# Achieving Full Diversity over the MIMO Fading Channel with Space-Time Precoders and Iterative Linear Receivers

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***Abstract*—We study the performance of space-time bit-interleaved coded modulation (ST-BICM) over quasi-static multiple-antenna fading channels with linear receivers. We show that, under iterative linear detection and decoding, an ST-BICM can achieve full diversity with a special class of space-time precoders. We then study the outage probability at the output of the linear detector that determines the theoretical performance of coded modulations with such receivers. Finally, symbol and word error rate performances under Monte Carlo simulations are shown.**

***Index Terms*—Diversity techniques, iterative decoding, MMSE receivers, space-time precoding.**

## I. Introduction

MULTIPLE-ANTENNA techniques have been shown to provide high transmission rates over fading channels [1] and help in increasing the diversity order of signals over slow fading channels [2]. For uncoded systems, *i.e.* systems not employing error correction coding, achieving maximum diversity requires the use of space-time codes together with maximum likelihood (ML) receivers [3]. For coded systems, *a posteriori* probability detectors are required to recover maximum diversity at the receiver end [4], [5], [6]. In both cases, the complexity at the detector increases exponentially with the number of transmit antennas. On the other hand, linear receivers for uncoded transmission over multiple-antenna quasi-static fading channels have been extensively studied [7], [8], [9], [10], [11], and the achieved diversity orders are far from being optimal even with full-rate space-time precoders [12], [13]. With orthogonal space-time codes, linear receivers can achieve full diversity at the cost of low rate transmission [14]. In this work, we show that by concatenating coded modulations with full-rate space-time precoders, an iterative receiver can recover maximum diversity with a soft-input soft-output (SISO) linear detector [15]. We then study the outage probability [16] of such receivers that provide an information theoretic lower bound on the performance and thus give insight on the achievable diversity orders. The paper is organized as follows: Section II gives the system model and the notations used in the rest of the paper. In Section III, we give the equations for the linear detectors. In Section IV, an analysis of the diversity order for iterative linear receivers is provided, and the design of space-time precoders for such receivers is proposed. Section V gives the outage probability analysis of linear receivers. In Section VI, we show Monte Carlo simulations with iterative receivers and Section VII gives the concluding remarks.

Manuscript received April 26, 2010; revised August 19, 2010 and March 5, 2011; accepted April 15, 2011. The associate editor coordinating the review of this paper and approving it for publication was H.-C. Yang.

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This work has been supported by the Research Council of Norway under the project WILATI+ within the NORDITE framework.

Digital Object Identifier 10.1109/TWC.2011.061311.100699

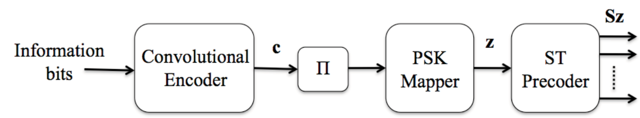


Fig. 1. ST-BICM encoder.

## II. System Model and Notations

We consider a system transmitting via a space-time bit-interleaved coded modulation [17] over a quasi-static frequency non-selective MIMO fading channel with $n_t$ transmit and $n_r$ receive antennas, *i.e.* a codeword undergoes one temporal channel realization. The channel model is written as:

$$\mathbf{y} = \mathbf{H}\mathbf{S}\mathbf{z} + \mathbf{w} \ , \tag{1}$$

where $\mathbf{y} \in \mathbb{C}^{sn_r}$ and $\mathbf{z} \in \Omega = (2^m - \text{PSK})^{sn_t}$ are the vectors of received and transmitted complex symbols, respectively, $m$ being the number of bits per Phase Shift Keying (PSK) modulated symbol, and $s$ being the time spreading of a linear unitary space-time precoding matrix $\mathbf{S}$ of dimensions $sn_t \times sn_t$, also called space-time rotation. The MIMO channel matrix $\mathbf{H}$ with dimensions $sn_r \times sn_t$ is diagonal and given by:

$$\mathbf{H} = \mathbf{I}_s \otimes \mathcal{H} \ , \tag{2}$$

where $\otimes$ denotes Kronecker product, $\mathbf{I}_s$ is the $s \times s$ identity matrix, and $\mathcal{H}$ has dimensions $n_r \times n_t$ and has independent complex Gaussian entries $h_{ij}$ with zero mean and unit variance representing the channel gain from the $j$th transmit to the $i$th receive antennas. The length-$sn_r$ vector of additive white Gaussian noise components $\mathbf{w}$ is assumed to be circularly symmetric with zero mean and variance $N_0$. Digital transmission operates as shown in Fig. 1: information bits are fed to an encoder of rate $R_c \leq 1$. The codeword $\mathbf{c}$ is first interleaved, fed to a $2^m$-PSK mapper, and then space-time precoded. The space-time precoder combines $sn_t$ PSK symbols over the $n_t$ transmit antennas and over $s$ time periods. The resulting

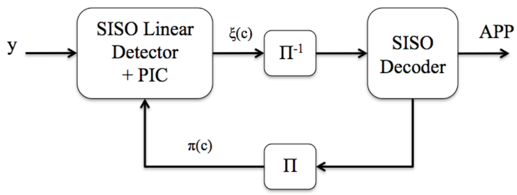


Fig. 2. ST-BICM iterative linear detector and decoder.

frame is sent over the fading channel through the $n_t$ transmit antennas, and the total transmission rate is : $R = mn_tR_c$. The channel coefficients are supposed to be perfectly known to the receiver, but not to the transmitter. At the receiver, a linear detector, following parallel interference cancellation (PIC), provides extrinsic probabilities on modulated symbols based on the received symbols, the channel matrix, and the *a priori* probabilities on coded bits fed back from the SISO decoder, as shown in Fig. 2.

## III. Iterative Linear Detectors

At the output of the channel, the vector $\mathbf{y}$ is fed to a linear SISO detector [15] [18] that takes into account the *a priori* information fed from the SISO channel decoder. The detector can be a Zero Forcing (ZF) or Minimum Mean-Square Error (MMSE) detector. In the sequel, we will consider:

$$\mathbf{G} = \mathbf{HS} \ , \tag{3}$$

to be the channel seen by the PSK symbols. Now let $\bar{z}_j$ denote the *a priori*-based mean of the complex transmitted symbol $z_j$ $(j = 1, \cdots, sn_t)$ computed using the *a priori* probabilities $\pi(c_b)$ on coded bits fed back from the SISO decoder as :

$$\bar{z}_j = \sum_{z_j} z_j \prod_{b=m.j}^{m.j+m-1} \pi(c_b) \ , \tag{4}$$

and define $\bar{\mathbf{z}}_j = \bar{\mathbf{z}} - \bar{z}_j \mathbf{e}^j_{sn_t}$ a vector containing the interference experienced by the $j$th symbol from the $sn_t - 1$ other symbols in the space-time vector. The vector $\mathbf{e}^j_{sn_t}$ of length $sn_t$ contains a value of 1 at position $j$ and 0 elsewhere. The unbiased estimation of $z_j$, obtained via PIC followed by linear MMSE filtering [18], is:

$$\tilde{z}_j = \mathbf{g}_j^\dagger \left( \mathbf{G}\Gamma_j\mathbf{G}^\dagger + \alpha \mathbf{I}_{sn_r} \right)^{-1} \tilde{\mathbf{y}}_j \ , \tag{5}$$

where $.^\dagger$ is the transpose conjugate operator, $\mathbf{g}_j = \mathbf{G}\mathbf{e}^j_{sn_t}$ and $\tilde{\mathbf{y}}_j = \mathbf{y} - \mathbf{G}\bar{\mathbf{z}}_j$ represents the residual term after interference cancellation, and $\Gamma_j$ is a diagonal matrix whose $n$th entry is given by:

$$\Gamma_{jn} = \begin{cases} 1 - |\bar{z}_n|^2 & n \neq j \\ 1 & n = j \end{cases} \ ,$$

and is computed at every iteration from the *a priori* probabilities fed from the SISO decoder. Note that $\alpha = 2N_0$ in the case of MMSE detection. However, although it should be zero for ZF detectors, we will show in the sequel that this is not possible in most cases.

## IV. Diversity of Linear Receivers over Quasi-Static Fading Channels

### A. Linear receivers for uncoded systems

For uncoded systems, the linear detector has exactly the same expression as that of (5) without any prior knowledge on the symbol estimates, *i.e.* $\bar{z}_n = 0 \ \forall \ n$. The signal-to-interference-and-noise ratio (SINR) of the $j$th symbol is thus computed as [7] :

$$\gamma_j = \mathbf{g}_j^\dagger \left( \hat{\mathbf{G}}_j \hat{\mathbf{G}}_j^\dagger + \alpha \mathbf{I}_{sn_r} \right)^{-1} \mathbf{g}_j \ , \tag{6}$$

where $\hat{\mathbf{G}}_j \in \mathbb{C}^{sn_r \times (sn_t - 1)}$ is obtained by removing column $\mathbf{g}_j$ from $\mathbf{G}$. The random variable in (6) has been shown to follow a $\chi^2$ distribution with $2(n_r - n_t + 1)$ degrees of freedom [8], [9], [10], [11]. Hence, the maximum diversity order of uncoded linear receivers over quasi-static channels is:

$$d_u = \lim_{SINR \to \infty} \frac{-\log(P_e)}{\log(SINR)} = n_r - n_t + 1 \ , \tag{7}$$

where $P_e$ is the error probability. Even with the use of a space-time code, uncoded MMSE receivers cannot attain full diversity [12], [13]. Only for small transmission rates (*i.e.* $R < n_t \log\left(\frac{n_t}{n_t - 1}\right)$ ), MMSE receivers can recover maximum diversity $d_{max} = n_t n_r$.

### B. Iterative linear receivers for coded systems

With an iterative receiver, soft information is exchanged between the detector and the channel decoder, and interference between transmit antennas can be removed more efficiently. First, let us consider the channel seen at the output of the linear detector for every symbol, provided that the interference is totally removed between symbols. We have that:

$$\tilde{z}_j = \mu_j z_j + w_j \ , \tag{8}$$

with the mean $\mu_j$ written as:

$$\mu_j = \mathbf{g}_j^\dagger \left( \mathbf{g}_j \mathbf{g}_j^\dagger + \alpha \mathbf{I}_{sn_r} \right)^{-1} \mathbf{g}_j \tag{9}$$

$$= \mathbf{g}_j^\dagger \Omega_j^{-1} \mathbf{g}_j \ . \tag{10}$$

and $w_j$ being the equivalent noise term with variance $\sigma_{w_j} = \mu_j (1 - \mu_j)$ [19]. The matrix $\Omega_j$ being a $sn_r \times sn_r$ Hermitian matrix, one can see that, for $n_r > 1$, we should have $\alpha > 0$ for the matrix to be non-singular and thus invertible. Hence, a classical ZF detector where $\alpha = 0$ is not feasible for iterative receivers. If $\alpha > 0$, and after eigenvalue decomposition, $\Omega_j$ can be written as:

$$\Omega_j = \mathbf{D}_j \Lambda_j \mathbf{D}_j^{-1} \ , \tag{11}$$

where $\mathbf{D}_j$ is the $sn_r \times sn_r$ matrix containing the eigenvectors of $\Omega_j$, and the non-zero diagonal entries $\lambda_i$ of $\Lambda_j$ are the corresponding eigenvalues. Now by inverting $\Omega_j$, we obtain:

$$\Omega_j^{-1} = \mathbf{D}_j^{-1} \Lambda_j^{-1} \mathbf{D}_j = \mathbf{D}_j^\dagger \Lambda_j^{-1} \mathbf{D}_j \ , \tag{12}$$

where the second equality holds because $\Omega_j$ is Hermitian, thus $\mathbf{D}_j$ is a rotation. Now inserting (12) into (10), we obtain:

$$\mu_j = \mathbf{g}_j^\dagger \mathbf{D}_j^\dagger \Lambda_j^{-1} \mathbf{D}_j \mathbf{g}_j = \mathbf{f}_j^\dagger \Lambda_j^{-1} \mathbf{f}_j \ . \tag{13}$$

The distribution of $\mathbf{g}_j$ being invariant after rotation, $\mathbf{g}_j$ and $\mathbf{f}_j$ have the same distribution, and we thus obtain that:

$$\mu_j = \mathbf{f}_j^\dagger \Lambda_j^{-1} \mathbf{f}_j = \sum_{i=1}^{sn_r} \frac{|f_i|^2}{\lambda_i} \tag{14}$$

follows a $\chi^2$ distribution with $2sn_r$ degrees of freedom. This means that, if the *a priori* information fed from the channel decoder is perfectly reliable, maximum diversity $n_t n_r$ is attained when using a full-spreading (*i.e.* $s = n_t$) space-time rotation. In other words, without a space-time rotation, the receiver sees $n_t$ interfering single-input multiple-output (SIMO) fading sub-channels carrying each a diversity order of $1 \times n_r$. With a space-time precoder, the receiver sees $sn_t$ interfering SIMO fading sub-channels having each a diversity order of $1 \times sn_r$. Hence, all PSK symbols can potentially achieve full diversity at a cost of a stronger interference. For this reason, the choice of the space-time rotation that allows for interference suppression is crucial.

### *C. Design of space-time precoders for iterative linear receivers*

As explained in the previous section, every transmitted symbol $z_j$ achieves full diversity when a full-spreading space-time rotation is used, provided interference is totally removed. For this reason, the critical part of the detector in (5) is the interference cancellation operation, given by:

$$\begin{aligned} \tilde{\mathbf{y}}_j &= \mathbf{y} - \mathbf{G}\bar{\mathbf{z}}_j & (15)\\ &= \mathbf{G}\left(\mathbf{z} - \bar{\mathbf{z}}_j\right) + \mathbf{w} \ . & (16) \end{aligned}$$

To have a closer look on $\mathbf{G}$, let us first write the space-time rotation matrix as:

$$\mathbf{S} = \begin{bmatrix} \theta_{1,1}^1 & \theta_{1,2}^1 & \cdots & \theta_{1,sn_t}^1 \\ \vdots & \vdots & \vdots & \vdots \\ \theta_{n_t,1}^1 & \theta_{n_t,2}^1 & \cdots & \theta_{n_t,sn_t}^1 \\ \vdots & \vdots & \vdots & \vdots \\ \theta_{1,1}^s & \theta_{1,2}^s & \cdots & \theta_{1,sn_t}^s \\ \vdots & \vdots & \vdots & \vdots \\ \theta_{n_t,1}^s & \theta_{n_t,2}^s & \cdots & \theta_{n_t,sn_t}^s \end{bmatrix} , \tag{17}$$

where the entries $\theta_{u,v}^\ell$ are complex numbers. We then have (18) at the top of the next page, where the $\Phi_p$ are $n_t \times sn_t$ matrices. Now let us compute the covariance matrix of the equivalent channel matrix $\mathbf{G}$ as:

$$\mathbf{C} = \mathbb{E}\left[\mathbf{G}\mathbf{G}^\dagger\right] = \mathbb{E}\begin{bmatrix} \Phi_1\Phi_1^\dagger & \cdots & \Phi_1\Phi_s^\dagger \\ \vdots & \vdots & \vdots \\ \Phi_s\Phi_1^\dagger & \cdots & \Phi_s\Phi_s^\dagger \end{bmatrix} . \tag{19}$$

In order for the interference encountered by the $sn_t$ PSK symbols to be independent, the covariance matrix $\mathbf{C}$ should be a scaled identity matrix. First, by considering the block-diagonal part of $\mathbf{C}$, *i.e.* the entries at $\Phi_p\Phi_p^\dagger$, only diagonal terms remain as channel coefficients are independent. Second, for the off-block-diagonal entries of $\mathbf{C}$ to be null, *i.e.* for $\Phi_p\Phi_q^\dagger$ $\forall\ p \neq q$, the $s$ sub-parts of every column in $\mathbf{S}$ should be orthogonal to each other, *i.e.*:

$$< \mathbf{\Theta}_r^b, \mathbf{\Theta}_r^d >= 0 \ \ \forall\ r, b \neq d \ , \tag{20}$$

with

$$\mathbf{\Theta}_r^b = \left[\theta_{1,r}^b, \cdots, \theta_{n_t,r}^b\right]^t \ , \tag{21}$$

where $.^t$ denotes the transpose operator. Finally, the space-time precoder should ensure that the $sn_t$ transmitted symbols encounter the same residual interference-plus-noise variance. The reason is that usually, in multi-user detection, PIC is most efficient when all the users have equal power [20]. It is thus necessary that the symbols at the output of the equivalent channel $\mathbf{G}$ have equal power so that, after PIC, they face residual interference having the same variance. This property is ensured if the $s$ sub-parts of every column of the space-time rotation $\mathbf{S}$ have equal norm, and we thus obtain that $\mathbf{G}\left(\mathbf{z} - \bar{\mathbf{z}}_j\right) \sim \mathcal{N}\left(0, \left(\mathbf{\Theta}_r^b\right)^\dagger \mathbf{\Theta}_r^b \mathbf{I}_{sn_t}\right) \forall\ r, b$. If this property is not ensured, a situation similar to the near-far problem in multi-user detection will occur, in which the dominance of certain sub-channels over others leads to performance degradation [21]. Moreover, the equal-norm property of the subparts of every column leads to the fact that the eigenvalues of $\Omega_j$ will be all equal and thus the coding gain is enhanced, as shown with *a posteriori* probability (APP) detectors in [4]. The equal-norm and orthogonal sub-parts conditions, called the "Genie conditions" in [22], are also necessary for optimal iterative APP detection and decoding over MIMO channels. They also correspond to the properties of Unitary Trace-Orthogonal Space-Time Block Codes [12] that allow for optimal uncoded performance of MMSE receivers. Finally, it should be noted that, as detection complexity in (5) is proportional to $s$, partial-spreading rotations (with $s < n_t$) can be used at the cost of lower diversity orders. *Dispersive Nucleo Algebraic* (DNA) space-time precoders [4], that exist for $s \in [1, \cdots, n_t]$ and satisfy the "Genie conditions", allow to achieve a diversity order of $sn_r$ over MIMO channels with linear detectors as well.

## V. Outage Probability of MMSE SISO Detectors

In this section, we provide an information-theoretic lower bound on the performance of iterative MMSE receivers. Namely, we compute the Gaussian input outage probability [16] at the output of the MMSE detector. As discussed in [10], [11], two encoding strategies can be considered: separate spatial encoding where symbols are encoded by independent codes, and joint spatial encoding, where a single stream of coded bits is demultiplexed onto $sn_t$ streams and sent through PSK symbols. The first strategy is the most vulnerable in terms of outage probability, as a global outage event occurs if any of the $sn_t$ streams is in outage. For this reason, we will consider the study of outage probability under separate spatial encoding for the SISO MMSE detector to show its superiority even in the worst case scenario. Based on (4), we consider that the symbol $\tilde{z}_j$ sees an equivalent Additive White Gaussian Noise (AWGN) channel with mean $\mu_j$ as in (14) and variance $\sigma_j^2 = \mu_j\left(1 - \mu_j\right)$ [19]. The outage probability per symbol is thus given by [10]:

$$\begin{aligned} P_{out} &= 1 - \left(P\left[\mathcal{I}\left(z;\tilde{z}\right) \geq \frac{sR}{sn_t}\right]\right)^{sn_t} & (22)\\ &= 1 - \left(P\left[\log_2\left(1 + \rho_j\mu_j\right) \geq R_c m\right]\right)^{sn_t} \ , & (23) \end{aligned}$$

$$\mathbf{G} = \mathbf{HS} = \begin{bmatrix} \Phi_1 \\ \vdots \\ \Phi_s \end{bmatrix} = \begin{bmatrix} \sum_{a=1}^{n_t} h_{1,a}\theta^1_{a,1} & \sum_{a=1}^{n_t} h_{1,a}\theta^1_{a,2} & \cdots & \sum_{a=1}^{n_t} h_{1,a}\theta^1_{a,sn_t} \\ \vdots & \vdots & \vdots & \vdots \\ \sum_{a=1}^{n_t} h_{n_r,a}\theta^1_{a,1} & \sum_{a=1}^{n_t} h_{n_r,a}\theta^1_{a,2} & \cdots & \sum_{a=1}^{n_t} h_{n_r,a}\theta^1_{a,sn_t} \\ \vdots & \vdots & \vdots & \vdots \\ \sum_{a=1}^{n_t} h_{1,a}\theta^s_{a,1} & \sum_{a=1}^{n_t} h_{1,a}\theta^s_{a,2} & \cdots & \sum_{a=1}^{n_t} h_{1,a}\theta^s_{a,sn_t} \\ \vdots & \vdots & \vdots & \vdots \\ \sum_{a=1}^{n_t} h_{n_r,a}\theta^s_{a,1} & \sum_{a=1}^{n_t} h_{n_r,a}\theta^s_{a,2} & \cdots & \sum_{a=1}^{n_t} h_{n_r,a}\theta^s_{a,sn_t} \end{bmatrix}, \tag{18}$$

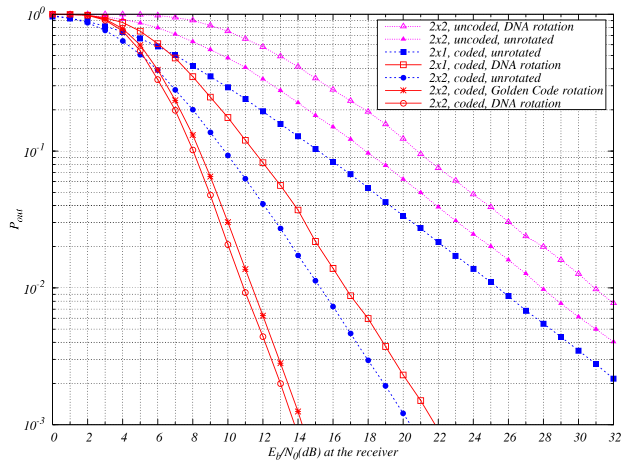


Fig. 3. MMSE detector, outage probability with separate encoding, Gaussian input, $R = 12$ bits/s/Hz.

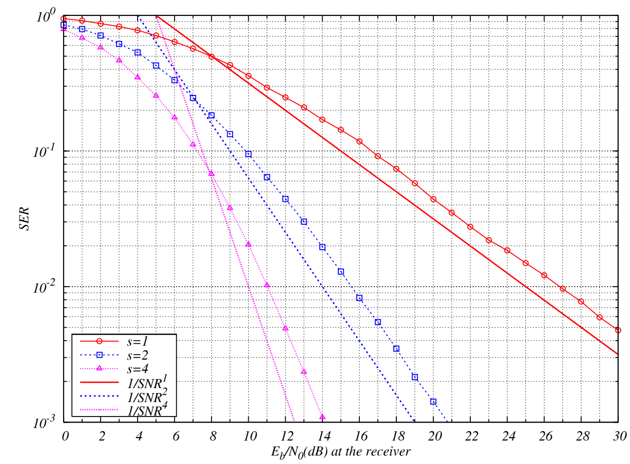


Fig. 4. MMSE detector, uncoded transmission, ideal (genie-aided) interference cancellation, $n_t = 4$, $n_r = 1$, DNA precoders.

where $\rho_j = \frac{mR_c}{\sigma_j^2}$ . Fig. 3 shows the outage probability comparisons between uncoded and coded MMSE receivers over quasi-static MIMO fading channels with $n_t = 2$. The overall rate is $R = 12$ bits/s/Hz. For uncoded MMSE detectors, even with a space-time rotation, a diversity order of $n_r - n_t + 1$ is achieved, as shown by the violet curves. For coded systems, a diversity order of $n_r$ is achieved for systems not employing a space-time precoder (*i.e.* unrotated systems) (blue curves) and $d_{max} = n_t n_r$ is achieved with a full-spreading space-time rotation (red curves). A space-time precoder thus leads every modulated symbol to achieve full diversity provided the interference between symbols is totally removed. It should be noted that the Golden Code rotation [23], that is optimal with ML detection, does not provide optimal performance with coded MMSE transmission. The reason is that it does not satisfy the equal-norm property, leading to the fact that the eigenvalues of $\Omega_j$ are not equal.

## VI. Simulation Results

In this section, symbol and word error rate performance of ST-BICM with linear receivers are shown. The code used is the 16-state non-recursive non-systemmatic convolutional (NRNSC) code with generator polynomials $(23, 35)_8$, and the interleavers are pseudo-randomly generated. We consider Quadrature Phase Shift Keying (QPSK) modulation with Gray mapping, so the extrinsic probabilities at the output of the detector are given by:

$$\xi(c_{2n}) = 2\Re(\tilde{z}_n) - 1 \ , \tag{24}$$

$$\xi(c_{2n+1}) = 2\Im(\tilde{z}_n) - 1 \ , \tag{25}$$

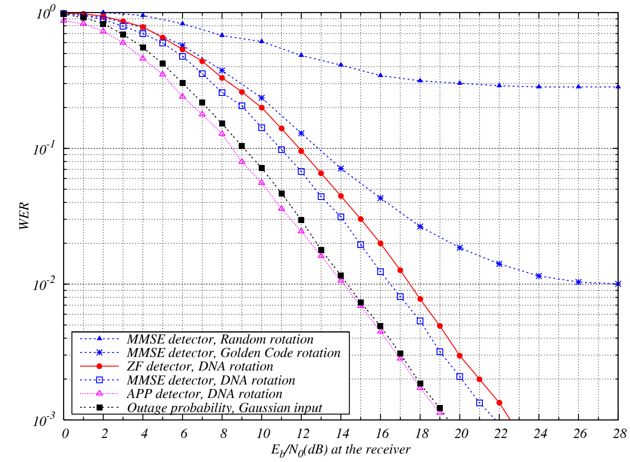


Fig. 5. ST-BICM with linear detectors, iterative interference cancellation, $n_t = 2$, $n_r = 1$, $N = 1024$.

with $n = 0, \cdots, N/2-1$, and where $N$ is the number of bits per codeword. The space-time precoders are the full-rate DNA precoders that satisfy the properties of Section IV-C. In Fig. 4, symbol error rate performance of the uncoded genie-aided MMSE detector is shown over the $4 \times 1$ quasi-static fading channel, in which interference was removed in the simulation (*i.e.* we set : $\bar{z}_j = z_j \ \forall j$ in (5)). The three curves represent transmission without rotation ($s = 1$), partial-spreading DNA rotation ($s = 2$), and full-spreading DNA rotation ($s = 4$). In each case, a diversity order of $sn_r$ is achieved, thus a full-spreading rotation is required to recover maximum diversity.

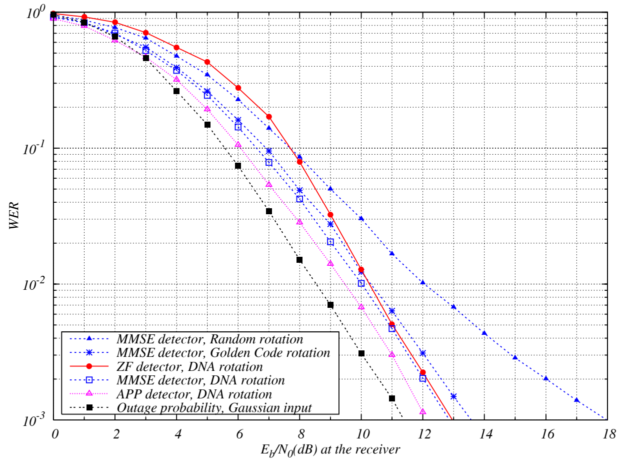


Fig. 6. ST-BICM with linear detectors, iterative interference cancellation, $n_t = n_r = 2$, $N = 1024$.

In Fig. 5 and 6, word error rate performance of ST-BICM with iterative interference cancellation for a codeword length of $N = 1024$ is shown for $2 \times 1$ (Fig. 5) and $2 \times 2$ (Fig. 6) MIMO quasi-static fading channels. The comparison of both MMSE and ZF receivers (with $\alpha = 0.05$) is made with outage probability for different space-time precoders. In both figures, maximum diversity $d_{max} = sn_r$ is achieved with DNA rotations and performance less than 2 dB from outage probability is achieved, while the receivers fail to remove the interference with both a random rotation [24] and the Golden code rotation. Moreover, comparison is made with optimal DNA precoded ST-BICM with APP detectors [4], [5]. The MMSE detectors are obviously outperformed by these detectors, but the APP detectors have a complexity that is exponential in the number of transmit antennas (*i.e.* proportional to $2^{smn_t}$). In addition, the gap between the two types of detectors is reduced as the number of receive antennas is increased, reaching less than 1 dB with $n_r = 2$, as shown in Fig. 6.

## VII. Conclusions

We proposed space-time bit-interleaved coded modulations that achieve maximum diversity over a multiple-antenna channel with iterative linear receivers. We show that, assuming interference is totally removed between modulated symbols, the linear detector attains maximum diversity using a space-time rotation. Moreover, we show that, under specific properties of the space-time rotation, the channel decoder is capable of removing the interference between the transmitted symbols. The outage probability of these receivers is then studied in order to provide an information-theoretic bound on the performance. We finally show symbol and word error rate performances under Monte Carlo simulations.

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