Array Processing for Simplified Turbo Decoding of Interleaved Space-Time Codes

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Abstract- **Space-time coded architectures that deploy multiple transmit-receive antennas with interleaved space-time codes are considered. Decoding is based on the iterative concatenation of a demapping stage with a decoding stage. We describe several simplified demapping algorithms based on array processing methods that can include either soft or hard feedback information. These algorithms are particularly suited for application in space-time architectures that deploy bit-interleaved convolutional codes and transmission over block fading channels. Performance results show that space-time bit-interleaved convolutional codes with turbo array processing provide high coding, and spatial/temporal diversity gains.**

I. INTRODUCTION

It is well known that multiple-transmit multiple-receive antenna architectures can increase the capacity of wireless communication systems [6], [7]. To approach the capacity limits we need to deploy powerful space-time coding schemes that exploit the spatial-temporal diversity of rich scattering environments. For low transmission rates, and small number of transmit antennas several space-time coding schemes have been proposed. For instance the space-time trellis codes in [8], or the space-time bit-interleaved codes in [9]-[13]. The space-time coding problem becomes more complicated if very high data rates have to be achieved, i.e., we deploy a high number of transmit antennas with high order modulation. On one hand, this is due to the difficulty of designing coding schemes that are capable of achieving full spatial diversity and delivering high coding gains. On the other hand, the optimum decoding algorithm that is based on the maximum-likelihood principle is characterized by a complexity that limits the practical implementation. However, when multiple receive antennas are available array processing can significantly simplify decoding complexity.

In this paper we consider space-time architectures that deploy space-time bit-interleaved codes. Essentially, in these architectures bit interleaving separates coding and modulation. The encoder can be a block, a convolutional, or a turbo-encoder. Coding/interleaving can be done only in time, i.e., each layer is independently coded/interleaved, as well as in both space and time, i.e., across layers.

We consider transmission over block fading channels, i.e., blocks of coded information are transmitted over a small number of bursts that experience uncorrelated frequency non-selective fading. The block fading channel model is appropriate for wireless communication systems that for instance deploy slow frequency hopping, e.g., the GSM/EDGE system [13].

Practical decoding is based on a two steps procedure. First, the received samples are demapped to generate soft information on the coded bits. Then, after deinterleaving decoding takes place. Further, demapping and decoding are concatenated in an iterative fashion following the turbo decoding approach.

Optimal demapping in flat fading [9], [12] requires to compute the a posteriori probabilities of all coded bits, which has a complexity that grows exponentially with the number of transmit antennas and the number of bits per modulation symbol on each antenna. Therefore, decoding may be too complex even for few antennas and low modulation orders.

The coding/decoding problem in inter-symbol interference channels is addressed in [10], [11], [13].

Assuming that multiple receive antennas are available, we seek complexity reduction through array processing [1], [3], [4]. We describe an array processor that performs spatial decorrelation by including second order statistics of the coded symbols. We consider a simplification of it that requires only estimation of the mean and power of the transmitted symbols. In turn, this estimation can be performed with a conventional soft-out MAP decoder. We also highlight the conditions under which this SOFT spatial decorrelator becomes similar to the known MMSE/ZF array processor, and the successive interference cancellation processor [6].

Finally, we study the maximum attainable diversity bounds and the outage probability performance of these architectures. We report simulation results for several ST bit-interleaved architectures with minimum complexity convolutional codes and transmission at 2 and 4 bits/s/Hz.

II. TRANSMITTER, CHANNEL, AND DECODER MODELS

A. Coded Architectures with Multiple Transmit Antennas

We consider a space-time coded architecture with N_T transmit and N_R receive antennas. A block of information bits $\underline{b} = [b_1 \dots b_{N_b}]$ is encoded and modulated into N_T blocks of complex symbols (one per transmit antenna/layer) that belong to the M-PSK or M-QAM signal set.

Mainly two coded architectures are possible. In the first one, coding is done independently on each transmit antenna. Therefore, each block of bits $\underline{b}$ is first S/P converted into N_T sub-blocks that are coded and modulated separately. We refer to it as *temporal coding* of a multiple transmit antenna system. In a second and more general approach coding is done across layers and it is referred to as *space-time coding*.

In terms of coding we can follow either the trellis coded modulation (TCM) approach, or the bit-interleaved coded

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modulation (BICM) approach. In the former coding and modulation are done in a single step. In the latter coding is separated from modulation by a bit interleaver. The extension to multiple transmit antenna systems is respectively referred to as space-time trellis coded modulation ST-TCM [8] and space-time bit-interleaved coded modulation ST-BICM [9].

In this work we consider the deployment of bit-interleaved codes. When coding is performed on independent layers the resulting architecture is depicted in Fig. 1 and referred to as T-BICM. When coding is performed across layers the resulting architecture is depicted in Fig. 2 and referred to as ST-BICM.

The encoder can be a block encoder, a convolutional encoder, or a turbo-encoder. After bit interleaving the encoded bits are mapped into constellation symbols. Transmission from the antennas is simultaneous.

B. Channel Model

We assume a channel model where the fading is considered static over a number of transmitted symbols and then randomly changes. In other words, a block of coded symbols is transmitted over N_B bursts that are independently faded [12]. It is interesting to note that such a model applies in systems that use frequency hopping, e.g., the GSM/EDGE system.

The complex symbol that is transmitted by antenna t at time instant kT on burst l, is obtained by memoryless mapping of N coded and interleaved bits [9], i.e., $x_k^{t,l} = \mu^t(\underline{d}_k^{t,l})$, with $\underline{d}_k^{t,l} = [d_{Nk+1}^{t,l}, ..., d_{Nk+N}^{t,l}]$ being the vector of bits to be mapped.

With Nyquist filtering, and transmission over a frequency non-selective fading channel, the *k-th* matched filter output sample of burst l, and receive antenna r is

$$y_k^{r,l} = \sum_{t=1}^{N_T} (h^{r,t,l} x_k^{t,l}) + n_k^{r,l} . \qquad (1)$$

The equivalent channel impulse response during burst l of the link between the receive antenna r and the transmit antenna t, is given by $h^{r,t,l}$. It is here assumed to be complex Gaussian with zero mean and unit-variance (Rayleigh fading model). No channel variation is assumed over a given burst, however distinct bursts experience independent fading. Further, the fading is uncorrelated across distinct antenna links. In (1) $n_k^{r,l}$ is the AWGN contribution that is assumed to have mean zero, variance N_0, and to be independent across the receive antennas.

Frequency selective channels are considered in [10], [11], [13].

C. Matrix Representation

The MIMO system can be represented in matrix notation by $\underline{y}_k^l = \underline{H}(l)\underline{x}_k^l + \underline{n}_k^l$. Since the demapper operates by observing samples at a given time instant (see below), we can drop the dependency from index k, and l, and write

$$\underline{y} = \underline{H}\,\underline{x} + \underline{n} . \qquad (2)$$

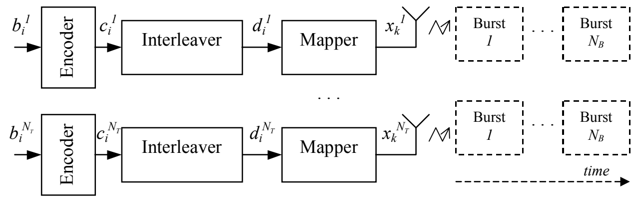

Fig.1. T-BICM: *Time-coded* bit-interleaved modulation with multiple transmit antennas.

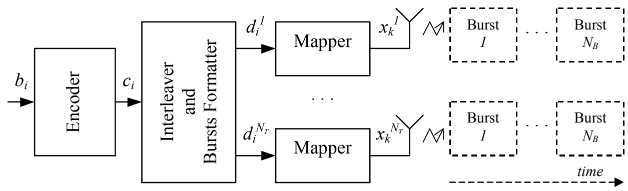

Fig.2. ST-BICM: *Space-time coded* bit-interleaved modulation with multiple antennas.

$$\underline{H} = \begin{bmatrix} h^{1,1} & ... & h^{1,N_T} \\ ... & ... & ... \\ h^{N_R,1} & ... & h^{N_R,N_T} \end{bmatrix} \ (3) \quad \underline{x} = \begin{bmatrix} x^1 \\ ... \\ x^{N_T} \end{bmatrix} \ (4) \quad \underline{n} = \begin{bmatrix} n^1 \\ ... \\ n^{N_R} \end{bmatrix} \ (5)$$

Further,

$$\underline{y} = \underline{h}_{(t)}\, x^t + \underline{H}_{(t)}\, \underline{x}_{(t)} + \underline{n} = \underline{h}_{(t)}\, x^t + \underline{z}_{(t)} \qquad (6)$$

where $\underline{h}_{(t)}$ is the k-th column of (3), $\underline{H}_{(t)}$ is the matrix obtained from (3) suppressing the k-th column, $\underline{x}_{(t)}$ is the interfering symbols vector, x^t is the symbol of layer t. Finally, $\underline{z}_{(t)}$ is the interference-plus-noise vector seen by layer t.

D. Turbo Decoding

The optimal decoder operates by searching for the maximum likelihood solution and takes into account the encoder, interleaver, and modulator structures. Sub-optimal decoding is based on the turbo concept of concatenating in an iterative fashion a demapper with a decoder [9], [12].

The goal of the demapper is to compute soft information on the coded bits (either a posteriori probabilities, or log-likelihoods, or log-likelihood ratios) by observing the antenna samples. After re-arranging and de-interleaving, the soft information is fed to the decoder.

The goal of the decoder is to recover the block of transmitted information bits. It is assumed it can provide soft or hard feedback information on the coded bits to the demapper. For instance a MAP/BCJR soft-out decoder [2] or a hard-out Viterbi decoder can be used.

Multiple demapping/decoding iterations should approach the optimal ML solution.

III. DEMAPPING

The goal of the demapper is to compute the a posteriori probabilities of the coded bits by observing $N_S N_B$ channel samples from N_R antennas, i.e.,

$$\lambda(d_{Nk+i}^{t,l} = \pm 1) = P[d_{Nk+i}^{t,l} = \pm 1 \mid y_k^{1,l}, ..., y_k^{N_R,l}] \qquad (7)$$

for all $t = 1,...,N_T$, $l = 1,...,N_B$, $k = 1,...,N_S$, $i = 1,...,N$.

Since the demapper operates on a sample by sample basis, for easy of notation we can drop the dependency from index k and l, and simply write for the a posteriori log-likelihood

$$\Lambda(d_n^t = \pm 1) = \ln \lambda(d_n^t = \pm 1)\,. \tag{8}$$

Several optimum and sub-optimum algorithms can be used to compute (8). We consider the optimum a posteriori probabilities calculator (MIMO-APP), and a soft-in soft-out spatial decorrelator (SOFT-DCR). We show when the latter coincides with MMSE/ZF array processing and with successive interference cancellation. All these algorithms require estimation of the channel matrix $\underline{H}$.

A. MIMO APP Calculator

Details on the optimal computation of (8) can be found in [9], [12]. The log-likelihood of bit d_n^t being +1 is

$$\Lambda(d_n^t = +1) = \ln \sum\nolimits_{\underline{\hat{x}} \in \underline{X}(d_n^t = +1)} (\, p[\underline{y} \,|\, \underline{\hat{x}}] P[\underline{\hat{x}}]) \tag{9}$$

where $\underline{X}(d_n^t = +1)$ is the set of all possible combinations of transmitted symbols with the constraint that symbol x^t, transmitted from antenna t, has bit n equal to +1. The conditional probability density function $p[\underline{y} \,|\, \underline{\hat{x}}]$ under the assumption of knowing the channel propagation matrix, is Gaussian [9], [12]. The a priori probability $P[\underline{\hat{x}}]$ can be approximated with the extrinsic information provided by a soft-out decoder in the previous decoding iteration.

It should be noted that samples from multiple receive antennas are combined in a maximal ratio combining fashion. However, the algorithm does not require multiple receive antennas.

From (9) it is clear that the complexity of the algorithm grows exponentially with the number of transmit antennas, and the number of bits per symbol.

B. Soft-Input Soft-Output Spatial Decorrelator

Let us assume to have $N_R \geq N_T$. Then, the interference-plus-noise vector seen by layer t is, from (6), $\underline{z}_{(t)} = \underline{y} - \underline{h}_{(t)}\, x^t$. Its covariance matrix is[1]

$$\underline{K}_{\underline{z}_{(t)}} = E\{(\underline{z}_{(t)} - \underline{m}_{\underline{z}_{(t)}})(\underline{z}_{(t)} - \underline{m}_{\underline{z}_{(t)}})^{\perp}\} = \underline{H}_{(t)} \underline{K}_{\underline{x}_{(t)}} \underline{H}_{(t)}^{\perp} + \underline{K}_n \tag{10}$$

where $\underline{K}_n$ is the covariance of the thermal noise, while the covariance of the interference vector is

$$\underline{K}_{\underline{x}_{(t)}} = E\{(\underline{x}_{(t)} - \underline{m}_{\underline{x}_{(t)}})(\underline{x}_{(t)} - \underline{m}_{\underline{x}_{(t)}})^{\perp}\}\,. \tag{11}$$

Therefore, if we model the impairment with a multivariate Gaussian process, the logarithm of the probability density function conditioned on a given hypothetical symbol $\hat{x}^t$ is

$$\ln p(\underline{z}_{(t)} \,|\, \hat{x}^t) \sim -\frac{1}{2}(\underline{y} - \underline{h}_{(t)}\, \hat{x}^t - \underline{m}_{\underline{z}_{(t)}})^{\perp} \underline{K}_{\underline{z}_{(t)}}^{-1} (\underline{y} - \underline{h}_{(t)}\, \hat{x}^t - \underline{m}_{\underline{z}_{(t)}})\,. \tag{12}$$

From (12) the log-likelihood can be approximated as

$$\Lambda(d_n^t = +1) \cong -\min_{\hat{x}^t \in X(d_n^t = +1)} \{(\,\underline{\hat{z}}_{(t)} - \underline{m}_{\underline{z}_{(t)}})^{\perp} \underline{K}_{\underline{z}_{(t)}}^{-1} (\,\underline{\hat{z}}_{(t)} - \underline{m}_{\underline{z}_{(t)}})\} \tag{13}$$

where $X(d_n^t = +1)$ is the set of 2^{N-1} symbols x^t that have bit n equal to +1, and $\underline{\hat{z}}_{(t)} = \underline{y} - \underline{h}_{(t)} \hat{x}^t$.

In the absence of a priori information we can assume the transmitted bits/symbols to be i.i.d. with zero mean. This implies $\underline{K}_{\underline{x}_{(t)}} = E_S / N_T\, \underline{I}$, and the algorithm coincides with the spatial decorrelator described in [5] and [14] for flat fading.

On the contrary, whenever a priori information is available about the covariance of the impairment, this can be included in the computation of (10). The covariance of the interference can be estimated with an ad hoc decoder. In this case we refer to the algorithm as *soft-input soft-output spatial decorrelator*.

A simpler way to proceed is described in what follows. At the first demapping pass we assume the symbols i.i.d. with power E_S / N_T and zero mean. In the following demapping passes we can still assume the symbols i.i.d. but with power and mean estimated by a soft-output decoder. Thus, the covariance of the interference vector is diagonal with elements $\underline{K}_{\underline{x}_{(t)}}(j,j) = (M_j - |m_j|^2)$, with $M_j = E\{|x^j|^2\}$, and $m_j = E\{x^j\}$. It is simple to obtain an estimation of the mean/power of the coded symbols by using the soft-outputs of a conventional MAP-BCJR decoder [2]. In general the covariance (10) needs to be computed and inverted at any given time instant and at each demapping pass. When the symbols are constant amplitude the covariance inversion takes place only once per burst.

If we delete constant terms that do not depend on the hypothetical symbol, the computation of (13) yields

$$\Lambda(d_n^t = +1) \sim \max_{\hat{x}^t \in X(d_n^t = +1)} \{\mathrm{Re}\left\{\hat{x}^{t^*} \underline{w}_{(t)}^T \left(2\underline{y} - 2\underline{m}_{\underline{z}_{(t)}} - \hat{x}^t \underline{h}_{(t)}\right)\right\}\} \tag{14}$$

having defined the receive array weights as $\underline{w}_{(t)}^T = \underline{h}_{(t)}^{\perp} \underline{K}_{\underline{z}_{(t)}}^{-1}$.

C. MMSE Array Processor and ZF Array Processor

In this section we investigate the relationship among the SOFT spatial decorrelator and the well known minimum-squared-error and zero forcing array processors [15].

Let $N_R \geq N_T$, then under the hypothesis of i.i.d. zero mean transmitted symbols, the MMSE array processor detects layer t by weighting the receive antennas with

$$\underline{w}_{(t),MMSE}^T = \underline{h}_{(t)}^{\perp} \underline{R}_y^{-1} \tag{15}$$

where the correlation matrix of the received vector is

$$\underline{R}_y = E\{\underline{y}\, \underline{y}^{\perp}\} = \underline{H}\, \underline{R}_x\, \underline{H}^{\perp} + \underline{R}_n = E_S / N_T\, \underline{H}\, \underline{H}^{\perp} + N_0 \underline{I}\,. \tag{16}$$

It follows that

$$\Lambda(d_n^t = +1) \sim \max_{\hat{x}^t \in X(d_n^t = +1)} \{\mathrm{Re}\left\{\hat{x}^{t^*} \underline{w}_{(t)}^T \left(2\underline{y} - \hat{x}^t\right)\right\}\}\,. \tag{17}$$

It can be shown that the MMSE array processor and the

[1] Superscript $^{\perp}$ denotes transpose conjugate, while ^{T} denotes transpose.

spatial decorrelator are identical when there is no feedback information, and the symbols are assumed to be i.i.d., and to have zero mean and constant amplitude [5].

An alternative method to combine the receive array is obtained with the ZF criterion [15]. No estimation of the noise variance is required although worse performance is achieved. The weights are obtained from (16) for $N_0 \to 0$.

Note that if the covariance/correlation matrices are singular, the weights can be obtained with pseudo-inverse methods.

D. Successive Interference Cancellation: SIC Processor.

Successive interference cancellation in the context of multiple transmit-receive antennas systems has been described in [6]. The basic concept is as follows. Detect a layer by deploying an MMSE or ZF array processor. Regenerate the corresponding signal. Subtract it from the received signals. Detect another layer assuming the presence of only the remaining layers. Continue till all layers are detected. Further, a performance improvement is achieved by ordering the detection process in a way such that we detect the layer that has the best signal-to-noise ratio.

This idea can be combined with turbo detection, if we simply do canceling using the decoder hard/soft outputs. As noted in [1] when coding is done separately on the layers, each layer can be fully detected and decoded without requiring the detection of the others. This might help to avoid error propagation. Further, soft interference cancellation can be obtained in a simple manner by subtracting the mean values of the symbols. The mean values are easily computed from the soft-outputs provided by a MAP-BCJR decoder.

Once we have decoded all layers, we can restart detection of a given layer by applying soft cancellation followed by maximal ratio combining. This is exactly what the simplified soft decorrelator does, assuming constant amplitude signals.

Finally, note that hard feedback decisions on the coded bits can be used in the MIMO APP calculator as shown in [12].

IV. DIVERSITY ORDER AND OUTAGE PROBABILITY

With N_T transmit N_R receive antennas and transmission over N_B bursts, the diversity resources are limited to $N_T N_R N_B$.

In a ST-BICM system with iterative decoding the maximum attainable diversity L' satisfies the bound (Singleton) [12]

$$L' \leq N_R \left(1 + \lfloor N_B (N_T - R/\log_2 M) \rfloor \right) \quad (18)$$

where M is the modulation order, and R is the transmission rate in Bits/s/Hz.

If coding is done independently on each layer, i.e. we deploy a T-BICM architecture the bound becomes

$$L' \leq N_R \left(1 + \lfloor N_B (1 - R/N_T/\log_2 M) \rfloor \right). \quad (19)$$

Therefore lower diversity levels are achievable with the T-BICM architecture than with the ST-BICM architecture for the same transmission rate and modulation order.

From a generalization of the Shannon capacity formula [7] the outage probability, i.e., the probability that the channel does not sustain the rate, is

$$P_{out}(C) = P[\prod_{l=1}^{N_B} \det(\underline{I} + \frac{E_S}{N_T N_0} \underline{H}(l) \underline{H}^{\perp}(l)) < 2^{C N_B}]. \quad (20)$$

We can also upper bound the outage probability by using Foschini capacity bound when $N_T \geq N_R$ [7] :

$$P_{out,FS}(C) = P[\prod_{l=1}^{N_B} \prod_{i=N_T-N_R+1}^{N_T} (1 + \frac{E_S}{N_T N_0} \chi_{2N_R}(i,l)) < 2^{C N_B}] \quad (21)$$

where $\chi_{2N_R}(i,l)$ is a sequence of chi-squared independent random variables with $2N_R$ degrees of freedom.

We will use (20)-(21) to bound the performance curves in Section V.

V. PERFORMANCE RESULTS

We consider both ST-BICM and T-BICM architectures with convolutional codes and QPSK with Gray mapping.

In ST-BICM we use one encoder. Its output bits are parsed into N_T streams and randomly interleaved into N_B (Rayleigh faded) bursts. The code polynomials have been obtained from computer search [12] and correspond to minimum constraint length codes that achieve full asymptotic spatial-temporal diversity for a given transmission rate and modulation order. For clarity we tabulate the polynomials in Table 1.

In T-BICM each layer is independently coded and randomly interleaved into N_B bursts.

Block error rate performance has been collected over 20000 coded blocks. Demapping has been performed using the APP calculator with soft feedback (APP-SF), with hard feedback (APP-HF), with exact feedback (APP-EF), and using the simplified soft-in soft-out spatial decorrelator (SOFT-DCR). The channel matrix and noise variance are assumed known.

	2 Tx Antennas 2 Bits/s/Hz		4 Tx Antennas 2 Bits/s/Hz	4 Tx Antennas 4 Bits/s/Hz	
	N_B=1	N_B=4	N_B=2	N_B=1	N_B=2
K	**3**	**4**	**4**	**3**	**4**
Poly	**(5,7)**	**(11,15)**	**(7,11,13,15)**	**(5,7)**	**(11,15)**
L	**2**	**5**	**7**	**3**	**5**
X^2_{min}	**4.90**	**2.00**	**3.91**	**3.18**	**2.00**

Table 1. Convolutional codes. *K*: constraint length. $L=L'/N_R$: transmit diversity order. X^2_{min}: minimum squared product distance with QPSK and Gray mapping.

A. Transmission at 2 Bit/s/Hz with 2 Tx and 2 Rx Antennas

In Fig. 4 we consider a ST-BICM architecture (Fig. 2) for transmission at 2 bit/s/Hz with 2 Tx and 2 Rx antennas. The block of information bits has length 240. The convolutional codes that we deploy are according to Table 1.

The various detectors perform similarly at the 6-th decoding iteration (IT=6 in the legend). Note that APP-SF converges to the APP-EF performance, showing the high reliability of the soft feedback. The performance is improved with transmission over 4 bursts due to the increased diversity.

Full spatial and temporal diversity is exploited. As reported in [12] these ST-BI convolutional codes are the optimal minimum constraint length codes for transmission over 1 and 4 bursts. Further, it is possible to improve performance by choosing bit-mappings with larger Euclidean distance [12].

Note that we are only 2 dB in Fig. 4.A and 3-4 dB in Fig. 4.B from the outage probability although we deploy a simple

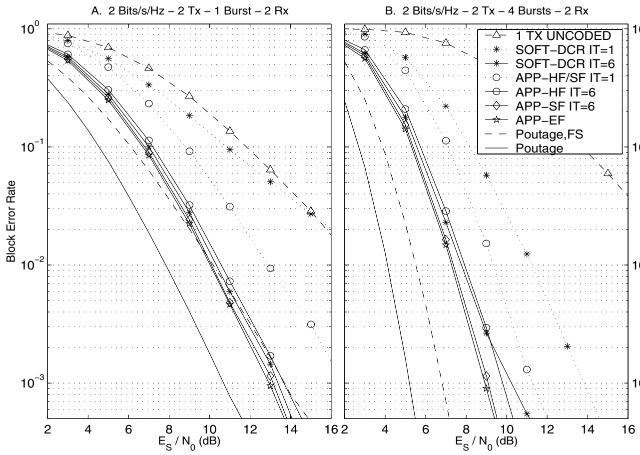

Fig.4. A) ST-BICM at 2 Bits/s/Hz over 1 burst with 2Tx-2Rx antennas. B) Transmission at 2 Bits/s/Hz over 4 independently faded bursts. Blocks of 240 information bits.

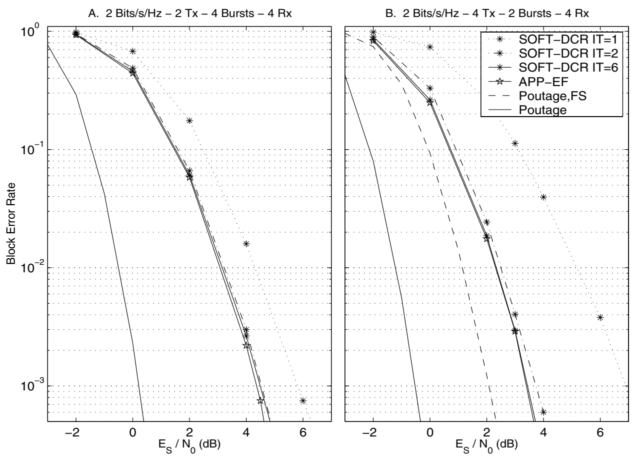

Fig.5. A) ST-BICM at 2 Bits/s/Hz over 4 bursts with 2Tx-4Rx antennas.
B) Transmission at 2 Bits/s/Hz with 4Tx-2Rx antennas over 2 bursts. Blocks of 240 bits.

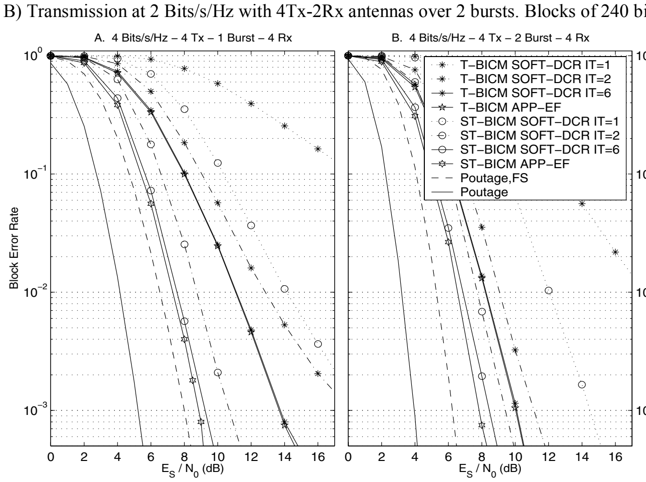

Fig.6. A) ST-BICM and T-BICM at 4 Bits/s/Hz with 4Tx-4Rx antennas over one burst. B) Transmission at 4 Bits/s/Hz over 2 independently faded bursts. Blocks of 480 bits.

code. In Fig. 1.A we exceed the Foschini outage bound (21).

B. Transmission at 2 Bit/s/Hz with 2-4 Tx and 4 Rx Antennas

In Fig. 5 we consider a ST-BICM scheme for transmission at 2 bit/s/Hz with 2 and 4 Tx antennas. The number of Rx antennas is 4. The block size is 240 bits. When using 2 Tx antennas the information is interleaved across 4 independently faded bursts. When using 4 transmit antennas interleaving is across 2 bursts. With QPSK modulation the maximum attainable transmit diversity $L=L'/N_R$ is respectively 5 and 7. Fig. 5 confirms that with (N_T=4, N_B=2) we have better performance than with (N_T=2, N_B=4). In both cases the turbo SOFT-DCR converges to the APP-EF curves.

C. Transmission at 4 Bit/s/Hz with 4 Tx and 4 Rx Antennas

In Fig. 6 we consider transmission at 4 bit/s/Hz with 4 Tx and 4 Rx antennas. We compare the performance of T-BICM and ST-BICM using the turbo soft spatial decorrelator. The blocks have length 480 information bits. In both Fig. 6.A, and B better performance is obtained with ST-BICM due to the increased diversity level. However, the advantage diminishes when higher temporal diversity is available as Fig. 6.B shows.

VI. CONCLUSIONS

We have presented a space-time bit-interleaved coded approach with turbo decoding based on several demapping strategies among which soft-in soft-out array decorrelation.

The deployment of 'good' ST-BI convolutional codes with turbo array processing allows for the exploitation of the spatial and temporal diversities in systems that combine, for instance, multiple transmit antennas and slow frequency hopping. Block-error-rate performance is within 2-4 dB from the outage probability and 1-2 dB from the bound on outage probability obtained from Foschini capacity lower bound.

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