On Turbo Equalization of Interleaved Space-Time Codes

Andrea M. Tonello

University of Padova - DEI Department of Electronics
Via Gradenigo 6/A - 35131 - Padova – Italy

Lucent Technologies Bell Labs - Whippany Laboratory
67 Whippany Rd - Whippany - NJ 07981 – USA

e-mail: tonello@dei.unipd.it

Abstract - **Space-time coding over frequency selective time-varying fading channels is considered. The approach is based on deploying bit/symbol interleaved space-time codes. Decoding is based on the iterative concatenation of a soft-in soft-out MIMO equalizer with a soft-in soft-out decoder. The equalizer structure comprises a bank of matched filters followed by a maximum a posteriori processor with an appropriate metric. It does not require multiple receive antennas. Methods to reduce its complexity are devised. Finally, performance results are shown for ST bit-interleaved convolutional codes designed for the EDGE air interface with two transmit - single receive antennas.**

I. INTRODUCTION

In this paper we address the decoding problem of space-time codes deployed over frequency selective fading channels that introduce inter-symbol interference (ISI). In particular we consider a space-time interleaved coded system. Interleaving takes place at the output of the space-time encoder. In principle we could use a space-time trellis code [9] followed by symbol interleaving. However, it is preferable to use space-time bit-interleaved codes [10]-[14] where bit interleaving separates coding from modulation. This allows for a better exploitation of the spatial, temporal and frequency diversity resources available in the system. Further, the system can be treated as a serially concatenated coded system [4]. Coding, interleaving, and modulation can be done independently on each transmit layer, or more in general coding and interleaving take place across layers [1], [14].

Optimal maximum likelihood decoding requires taking into account the structure of the space-time encoder, the interleaver, and the ISI channel. Here we follow the turbo decoding approach through the iterative concatenation of a space-time equalizer and a decoder [11], [12].

In this contribution we are interested on space-time coded architectures that do not necessarily require multiple receive antennas. This can be the scenario in applications where the receiver is a terminal that has to fulfill stringent cost and size requirements. Therefore, we investigate joint equalization of multiple transmitted signals [16].

For single-input single-output inter-symbol interference channels, two well-known formulations of the MLSE/MAP (maximum likelihood/maximum a posteriori) receiver have been developed. In the first formulation, by Forney, a sequence estimation algorithm that uses the Euclidean distance metric follows a whitening matched filter [8]. In a second formulation, by Ungerboeck, the sequence estimation algorithm operates directly on the matched filter output using a modified metric [15]. The equivalence of the two receivers has been shown in [5].

We consider the Ungerboeck's approach and extend it to the space-time coded scenario that we consider. This yields an equalizer structure that comprises a bank of matched filters followed by a processor that runs the maximum a posteriori algorithm with a modified metric [12]. The MAP equalizer is formulated for both time-varying and static MIMO channels.

Recently, turbo space-time processing has independently been proposed in [3] for decoding of space-time block/trellis codes with symbol interleavers, and in [1] for decoding of space-time bit-interleaved codes. However, both [1] and [3] follow the approach of deploying a spatial-temporal whitening multiple-input multiple-output (MIMO) prefilter. Further, the layered space-time architecture described in [1] requires multiple receive antennas. Instead, in the approach that is proposed in this paper, the spatial-temporal whitening operation as well as multiple receive antennas are not required.

Although this equalization approach removes the constraint of having multiple receive antennas, it may suffer from high complexity. The equalizer has a number of states that grows exponentially with the number of transmit antennas and the channel constraint length, while the number of transitions to/from each state grows exponentially with the number of transmit antennas. In order to lower such a complexity we investigate the application of techniques that allow for reducing the number of states and transitions in the equalizer.

This space-time coding/decoding approach is evaluated over the EDGE (enhanced data rates for GSM evolution) air interface [18]. Simulation results show that ST bit-interleaved convolutional codes provide high gains when deploying two transmit antennas while keeping single receive diversity.

II. SPACE-TIME BIT-INTERLEAVED CODED SYSTEM

We consider a space-time bit-interleaved coded system [10]-[14] as depicted in Fig. 1. An encoder encodes a sequence of bits $\{b_i\}$ into a coded bit sequence $\{c_i\}$. It can be a block, a convolutional, or a turbo encoder. Here, we consider convolutional coding.

The coded bit sequence is appropriately interleaved, and parsed into N_T branch sequences $\{d_i^a\}, a = 1,...,N_T$. Each branch bit sequence is mapped into a complex symbol sequence $\{x_k^a\}$ e.g., with a M-PSK/M-QAM mapper. After pulse shaping and RF modulation the signals are simultaneously transmitted by the transmit antennas.

The bit interleaver is deployed for the following reasons. First, it is used to remove the correlation in the sequence of

0-7803-7005-8/01/$10 © 2001 IEEE

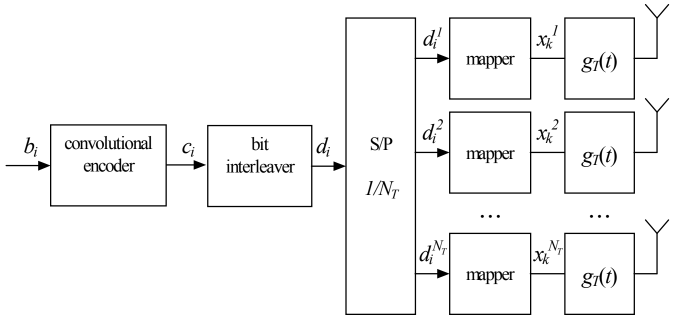

Fig. 1. Space-time bit-interleaved coded system.

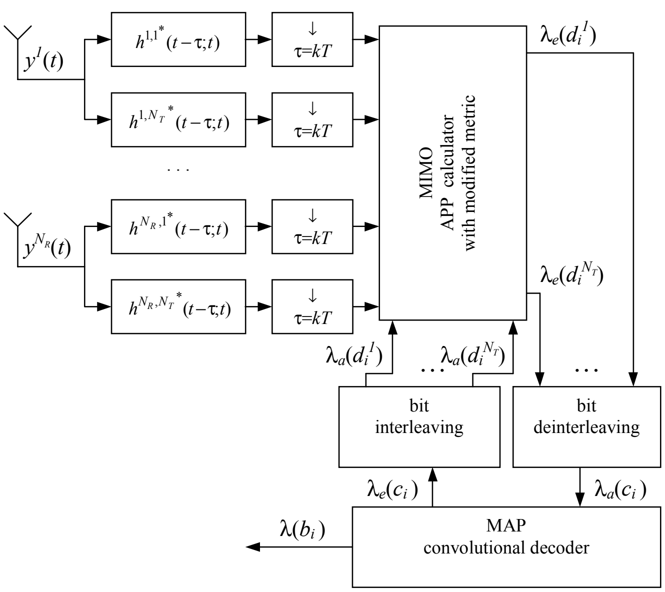

Fig. 2. Iterative (turbo) equalization and decoding receiver structure for space-time bit-interleaved coded modulation.

convolutionally encoded bits, which is an essential condition for turbo decoding. Second, it can be used to de-correlate the fading channel and maximize the diversity order of the system. Finally, when dealing with channels with memory, the overall system can be treated as a serially concatenated and interleaved coded system [4].

Let $g_T(t)$ be the pulse shaping filter, then, the complex signal transmitted by antenna a is

$$x^a(t) = \sum_{k=-\infty}^{+\infty} x_k^a g_T(t-kT) . \quad (1)$$

Assuming to deploy an array of N_R receive antennas, Fig. 2, the base band complex received signal of receive antenna b is

$$y^b(t) = \sum_{k=-\infty}^{+\infty} \sum_{a=1}^{N_T} x_k^a h^{b,a}(t-kT;t) + \eta^b(t) \quad (2)$$

where $h^{b,a}(\tau;t)$ is the convolution, at time *t*, of the pulse shaping filter with the time-varying channel impulse response $g_{ch}^{b,a}(\tau,t)$ of the link between receive antenna b, and transmit antenna a.

The channel is assumed in general to be time-varying and frequency selective, such that inter-symbol interference may arise. The correlation between the fading channels of distinct antenna links is a function of the antenna configuration and the propagation environment. It is often assumed to be zero. The noise processes $\eta^b(t)$ are assumed to be stationary zero-mean white Gaussian, with power spectral density $2N_0$, and independent from receive branch to receive branch.

Guidelines for the design of space-time bit-interleaved codes that achieve full spatial/temporal diversity are described in [10], [13], and [14].

III. Iterative (Turbo) Decoding

Concatenating (Fig. 2) in an iterative fashion a space-time soft-out equalizer with a soft-in soft-out convolutional decoder performs decoding. The equalizer delivers to the decoder the a posteriori probabilities of the coded bits[1] by observing the received signals over a finite time interval, as it is described in Section IV. The decoder can be implemented with the maximum a posteriori algorithm (MAP) [2], and provides new and improved a posteriori probabilities of the coded bits by exploiting the redundancy of the code. According to the turbo principle, extrinsic information ($\lambda_e(d_i^a)$, $\lambda_e(c_i)$ in Fig. 2) has to be exchanged between the equalizer and decoder in order to minimize the correlation with previously computed information [6]. The extrinsic information on a given bit is defined as the a posteriori probability of that bit by taking into account all transmitted bits except the bit under consideration. If we operate in the logarithm domain, then the extrinsic information is obtained by subtracting the input log-likelihoods from the output log-likelihoods. In the final iteration the decoder provides the decoded information bit sequence.

IV. Space-Time MIMO MAP Equalizer

The goal of the equalizer is to compute the a posteriori probabilities $\lambda(d_i^m = \pm 1)$ of each coded and interleaved bit d_i^m, $m = 1,...,N_T$, observing the received signals $y^b(t)$, $b = 1,...,N_R$, over a finite time interval I, i.e.,

$$\lambda(d_i^m = \pm 1) = P[d_i^m = \pm 1 \,|\, \underline{y}(t), t \in I] \quad (3)$$

where $\underline{y}(t) = \{y^1(t),...,y^{N_R}(t)\}$.

Now, consider the a posteriori probability of a given coded bit sequence $\{\hat{d}_i\} = \{\hat{d}_i^a, a = 1,...,N_T\}$

$$P[\{\hat{d}_i\} \,|\, \underline{y}(t)] = P[\underline{y}(t) \,|\, \{\hat{d}_i\}] P[\{\hat{d}_i\}] / P[\underline{y}(t)] . \quad (4)$$

Since there is a one to one correspondence between the

[1] The operations can take place in the logarithm domain by exchanging a posteriori log-likelihoods between the equalizer and the decoder.

sequence of coded bits $\{\hat{d}_i\}$ transmitted over the interval I, and the sequence of symbols $\{\underline{\hat{x}}_k\} = \{\hat{x}_k^1,...,\hat{x}_k^{N_T},\ kT \in I\}$, we have $P[\{\hat{d}_i\} \mid \underline{y}(t)] = P[\{\underline{\hat{x}}_k\} \mid \underline{y}(t)]$, with

$$P[\{\underline{\hat{x}}_k\}|\underline{y}(t)] = K\, P[\underline{y}(t)|\{\underline{\hat{x}}_k\}] \prod_{kT\in I} \prod_{a=1}^{N_T} \prod_{l=1}^{N} P[\hat{d}^{\,a}_{Nk+l}] \quad (5)$$

$$P[\underline{y}(t)|\{\underline{\hat{x}}_k\}] \sim \exp\{\frac{-1}{4N_0}\sum_{b=1}^{N_R}\int_I |\, y^b(t) - \sum_{a=1}^{N_T}\sum_{kT\in I} \hat{x}_k^a\, h^{b,a}(t-kT;t)|^2\, dt\} \quad (6)$$

In (5) the bits are assumed to be independent, due to bit interleaving. Furthermore, $P[\hat{d}^{\,a}_{Nk+l}]$ is the a priori probability that bit $d^{\,a}_{Nk+l}$ equals $\hat{d}^{\,a}_{Nk+l}$, $N = \log_2 M$ is the number of bits per antenna symbol, and $P[\underline{y}(t)] = K$.

Let us define the state at time $(k\text{-}1)T$ for some finite N_S as

$$\hat{S}_{k-1} = (\hat{x}^1_{k-N_S+1},....,\hat{x}^{N_T}_{k-N_S+1},....,\hat{x}^1_{k-1},....,\hat{x}^{N_T}_{k-1})\,. \quad (7)$$

Following the Ungerboeck equalizer formulation [15], it is shown in [12] that (5) can be factored (apart from constant factors) as

$$P[\{\underline{\hat{x}}_k\}|\underline{y}(t)] \sim \prod_{kT\in I} \underbrace{\gamma_{ch}(\hat{S}_{k-1},\hat{S}_k)P(\hat{S}_k\,|\,\hat{S}_{k-1})}_{\gamma_k(\hat{S}_{k-1},\hat{S}_k)} \quad (8)$$

where the *a priori transition probability* is defined as

$$P(\hat{S}_k\,|\,\hat{S}_{k-1}) = \prod\nolimits_{a=1}^{N_T} \prod\nolimits_{l=1}^{N} P[\hat{d}^{\,a}_{Nk+l}] \quad (9)$$

while the *channel transition probability* is defined as

$$\gamma_{ch}(\hat{S}_{k-1},\hat{S}_k) = \exp\{\frac{1}{4N_0}\sum_{a=1}^{N_T}\mathrm{Re}\left\{\hat{x}_k^{a*}\sum_{b=1}^{N_R}\left[2z^{b,a}(kT) + \right.\right. \\ \left.\left. -\sum_{c=1}^{N_T}\hat{x}_k^c s^{b,a,c}(kT;kT) - 2\sum_{c=1}^{N_T}\sum_{m>0}\hat{x}^c_{k-m}s^{b,a,c}(kT;(k-m)T)\right]\right\}\} \quad (10)$$

Assuming knowledge of the equivalent time-varying channel impulse responses, the *z-parameters* and *s-parameters* in (10) are computed as follows

$$z^{b,a}(kT) = \int_I h^{b,a^*}(t-kT;t)\,y^b(t)\,dt \quad (11)$$

$$s^{b,a,c}(kT;nT) = \int_I h^{b,a^*}(t-kT;t)\,h^{b,c}(t-nT;t)\,dt\,. \quad (12)$$

Given the factorization in (8), the a posteriori probabilities of the coded bits can be computed by the application of the MAP/BCJR algorithm [2]. For clarity, we describe the fundamental steps in Section C.

A. Static Channel Impulse Response

Let us assume the channel to be quasi static, i.e., it does not experience any time variation over a block of coded information. Therefore, the z and s parameters are calculated as follows

$$z^{b,a}(kT) = \int_I h^{b,a^*}(t-kT)\,y^b(t)\,dt \quad (13)$$

$$s^{b,a,c}(kT;nT) = \int_{t+kT\in I} h^{b,a^*}(t)\,h^{b,c}(t+kT-nT)\,dt \quad (14)$$

In this scenario, the s-parameters do not need to be updated over the burst but can be computed off-line once.

B. Nyquist Pulse Shaping and Slowly Time-Varying Discrete Multi-Path Channel

Let us assume a typical tapped delay line channel model, i.e., $g^{b,a}_{ch}(\tau;t) = \sum_{p=1}^{N_P} g^{b,a}_{ch}(p;t)\delta(\tau-\tau_p)$. Under the hypothesis of

1. Nyquist pulse shaping and no excess bandwidth,
2. slow time variations compared to the duration of the transmit pulse,
3. symbol spaced taps, i.e., $\tau_p = pT$,

it can be shown that the metric (10) becomes

$$\gamma_{ch}(\hat{S}_{k-1},\hat{S}_k) = \exp\{\frac{-1}{2N_0}\sum_{b=1}^{N_R}|\,\hat{z}^b(kT) - \sum_{a=1}^{N_T}\sum_{p=1}^{N_P}\hat{x}^a_{k-p}g^{b,a}_{ch}(p;kT)|^2\} \quad (15)$$

where the sequence of samples is obtained by filtering with a filter matched to the transmit pulse, i.e.,

$$\hat{z}^b(kT) = \int_I g^*(t-kT)\,y^b(t)\,dt \quad (16)$$

This well-known metric is the one we used in [11].

C. Modified MAP/BCJR Algorithm

The a posteriori probabilities of coded bit d_i^m, transmitted during time period kT by antenna m, are obtained (ignoring a constant factor) as

$$\lambda(d_i^m = \pm 1) \sim \sum_{(S_{k-1},S_k)\in D(\pm 1)} \alpha_{k-1}(S_{k-1})\gamma_k(S_{k-1},S_k)\beta_k(S_k) \quad (17)$$

where the sum is computed over all state transitions corresponding to $d_i^m = \pm 1$ with $i=Nk+l$, for a given $l=1,...,N$. Further, $\alpha_k(S_k)$ and $\beta_{k-1}(S_{k-1})$ are recursively computed as

$$\alpha_k(S_k) = \sum_{S_{k-1}\in\Sigma}\alpha_{k-1}(S_{k-1})\gamma_k(S_{k-1},S_k) \quad (18)$$

$$\beta_{k-1}(S_{k-1}) = \sum_{S_k\in\Sigma}\beta_k(S_k)\gamma_k(S_{k-1},S_k) \quad (19)$$

with Σ being the set of all possible states. The cardinality of such a set is $|\Sigma| = M^{N_T(N_S-1)}$ with M modulation order, and N_S such that $s^{b,a,c}(kT;kT-mT) = 0$ for $m \geq N_S$. N_S is finite assuming time limited pulse shaping filters and channel impulse responses.

We can now summarize the fundamental operations for computing the a posteriori probabilities as follows.

1. Compute the z-parameters. These are obtained by convolving each received signal with a bank of time varying filters matched to each transmit antenna link. The matched filter outputs are then sampled at rate T.
2. Compute the s-parameters. These parameters are obtained by cross-correlating the impulse responses of the transmit-receive antenna links.
3. Consider a block of matched filter outputs of finite length L. Initialize $\alpha_0(S_0)$ and $\beta_L(S_L)$, setting them to 1 in correspondence to the known starting and ending states, and to zero otherwise.
4. Recursively compute $\alpha_k(S_k)$ and $\beta_k(S_k)$.
5. Once $\alpha_{k-1}(S_{k-1})$ and $\beta_k(S_k)$ have been computed, they can be multiplied by the appropriate $\gamma_k(S_{k-1}, S_k)$ to obtain $\lambda(d_i^m)$ according to (17). The evaluation of the transition probability $\gamma_k(S_{k-1}, S_k)$ requires the a priori probability of the coded bits that are associated with such a transition, i.e., $P(\hat{S}_k \mid \hat{S}_{k-1})$.

Note that each received signal is used in a maximal ratio combining fashion into the channel transition probability. Further the a priori probability (9) is in practice approximated with the interleaved extrinsic information provided by the outer decoder in the previous decoding iteration, i.e., $P[\hat{d}^a_{Nk+l}] = \lambda_a(\hat{d}^a_{Nk+l})$. Equally likely bits are assumed at the first equalization stage.

Finally, the extrinsic a posteriori probabilities at the equalizer output are computed as

$$\lambda_e(d_i^m = \pm 1) = \lambda(d_i^m = \pm 1) / \lambda_a(d_i^m = \pm 1) \qquad (20)$$

This corresponds to evaluate (17) by taking into account, in the transition probability $\gamma_k(\hat{S}_{k-1}, \hat{S}_k)$, the extrinsic information of all transmitted bits except the one under consideration.

D. Symbol Interleaving

Instead of computing the a posteriori probabilities of the coded bits, we can compute the a posteriori probabilities of the coded symbols $P[x_k^m \mid \underline{y}(t)]$ by simply computing the sum in (17) over the set of state transitions corresponding to the desired symbol. For instance, this is the goal in a space-time coded system that deploys block/trellis codes with symbol interleavers. In this case, the outer decoder has to provide the extrinsic state transition probabilities $P(\underline{S}_k \mid \underline{S}_{k-1})$, since the assumption of independence is valid at symbol level, and the bit-wise factorization in (9) does not hold.

V. Approaches to Reduce Complexity

The complexity of the algorithm is determined by the number of states, equal to $|\Sigma| = M^{N_T(N_S-1)}$, and the number of transition to/from each state, equal to M^{N_T}.

Complexity reduction in the equalization stage can be achieved by operating in the logarithm domain, and by introducing the MAX-LOG-MAP implementation of the BCJR algorithm [17]. Further, we should seek strategies to reduce the number of states and transitions.

States reduction can be obtained by applying the M-algorithm or the set partitioning with decision feedback technique [7]. Basically, as we proceed along the trellis we constrain the number of states by forcing hard decisions on past symbols. This can be done dynamically in the temporal and/or spatial direction.

Turbo processing allows for further simplifications since we can include, in the successive equalization stages, soft/hard feedback from the decoder and cut some of the transitions in the temporal and/or spatial direction.

Finally, it should be noted that the BCJR algorithm requires a forward and backward recursion on a full-state trellis. Nevertheless, it is possible to reduce the number of states by retaining for instance the reduced trellis of the forward direction also for the backward recursion.

VI. Example of Application to the EDGE System

As an example we consider the application of space-time bit-interleaved codes to the EDGE (enhanced data rates for GSM evolution) air interface. According to [18] transmission rates of 1, 2 and 3 bits/s/Hz are achieved by deploying bit-interleaved tail terminated convolutional codes, with 8-PSK modulation. A Gaussian filter (with normalized bandwidth equal to 0.3) is used for partial response pulse shaping. Blocks of coded bits are interleaved across four bursts. The bursts are frequency hopped. For low mobility applications, we can assume that such bursts experience independent Rayleigh fading, with no channel variation over a single burst.

We seek performance enhancements through the deployment of space-time bit-interleaved codes with two transmit antennas. This would allow minimal standard changes. We follow the code construction criteria in [13], such that the bits at the output of the convolutional encoder are parsed into two streams. Each antenna bit stream is randomly interleaved and transmitted over four bursts. Code rates up to 1/4 are obtained by puncturing (from right to left) of a mother code with polynomials in octal notation (133, 171, 145, 145) for both the single and double transmit antenna system. Code rate, modulation order, and information bit block size are indicated in Fig. 3. The mapping from bits to symbols is in accordance to the Gray rule.

Now, in Fig. 3 we report block error rate simulation results versus average signal energy to noise ratio (normalized over the number of transmit antennas). Perfect knowledge of the channel state information is assumed. The TU (typical urban) channel model is assumed with independent Rayleigh faded rays [18], and uncorrelated transmit-receive links. With such a channel model the ISI spans about three symbols requiring an equalizer with up to $|\Sigma| = 4096$ states with 8-PSK and two transmit antennas. Complexity reduction is obtained by using the MAX-LOG-MAP approximation in both the equalizer and the decoder. Further, the number of states in the equalizer is reduced through decision feedback. In particular, for the single transmit antenna case the reduced number of states is $|\Sigma_{red}|=8$. For the double transmit antenna case, $|\Sigma_{red}|=16$ with 4-PSK modulation, and $|\Sigma_{red}|=64$ with 8-PSK.

Curves with up to four iterations are shown for both the single and double transmit antenna system. At BLER=10^{-2} the double transmit antenna system gains 2, 4, 13 dB over the single transmit antenna system respectively for data rates of 1, 2, and 3 bits/s/Hz. An error floor appears at the first decoding iteration in the 2 and 3 bit/s/Hz curves. This is due to the deployment of decision feedback, however, it is removed through iterative processing, as Fig. 3 shows.

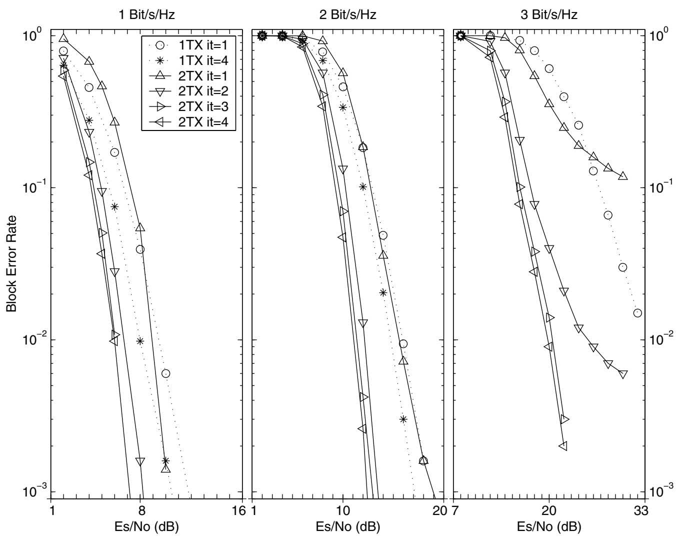

Fig.3. Block error rate performance of space-time bit-interleaved coded modulation with 2 transmit 1 receive antenna with turbo MAP equalization (solid lines). Performance with 1 transmit antenna is also shown (dashed lines). TU channel model. Transmission over four bursts with ideal frequency hopping and static fading on each burst. Uncorrelated fading across distinct antenna links. Block length, code rate, and modulation:

- **1 Bit/s/Hz** 400 info bits: **1 TX** R=1/3 with 8-PSK, **2 TX** R=1/4 with 4-PSK
- **2 Bit/s/Hz** 800 info bits: **1 TX** R=2/3 with 8-PSK, **2 TX** R=1/2 with 4-PSK
- **3 Bit/s/Hz** 1200 info bits: **1 TX** R=1/1 with 8-PSK, **2 TX** R=1/2 with 8-PSK

VII. CONCLUSIONS

We have considered the deployment of bit-interleaved space-time codes over frequency selective fading channels. Decoding is based on turbo MIMO equalization and decoding which is particularly suited for space-time coded systems that deploy symbol or bit interleavers. Following the Ungerboeck's approach, a maximum a posteriori equalizer for both static and time varying MIMO ISI channels has been described. The resulting structure comprises a bank of matched filters followed by an a posteriori probabilities calculator based on the BCJR algorithm with a modified metric. This space-time equalizer is optimum in the probabilistic sense, and does not necessarily require multiple receive antennas. Methods for reducing its complexity have also been considered.

We have evaluated the performance of space-time bit-interleaved convolutional codes over the EDGE air interface. This space-time coding approach allows for the exploitation of the temporal, spatial, and frequency diversity resources. As shown also in [14], it is an effective approach for systems that combine, for instance, multiple transmit antennas and slow frequency hopping as in the GSM/EDGE system.

ACKNOWLEDGMENTS

The author would like to thank Dr. J. P. Seymour, and Dr. N. Himayat of Lucent Technologies – Bell Labs for helpful discussions on the subject of this paper.

REFERENCES

[1] S. L. Ariyavisitakul, "Turbo space-time processing to improve wireless channel capacity", *IEEE Trans. on Comm.*, pp. 1347-1359, August 2000.

[2] L. R. Bahl, J. Cocke, F. Jelinek, J. Raviv, "Optimal decoding of linear codes for minimizing symbol error rate", *IEEE Tr. Info. Theory*, pp. 284-287, March 1974.

[3] G. Bauch, N. Al-Dhair, "Iterative equalization and decoding with channel shortening filters for space-time coded modulation", *Proc. of IEEE VTC-2000 Fall*, Boston, pp. 1575-1582, September 2000.

[4] S. Benedetto, D. Divsalar, G. Montorsi, F. Pollara, "Serial concatenation of interleaved codes: performance analysis, design, and iterative decoding", *IEEE Tr. on Info. Theory*, pp. 909-926, May 1998.

[5] G. E. Bottomley, S. Chennakeshu, "Unification of MLSE receivers and extension to time-varyimg-channels", *IEEE Trans. on Comm.*, pp. 464-472, April 1998.

[6] C. Douillard, A. Picart, *et al.* "Iterative correction of intersymbol interference: Turbo-equalization", *European Trans. on Telecomm.*, pp. 507-511, September/October 1995.

[7] M. V. Eyuboglu, S. U. H Qureshi, "Reduced-state sequence estimation with set partitioning and decision feedback", *IEEE Trans. on Comm.,* pp.13-20, January 1988.

[8] G. D. Forney, "Maximum-likelihood sequence estimation of digital sequences in the presence of inter symbol interference", *IEEE Trans. Inform. Theory*, pp. 363-378, May 1972.

[9] V. Tarokh, N. Seshadri, A. R. Calderbank, "Space-time codes for high data rate wireless communication: performance criterion and code construction", *IEEE Trans. Info. Theory*, pp. 744-765, March 1998.

[10] A. M. Tonello, "Space-time bit-interleaved coded modulation with an iterative decoding strategy", *Proc. of IEEE VTC-2000-Fall*, Boston, pp. 473-478, September 2000.

[11] A. M. Tonello, "Space-time bit-interleaved coded modulation over frequency selective fading channels with iterative decoding", *Proc. of IEEE Globecom-2000*, San Francisco, pp. 1616-1620, November 2000.

[12] A. M. Tonello, "MIMO MAP equalization for turbo decoding of interleaved space-time codes", *submitted to IEEE Trans. on Comm.*, January 2001.

[13] A. M. Tonello, "Performance of space-time bit-interleaved codes in fading channels with simplified iterative decoding", *Proc. of IEEE VTC-2001-Spring*, Rhodes, May 2001.

[14] A. M. Tonello, "Array processing for simplified turbo decoding of interleaved space-time codes", *Proc. of IEEE VTC-2001-Fall*, Atlantic City, October 2001.

[15] G. Ungerboeck, "Adaptive maximum likelihood receiver for carrier-modulated data transmission systems", *IEEE Trans. on Comm.*, pp. 624-636, May 1974.

[16] S. Verdu', "Minimum probability of error for asynchronous Gaussian multiple-access channels", *IEEE Trans. on Info. Theory*, pp. 85-96, January 1986.

[17] A.J. Viterbi, "An intuitive justification and a simplified implementation of the MAP decoder for convolutional codes", *IEEE Journal Select. Areas Comm.*, pp. 260-264, February 1998.

[18] "GSM Technical specifications 5.xx", ETSI 1999.