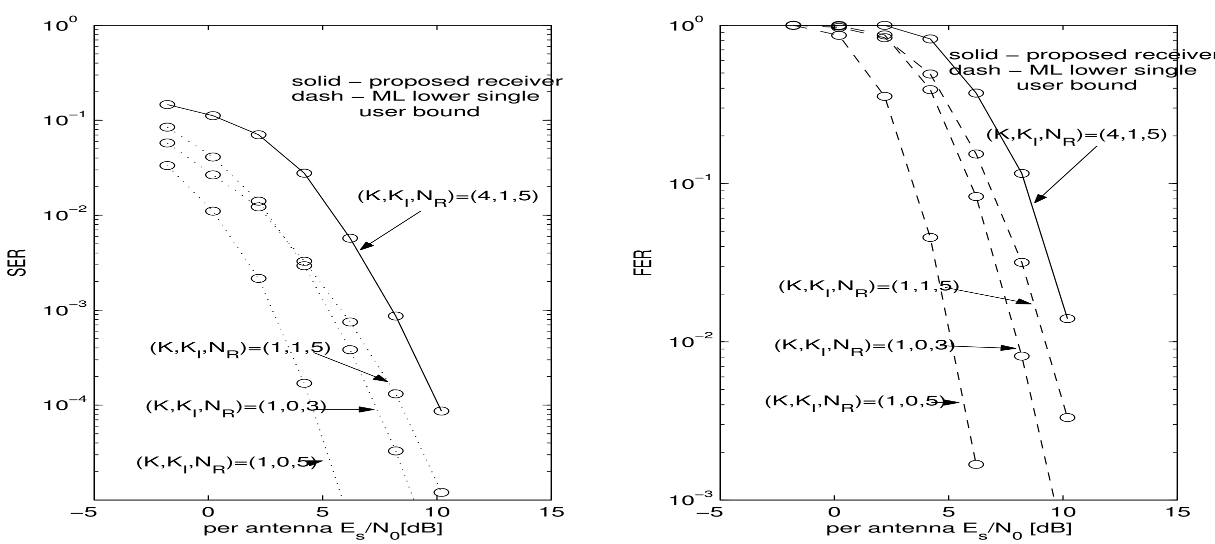

Fig. 3. (a)SER performance, (b) FER performance, $L = 2$, $B = 150$, ideal knowledge of covariance matrix $\mathbf{R}$