Performance of Space-Time Bit-Interleaved Codes in Fading Channels with Simplified Iterative Decoding

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Abstract

The performance of space-time bit-interleaved codes over block fading channels is considered. Decoding is based on iterative demapping and decoding. The demapper delivers soft information to the decoder, and accepts feedback from the decoder. Schemes with both soft and hard feedback are described. In particular, we consider the deployment of bit interleaved block/convolutional codes with memoryless mappers. Each block of coded information is transmitted over a number of bursts that experience independent fading. Code design criteria are devised through the analysis of the asymptotic pairwise error probability (i.e., with perfect extrinsic knowledge), and of the maximum attainable diversity bound. Several coding schemes that achieve full asymptotic diversity with convolutional codes of minimum complexity are given. Performance results from simulations show that convergence to the asymptotic error rate curves is achieved.

1. Introduction

In this paper we consider a space-time bit-interleaved coded wireless system that deploys multiple transmit-receive antennas. Coding and modulation are kept separated by a bit-interleaver, which extends the bit-interleaved coded modulation concept [1] from SISO channels to MIMO channels. This coding approach was considered in [2] and [3] over both static and fully interleaved fading channels. It differs from the space-time trellis coding [5] approach, and the orthogonal block coding [6] approach, since coding and modulation are not combined in a single entity.

In this work we consider transmission of coded information by multiple antennas over a small number of bursts that experience uncorrelated frequency non-selective fading. This is a realistic model for coded systems that deploy slow frequency hopping such as the GSM/EDGE (enhanced data rates for GSM evolution) system. In more detail, a block of coded bits, generated by a block or convolutional encoder, is appropriately interleaved, parsed into a number of antenna bit streams, and formatted into a number of bursts. Transmission by the antennas is simultaneous after mapping of the bits into M-PSK/M-QAM complex symbols. The bursts are transmitted sufficiently apart in time or frequency, such that, under the hypothesis of low mobility, the fading is static over a given burst, but uncorrelated among distinct bursts.

Although random symbols are transmitted over overlapping and independent fading channels, reliable decoding is possible through an iterative decoding procedure. First the received samples are demapped by computing soft information (e.g., a posteriori probabilities) of the coded bits. Then, after de-interleaving, the soft information is delivered to the decoder. By exploiting the redundancy of the coded bits, the decoder can provide feedback information on the coded bits to the demapper. Therefore, multiple demapping and decoding stages can be run following the turbo demapping concept [7], [8]. We consider either the case of providing soft or hard feedback. In the former case a MAP/BCJR decoder [9], [10] is required. In the latter case only a conventional hard output decoder (e.g., Viterbi decoder) is needed, which greatly simplifies complexity.

The deployment of bit-interleaved codes in a layered space-time architecture was considered also in [11]. However, the demapping approach herein considered is based on optimal joint demapping of the overlapping signals which allows decoding with single receive diversity. The application of interleaved space-time codes over frequency selective (i.e., ISI) fading channels has been addressed in [3] and [4], where decoding based on optimal iterative equalization is proposed.

We gain insight into the code construction criteria by studying the asymptotic pairwise error probability that is achieved when perfect extrinsic knowledge is available, and by extending the Singleton bound [12] to the multiple transmit antenna scenario under consideration. This bound shows that the maximum attainable transmit diversity is limited to a value lower or equal to the product of the number of transmit antennas and the number of independent bursts. It is a function of the code rate and modulation order. Further, we show that the distance properties of certain bit sub-sequences, and the bit-to-symbol mapping rule, play a key role in the maximization of the diversity gain and the coding advantage.

From computer search, several convolutional codes that achieve full asymptotic diversity with minimum constraint length have been found. The performance of several of these interleaved codes is evaluated by simulation.

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2. System Model

2.1. Transmitter and Channel Model

A block of information bits (Fig.1) $\underline{b} = [b_1 ... b_{N_b}]$ is either block or convolutionally encoded into a block of coded bits $\underline{c} = [c_1 ... c_{N_c}]$. The block of coded bits is bit-interleaved and parsed into N_T blocks $\underline{d}^t = [d_1^t ... d_{N_d}^t]$, $t=1,...,N_T$. Each antenna block is arranged into N_B bursts of $N_D=N_d/N_B$ bits each, $\underline{d}^{t,l} = [d_1^{t,l} ... d_{N_D}^{t,l}]$. The bits of each burst are memoryless mapped into complex symbols belonging to the M-PSK or M-QAM signal set with average power one.

Let $N = \log_2 M$ be the number of bits per symbol, then the complex symbol that is transmitted by antenna t at time instant kT on burst l, is a function of the mapping rule

$$x_k^{t,l} = \mu^t(d_{Nk+1}^{t,l},...,d_{Nk+N}^{t,l}) = \mu^t(\underline{d}_k^{t,l}) . \quad (1)$$

Assuming to deploy 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, $r=1,...,N_R$, can be expressed as

$$y_k^{r,l} = \sqrt{E_s} \sum_{t=1}^{N_T} h^{r,t,l} x_k^{t,l} + n_k^{r,l} . \quad (2)$$

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 (2) $n_k^{r,l}$ is the AWGN contribution that is assumed to have mean zero, variance N_0, and to be independent across the receive antennas.

2.2. Iterative Decoder

Decoding is based on concatenating in an iterative fashion a demapper with a decoder (Fig. 2). 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 $N_B N_S$ samples from N_R antennas. 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. Further, we assume it can provide feedback information on the coded bits to the demapper. Such feedback can be either soft or hard. In the former case a MAP/BCJR [9], [10] decoder can be deployed. In the latter case a conventional decoder (e.g., Viterbi decoder) can be used.

The feedback information from the decoder can be re-used in a new demapping stage to improve the quality of the soft information provided by the demapper. This helps to de-couple the signals from overlapping channels.

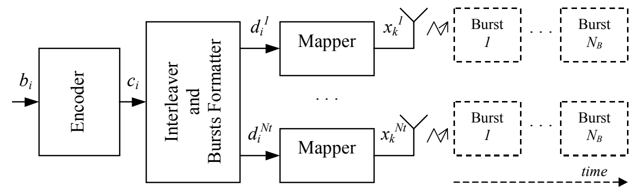


Fig.1. Bit-interleaved space-time coded transmitter.

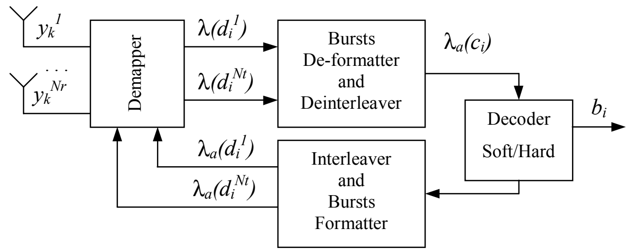


Fig.2. Iterative decoder. Feedback from the decoder can be soft or hard.

3. Demapping

The demapping algorithm is a function of whether soft or hard feedback is available. In either case the demapper computes the a posteriori probabilities of the coded bits by observing the channel samples from the N_R antennas

$$\lambda(d_{Nk+i}^{t,l} = \pm 1) = P[d_{Nk+i}^{t,l} = \pm 1 \,|\, y_k^{1,l},..., y_k^{N_R,l}] \quad (3)$$

for all $t = 1,..., N_T$, $l = 1,..., N_B$, $k = 1,..., N_S$, $i = 1,..., N$. We further assume to have perfect knowledge of the channel state information.

It is convenient to operate in the logarithm domain by exchanging between the demapper and decoder log-likelihoods or log-likelihood ratios, respectively defined as

$$\Lambda(d_{Nk+i}^{t,l} = \pm 1) = \ln \lambda(d_{Nk+i}^{t,l} = \pm 1) \quad (4)$$

$$L(d_{Nk+i}^{t,l}) = \Lambda(d_{Nk+i}^{t,l} = +1) - \Lambda(d_{Nk+i}^{t,l} = -1) . \quad (5)$$

3.1. Demapper with Soft Feedback

Let $\lambda_a(\tilde{d}_{Nk+i}^{t,l}) = P_a[d_{Nk+i}^{t,l} = \tilde{d}_{Nk+i}^{t,l}]$ be the a priori probability that bit $d_{Nk+i}^{t,l}$ equals $\tilde{d}_{Nk+i}^{t,l}$, then, the demapper (see also [2]) computes the a posteriori probabilities according to (neglecting constant factors)

$$\lambda(d_{Nk+i}^{t,l} = \pm 1) \sim$$

$$\sim \sum_{\tilde{\underline{d}}_k^l \in D(\pm 1)} e^{-\frac{1}{N_0} \sum_{r=1}^{N_R} \left| y_k^{r,l} - \sqrt{E_S} \sum_{p=1}^{N_T} h^{r,p,l} \mu^p(\tilde{\underline{d}}_k^{p,l}) \right|^2} \prod_{p=1}^{N_T} \prod_{n=1}^{N} \lambda_a(\tilde{d}_{Nk+n}^{p,l}) \quad (6)$$

where, for instance, $D(+1)$ is the set of all possible bit vectors $\tilde{\underline{d}}_k^l = [\tilde{d}_{Nk+1}^{1,l} ... \tilde{d}_{Nk+N}^{1,l} ... d_{Nk+i}^{t,l} = +1 ... \tilde{d}_{Nk+N}^{N_T,l}]$ that have bit $d_{Nk+i}^{t,l} = +1$. By taking the logarithm of (6), the a posteriori log-likelihoods can be approximated as

$$\Lambda(d_{Nk+i}^{t,l}=\pm 1)\cong \max_{\underline{\tilde{d}}_k^l\in D(\pm 1)}\left\{\frac{-1}{N_0}\sum_{r=1}^{N_R}\left| y_k^{r,l}-\sqrt{E_S}\sum_{p=1}^{N_T}h^{r,p,l}\mu^p(\underline{\tilde{d}}_k^{p,l})\right|^2 + \sum_{p=1}^{N_T}\sum_{n=1}^{N}\Lambda_a(\tilde{d}_{Nk+n}^{p,l})\right\} \quad (7)$$

The a priori probabilities/log-likelihoods of the coded bits, are approximated with the soft information provided by a soft-out decoder in the previous decoding iteration. Equally likely bits are assumed at the first demapping step, such that the product of the a priori probabilities is a constant.

To minimize the correlation with previously computed information, extrinsic information is exchanged between the demapper and decoder [2], e.g., in (6) we neglect the a priory probability of the bit under consideration.

Note that samples from multiple receive antennas are combined in a maximal ratio combining fashion. Further, the soft outputs are computed by considering all possible combination of simultaneously transmitted symbols (i.e. joint demapping of multiple transmitted signals).

3.2. Demapper with Hard Feedback

When hard feedback is available from the decoder, the a priori probabilities on the coded bits can assume only the value 0 or 1. Therefore, (6) and (7) simplify to

$$\lambda(d_{Nk+i}^{t,l}=\pm 1)\sim e^{-\frac{1}{N_0}\sum_{r=1}^{N_R}\left| y_k^{r,l}-\sqrt{E_s}\sum_{p=1}^{N_T}h^{r,p,l}\mu^p(\underline{\tilde{d}}_k^{p,l})\right|^2} \quad (8)$$

$$\Lambda(d_{Nk+i}^{t,l}=\pm 1)\cong -\frac{1}{N_0}\sum_{r=1}^{N_R}\left| y_k^{r,l}-\sqrt{E_s}\sum_{p=1}^{N_T}h^{r,p,l}\mu^p(\underline{\tilde{d}}_k^{p,l})\right|^2 \quad (9)$$

with $\underline{\tilde{d}}_k^{p,l}=[\tilde{d}_{Nk+1}^{p,l}...\tilde{d}_{Nk+N}^{p,l}]$ and $\tilde{d}_{Nk+n}^{p,l}=+1$ if $P_a[d_{Nk+i}^{p,l}=+1]=1$ and $\tilde{d}_{Nk+n}^{p,l}=-1$ otherwise, for all $p\neq t \wedge n\neq i$. Basically, the bits in the vectors $\underline{\tilde{d}}_k^{p,l}$ are determined by the codeword the decoder decides in favor of, with the exception of bit $d_{Nk+i}^{t,l}$ that is set to ± 1.

At the first demapping iteration we assume equally likely bits. Thus, we apply (6) or (7) and we neglect the a priori term. In the following iterations, we apply (8) or (9) by assuming a priori knowledge of all bits except the one for which we are computing the a posteriori probability.

If we deploy convolutional codes, decoding can be performed with the Viterbi algorithm, and the hard feedback is obtained by re-encoding the information bit sequence we decide in favor of.

4. Performance Analysis

The exact performance analysis is complicated due to iterative processing. We can gain insight by assuming that at the final iteration the feedback from the decoder to the demapper is exact [2]. A similar approach was followed also in [8] for the analysis of iterative decoding in single-input single-output channels. Bounds on the maximum attainable diversity (asymptotic diversity) can be derived by studying the pairwise error probability (the probability of transmitting the coded bit sequence $\underline{c}$ but erroneously deciding in favor of $\underline{\hat{c}}$), or extending to the multiple transmit antennas scenario the Singleton bound [12].

A Chernoff bound on the pairwise error probability was derived in [2] under the exact feedback assumption, and considering the transmission of coded information over a single burst that experiences either static or completely temporally uncorrelated fading. Here, we generalize those results by considering transmission over a number of independently faded bursts each experiencing static fading.

4.1. Asymptotic Pairwise Error Probability

The pairwise error probability is bounded as

$$P(\underline{c}\rightarrow\underline{\hat{c}})\leq\prod_{t=1}^{N_T}\prod_{l=1}^{N_B}\left(1+\frac{E_S}{4N_0}\sum_{n=1}^{N}d_E^{t,l,n}(\underline{c},\underline{\hat{c}})\right)^{-N_R} \quad (10)$$

with

$$d_E^{t,l,n}(\underline{c},\underline{\hat{c}})=\sum_{k=1}^{N_S}d_E^{t,l,n}(k) \quad (11)$$

$$d_E^{t,l,n}(k)=|\mu^t(\hat{d}_{Nk+1}^{t,l},...,d_{Nk+n}^{t,l},...,\hat{d}_{Nk+N}^{t,l})-\mu^t(\hat{d}_{Nk+1}^{t,l},...,\hat{d}_{Nk+n}^{t,l},...,\hat{d}_{Nk+N}^{t,l})|^2. \quad (12)$$

Eq. (12) gives the normalized squared Euclidean distance between the *k-th* constellation symbol that is transmitted by antenna *t*, over burst *l*, and the symbol that may differ only in bit $d_{Nk+n}^{t,l}$. It follows that (11) is the squared Euclidean distance between the symbol sequences associated to the pairwise error event, computed only over the sub-portion transmitted by antenna *t* on burst *l*, and assuming that they might differ in one bit at the time. Note that there is a one to one correspondence between coded bit sequences $\underline{c}$ and coded-interleaved bit sequences $\underline{d}$.

Let $d_{E,\min}^{t,l,n}$ be the minimum squared Euclidean distance between any pair of constellation symbols that differ only on bit in position *n* (the minimum being computed for all possible other bits). Further, let $d_H^{t,l,n}(\underline{c},\underline{\hat{c}})$ be the Hamming distance between the codewords $\underline{c}$ and $\underline{\hat{c}}$ (computed over the bits transmitted on antenna *t*, burst *l*, and bit position *n*) then (10) can be further bounded as follows

$$P(\underline{c}\rightarrow\underline{\hat{c}})\leq\prod_{t=1}^{N_T}\prod_{l=1}^{N_B}\left(1+\frac{E_S}{4N_0}\sum_{n=1}^{N}d_{E,\min}^{t,l,n}d_H^{t,l,n}(\underline{c},\underline{\hat{c}})\right)^{-N_R}. \quad (13)$$

For high SNRs (13) can be written as

$$P(\underline{c}\rightarrow\underline{\hat{c}})\leq(E_S/4N_0)^{-LN_R}\chi^2(\underline{c},\underline{\hat{c}})^{-LN_R} \quad (14)$$

with, $$\chi^2(\underline{c},\hat{\underline{c}}) = \Big(\underbrace{\prod_{t=1}^{N_T}\prod_{l=1}^{N_B}}_{d^{t,l}(\underline{c},\hat{\underline{c}})\neq 0} d^{t,l}(\underline{c},\hat{\underline{c}})\Big)^{1/L} \tag{15}$$

and $L \le N_T N_B$ being the number of nonzero $d^{t,l}(\underline{c},\hat{\underline{c}}) = \sum_{n=1}^{N} d_{E,\min}^{t,l,n} d_H^{t,l,n}(\underline{c},\hat{\underline{c}})$, i.e., nonzero bit squared Euclidean distances between the portions of the two codewords that are transmitted by antenna t on burst l.

From (14), we can see that a diversity gain of LN_R and a coding advantage of $\chi^2(\underline{c},\hat{\underline{c}})^{LN_R}$ is achieved over an uncoded system with single transmit and single receive diversity.

4.2. Singleton Bound

Under the exact feedback assumption the following bound on the maximum transmit diversity L applies (Singleton bound [12])

$$L \le 1 + \lfloor N_B (N_T - R/\log_2 M) \rfloor \tag{16}$$

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

5. Space-time Bit-interleaved Code Construction Guidelines

Guidelines for the design of an interleaved space-time coded system can be derived from the analysis of Section 4. Although the analysis is based on the assumption of exact feedback, extensive computer simulations have shown that asymptotically the system with iterative decoding behaves as predicted by theory (Section 6). In particular with soft feedback and few decoding iterations the block error rate curves practically converge to the curves obtained with perfect extrinsic knowledge (referred to as block error rate floor) for moderate SNRs. With hard feedback we have found that the performance is improved by running multiple decoding iterations. However, due to error propagation, the convergence to the error rate floor, as a function of the SNR, is much slower.

From (14), a maximum diversity order of LN_R can be achieved. A necessary condition for full transmit diversity $L=N_BN_T$ is that any two codewords have nonzero Hamming distance when computed over the portion transmitted on a given burst and given antenna, for all bursts and antennas. Clearly, this Hamming distance depends upon the encoder, and the bit-interleaver/parser.

From (16), the shape of the constellations, and the mapping rule, do not affect the maximum attainable diversity. For a given transmission rate, if we increase the modulation order we can increase the attainable transmit diversity up to the limit given by N_BN_T.

From (14) and (15), the coding gain depends upon the constellation shape, the bit-to-symbol mapping rule, and the Hamming distances $d_H^{t,l,n}$ that are a function of the encoder and the interleaver. Fixed the constellation shape, the coding gain can be increased by choosing an appropriate mapping.

5.1. Space-time Bit-interleaved Convolutional Codes

We herein consider the deployment of convolutional codes, which can be interpreted as very long block codes that verify the bound in (16).

The encoded bits are first parsed in the natural order into N_T (i.e., number of transmit antennas) blocks. Then, each antenna block is further partitioned into N_B (i.e., number of bursts) blocks, inside which random interleaving is applied before mapping and transmission.

We have fixed the number of transmit antennas, the number of bursts per block, and the modulation order. Then, we have searched for codes that achieve full asymptotic diversity, i.e., with perfect extrinsic knowledge, with minimum constraint length (complexity). Further, Gray mapping is assumed. Some of these codes are listed in Table 1. Note that some of these codes achieve full diversity but have product distance $\chi^2_{\min}$<1. A mapping with larger $d_{E,\min}^{t,l,n}$ than Gray mapping, as well as a higher constraint length, shall increase such a product distance. Several codes in Table 1 coincide with the ones reported in [12] for binary modulation with single transmit antenna and transmission over N_BN_T independent bursts.

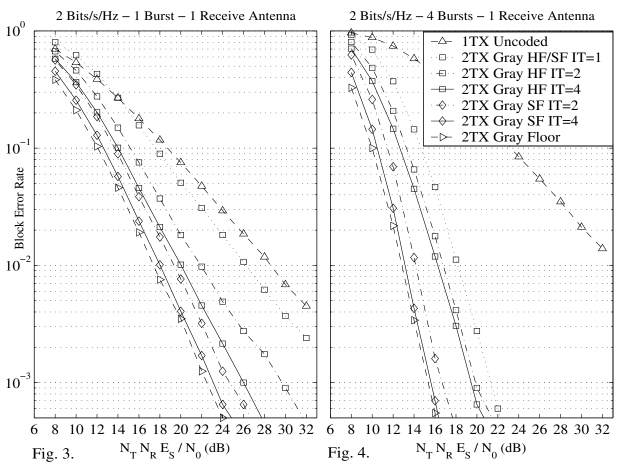
		2 TX Antennas 2 Bits/s/Hz		**4 TX Antennas 2 Bits/s/Hz**	**4 TX Antennas 4 Bits/s/Hz**	
	M - r	**4 – 1/2**	**8 – 1/3**	**4 – 1/4**	**4 – 1/2**	**16 – 1/4**
N_B=1	***K***	3	3	3	3	3
	Poly	(5,7)	(5,7,7)	(5,5,7,7)	(5,7)	(5,5,7,7)
	L	2	2	4	3	4
	X^2_{min}	4.90	2.27	4.90	3.18	0.98
N_B=2	***K***	3	3	4	4	4
	Poly	(5,7)	(5,7,7)	(7,11,13,15)	(11,15)	(7,11,13,15)
	L	3	3	7	5	7
	X^2_{min}	3.18	2.13	3.91	2.00	0.78
N_B=3	***K***	3	3			
	Poly	(5,7)	(11,15,17)			
	L	4	5			
	X^2_{min}	2.83	1.11			
N_B=4	***K***	4	4			
	Poly	(11,15)	(13,15,17)			
	L	5	6			
	X^2_{min}	2.00	0.89			

Table 1. Minimum constraint length K convolutional codes that achieve maximum asymptotic diversity L, and maximize the minimum product distance, $\chi^2_{\min}(\underline{c},\hat{\underline{c}})$. Code rates *r=1/2*, *1/3*, *1/4*, transmission over N_B bursts with *2* and *4* transmit antennas, modulation *M=4-PSK*, *8-PSK*, *16-QAM* with Gray mapping. Polynomial notation as in [13].

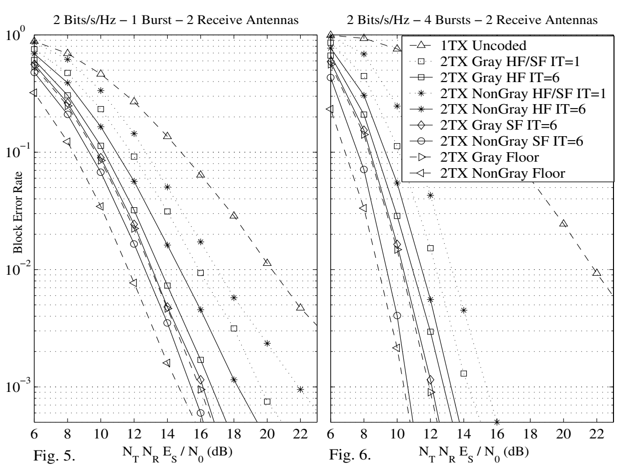
6. Simulation Results

Block error rate performance has been evaluated through simulations. We report only the results corresponding to the coding schemes in Table 1 for transmission at 2 Bits/s/Hz over one and four bursts, with 4-PSK modulation and two transmit antennas. The block length is 240 information bits with tail termination. The encoded block of bits is first partitioned into the bursts. Then, each burst is randomly interleaved. Further, distinct bursts use independent interleavers. Demapping is based on the approximations in (7) and (9) with perfect CSI. Soft-output decoding is based on a Max-Log-MAP decoder [10].

In Fig. 3, we assume Gray mapping, transmission over one burst, and single receive diversity. The performance is greatly improved with iterative decoding with both soft (SF) and hard (HF) feedback. At the fourth decoding iteration, the gain in BLER at 10^{-2} over the uncoded system with single transmit antenna, is 8.5 dB with HF, and 10.5 dB with SF. Further, SF is within 0.5 dB from the BLER floor that is obtained assuming the same coding scheme, and decoding based on perfect extrinsic knowledge (i.e., of all coded bits except the one for which we compute the a posteriori probability).



Block error rate versus average signal to noise ratio with double transmit and single receive diversity for transmission at 2 Bits/s/Hz of blocks of 240 information bits over one burst (Fig.3) or four independently faded bursts (Fig.4). Coding schemes of Table 1 with Gray mapping. Up to four decoding passes for both soft (SF) and hard (HF) feedback.



As in Fig. 3 and Fig. 4, with: double receive diversity, both Gray and Non Gray mapping, up to six decoding passes for both soft and hard feedback.

In Fig. 4, transmission is over four bursts. Due to time diversity the performance is greatly improved over the single burst transmission of Fig. 3. Again, the SF curve converges to the BLER floor. However, a larger gap is found with HF. Although not reported, we have found very small improvements when deploying non-Gray mapping and single receive diversity.

In Fig. 5 and Fig. 6 we consider double receive diversity with both Gray and non-Gray mapping. At the sixth decoding iteration non-Gray mapping yields better performance than Gray mapping if SF is deployed. Considering non-Gray mapping, SF is within 1 dB of the BLER floor in Fig. 5, while it exhibits convergence to the floor in Fig. 6 even for the SNR range under consideration.

7. Conclusions

We have considered coding with multiple transmit antennas over block fading channels, when deploying bit interleavers with iterative decoding based on both soft and hard feedback. Bounds on the maximum attainable diversity as well as an asymptotic expression for the pairwise error probability have been derived. From computer search, several convolutional codes with bit interleavers that achieve full asymptotic diversity with minimum complexity have been found. Simulation results show that with soft feedback and few iterations the block error rate curves converge to the error rate floor at moderate SNR levels. Hard feedback decoding is less complex but has slower convergence. Further, performance can be improved by choosing appropriate bit-mappings. Finally, we have found that as the information block length increases, and consequently with longer interleavers, the convergence to the BLER floor is achieved at lower SNRs.

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