A Comparison of Convolutional and Turbo Coding Schemes For Broadband FWA Systems

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Abstract— **The block fading characteristics of fixed wireless access (FWA) channels do not allow the exploitation of time diversity through the use of powerful codes combined with interleaving. We demonstrate that turbo codes are not necessarily the optimal solution in block fading channels. Convolutional codes, carefully selected so as to present the same decoding complexity as turbo codes, achieve similar performance when used in systems without antenna diversity. When antenna diversity is exploited, turbo codes outperform convolutional codes only for a large number of antennas.**

Index Terms—**FWA, turbo codes, convolutional codes, decoding complexity, OFDM.**

I. Introduction

Broadband fixed wireless access (FWA) systems can provide high data rate communications where traditional landlines are either unavailable or too costly to be installed [1]. Such systems may operate over links where the line-of-sight (LOS) component is very small or even nonexistent, serving residential or small office/home office (SOHO) subscribers.

Recently, the IEEE 802.16a standard has included orthogonal frequency-division multiplexing (OFDM) as one of the available transmission techniques to combat multipath delay spread, which is experienced in broadband FWA systems [2]. Moreover, turbo codes are also included as one of a number of possible channel codes for these systems. Turbo codes have proved to be powerful in both the additive white Gaussian noise (AWGN) channel [3] as well as the perfectly interleaved Rayleigh fading channel [4]. However the broadband FWA channel suffers from slow fading and consequently a large number of consecutive symbols in the same block experience identical fading. The block fading experienced in such a channel prevents the exploitation of time diversity in order to improve performance through the use of powerful coding combined with interleaving owing to delay considerations. Hoshyar *et al.* [5] showed that turbo and convolutional codes, combined with OFDM, have nearly the same bit error rate (BER) performance in block fading channels, when antenna diversity is not exploited. Lin *et al.* [6] showed that, when antenna diversity is exploited, turbo codes outperform convolutional codes only at high signal-to-noise ratios in Rayleigh slow-fading channels.

The motivation for this paper is to compare turbo-coded with convolutionally coded OFDM systems in realistic broadband FWA scenarios. Furthermore, in contrast to [5] and [6] complexity considerations are taken into account, i.e., the decoders are carefully configured so that the comparisons are performed with equal decoder complexity. In particular, we consider broadband FWA systems both with and without antenna diversity. Although turbo codes have already been proposed for multiple antenna systems [7], a thorough performance comparison between turbo coding and convolutional coding in such cases has not been performed.

II. System Model

In this paper, we compare the performance of turbo coding to that of convolutional coding in broadband FWA systems, both with and without antenna diversity. Six interim BFWA channel models have been adopted by IEEE 802.16a [8]. In this paper we consider the SUI3 model, which corresponds to average suburban conditions. The SUI3 model has three fading taps at delays of 0, 0.5 and 1 μs, with relative powers of 0 dB, -5 dB and -10 dB. The Doppler spread is 0.4 Hz and the RMS delay spread is 0.264 μs, when omni-directional antennas are used. When multiple transmit and multiple receive antennas are employed, the envelope correlation coefficient is 0.4.

A. Single-input, single-output (SISO) coded OFDM

The SISO coded OFDM model, which employs one transmit and one receive antennas, is shown in Fig. 1.

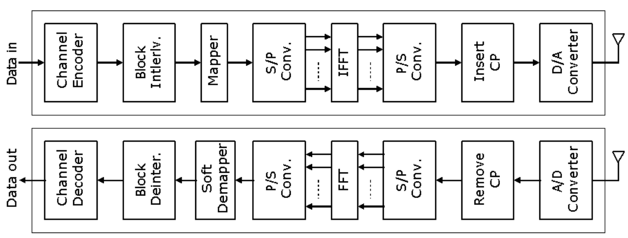

Fig. 1. SISO model of a coded OFDM system

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This work is sponsored by EPSRC under Grant GR/S46437/01.

The source bits are input to a channel encoder. In this paper we consider both convolutional coding and turbo coding. For the case of convolutional coding, the encoder uses a recursive systematic (RSC) code with code rate 1/2. In the case of turbo coding, the encoder is a parallel concatenation of two recursive systematic convolutional encoders, as described in [3]. The source bits feed the first constituent encoder, while the second encoder is fed by an interleaved version of the original data. The interleaver is assumed to be random and has a size of L bits. The output of the turbo encoder consists of the systematic bits of the first RSC encoder, the parity bits of the first RSC encoder and the parity bits of the second RSC encoder. In order to increase the code rate, puncturing of the parity bits can be applied. Puncturing patterns specify which bits are allowed to pass and which are rejected. For example, if patterns "10" and "01" are applied to the first and second encoders respectively, all bits located in even positions within the parity stream of the first RSC encoder and all bits located in odd positions within the parity stream of the second RSC encoder will be rejected, achieving an overall code rate of 1/2.

The coded bits are shuffled by a block interleaver, which is inserted in order to reduce the effects of the block-fading channel. After channel encoding and block interleaving, the binary signal is mapped onto modulation symbols.

The stream of modulation symbols is converted to N parallel streams. At each signaling interval, a block of N parallel symbols $\mathbf{S}=\{S_1,S_2,\ldots,S_N\}$ is modulated using an inverse fast Fourier transform (IFFT). The IFFT operation is denoted by an NxN matrix $\mathbf{Q}^{-1}$. The output of the IFFT at each signaling interval is a block of N channel symbols $\mathbf{s}=\{s_1,s_2,\ldots,s_N\}$, where $\mathbf{s}$ is given by:

$$\mathbf{s} = \mathbf{Q}^{-1}\mathbf{S} \tag{1}$$

Before transmission of $\mathbf{s}$ over the channel, a cyclic prefix is appended to prevent inter-symbol interference (ISI) and inter-channel interference (ICI).

At the receiver, the cyclic prefix of each OFDM block is removed, at each signaling interval. The relationship between the block of receive symbols $\mathbf{r}=\{r_1, r_2,\ldots, r_N\}$, after the cyclic prefix removal, and the block of transmit symbols $\mathbf{s}$, before the cyclic prefix insertion, can be expressed as:

$$\mathbf{r} = \mathbf{H}\mathbf{s} + \mathbf{n} \tag{2}$$

where $\mathbf{n}$ is a length N, white Gaussian noise signal block and $\mathbf{H}$ is a NxN circulant matrix that contains the channel impulse response. It is assumed that the receiver has perfect knowledge of the channel. Circulant matrices can be diagonalized by the Fourier transformation matrix. Therefore, if $\mathbf{Q}$ and $\mathbf{Q}^{-1}$ correspond to the FFT and IFFT operations, matrix $\mathbf{H}$ can be rewritten as:

$$\mathbf{H} = \mathbf{Q}^{-1}\mathbf{\Lambda}\mathbf{Q} \tag{3}$$

where $\mathbf{\Lambda}$ is a NxN diagonal matrix whose elements $\lambda(m)$ m=1,…,N, are the eigenvalues of $\mathbf{H}$ and correspond to the discrete Fourier transform (DFT) of the channel impulse response. The elements $\lambda(m)$ are also the channel gains experienced by the OFDM sub-carriers.

The FFT operation following the serial-to-parallel conversion, provides the received modulation symbols $\mathbf{R}=\{R_1,R_2,\ldots,R_N\}$ from the received block of symbols $\mathbf{r}$, since $\mathbf{R}=\mathbf{Q}\mathbf{r}$. Substituting for $\mathbf{r}$ gives:

$$\mathbf{R} = \mathbf{Q}\mathbf{r} = \mathbf{Q}(\mathbf{H}\mathbf{s}+\mathbf{n}) = \mathbf{Q}\left[\left(\mathbf{Q}^{-1}\mathbf{\Lambda}\mathbf{Q}\right)\mathbf{s}+\mathbf{n}\right] = \mathbf{\Lambda}\mathbf{S}+\mathbf{N} \tag{4}$$

since $\mathbf{S}=\mathbf{Q}\mathbf{s}$ and $\mathbf{N}=\mathbf{Q}\mathbf{n}$.

At each signaling interval, the soft demapper uses the received block of symbols $\mathbf{R}$ to calculate the a-posteriori log-likelihood ratios (LLRs) of the received bits, which are required by the channel decoder. We use the notation $b(\ell,m)$ to express the ℓ-th bit conveyed by the m-th OFDM sub-carrier, where ℓ depends on the modulation scheme used and m=1,…,N. The a-posteriori LLR L_D of $b(\ell,m)$, given vector $\mathbf{R}$ was received, is equal to:

$$L_D\left(b(\ell,m)\middle|\mathbf{R}\right) = ln\frac{Prob\left\{b(\ell,m)=1\middle|\mathbf{R}\right\}}{Prob\left\{b(\ell,m)=0\middle|\mathbf{R}\right\}} \tag{5}$$

L_D can be expressed as the sum of two terms, namely an a-priori LLR L_A and an extrinsic LLR L_E:

$$L_D\left(b(\ell,m)\middle|\mathbf{R}\right) = L_A\left(b(\ell,m)\right) + L_E\left(b(\ell,m)\middle|\mathbf{R}\right) \tag{6}$$

Since no a-priori knowledge is assumed in our model, L_A is equal to zero and so $L_D=L_E$. Taking into account the modulation scheme used, L_D can be further simplified. Due to space limitations, only the LLR expressions for QPSK modulation are given but similar expressions for higher modulation levels can also be easily derived. In QPSK two bits are grouped to form a symbol. The a-posteriori LLRs of $b(\ell=1,m)$ and $b(\ell=2,m)$, given vector $\mathbf{R}$ was received, is given by:

$$\begin{aligned} L_D\left(b(\ell=1,m)\middle|\,\mathbf{R}\right) &= 4\cdot SNR\cdot\left|\lambda(m)\right|\cdot Re\left\{e^{-j\angle\lambda(m)}R_m\right\} \\ L_D\left(b(\ell=2,m)\middle|\,\mathbf{R}\right) &= 4\cdot SNR\cdot\left|\lambda(m)\right|\cdot Im\left\{e^{-j\angle\lambda(m)}R_m\right\} \end{aligned} \tag{7}$$

After the LLR values are de-interleaved, they are used by the channel decoder to produce estimates of the transmitted bits. The convolutionally coded OFDM system uses the Viterbi algorithm (VA) to decode the received bits. The decoder of the turbo-coded OFDM system consists of two soft-input soft-output decoders. The first decoder uses a-priori information to produce soft estimates of the transmitted bits by processing the LLRs of the received systematic bits and the LLRs of the received parity bits of the first RSC encoder. Extrinsic information is extracted from the soft estimates and acts as a-priori information for the second decoder. Similarly, the second decoder processes the LLRs of the received interleaved systematic bits as well as the LLRs of the received parity bits of the second RSC encoder, to produce better estimates of the transmitted bits as well as extrinsic information, which will be used as a-priori information by the first decoder at the next iteration. The trellis-based decoding algorithms considered in this paper are the optimal Maximum

A-Posteriori algorithm in the log domain (log-MAP), also known as BCJR algorithm [9], the soft output Viterbi algorithm (SOVA) [10] and the max-log-MAP algorithm [11].

B. Multiple-input, multiple-output (MIMO) coded OFDM

The MIMO coded OFDM model, which employs N_T transmit antennas and N_R receive antennas, is shown in Fig. 2.

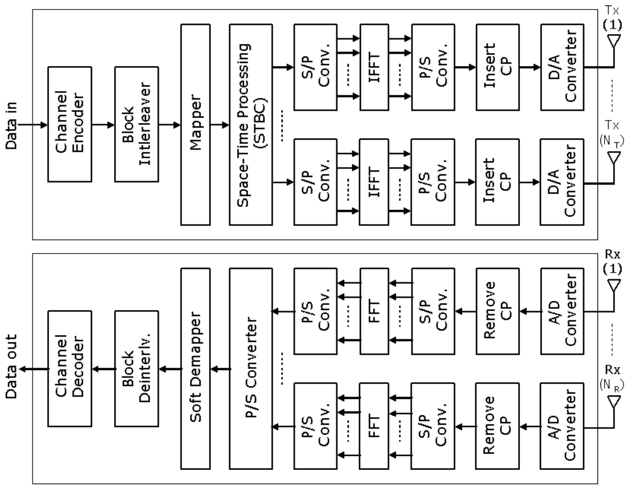

Fig. 2. MIMO model of a coded OFDM system

After channel encoding and block interleaving, the binary signal is mapped onto modulation symbols. For an arbitrary number of transmit antennas, Tarokh *et al.* [12] generalized the space-time coding scheme introduced by Alamouti [13]. For simplicity, we have implemented Alamouti-type space-time block coding (STBC) with N_T=2 transmit and N_R receive antennas. Assume that $\mathbf{S}$ and $\mathbf{S}'$ are two consecutive blocks of modulation symbols each of length N, conveyed during two consecutive signaling intervals, which are input to the STBC encoder. Each of the j outputs of the STBC encoder, where j=1 or 2 in the case of Amalouti-type STBC, is linked with a transmit antenna. During the first signaling interval, $\mathbf{S}_{j=1}=\mathbf{S}$ and $\mathbf{S}_{j=2}=\mathbf{S}'$ will be the blocks of symbols at streams j=1 and j=2 respectively, since N_T=2. During the second time interval $\mathbf{S}'_{j=1}=-(\mathbf{S}')^*$ and $\mathbf{S}'_{j=2}=(\mathbf{S})^*$ will be the blocks of symbols at streams j=1 and j=2 respectively, where $\mathbf{S}^*$ represents the conjugate of $\mathbf{S}$.

OFDM modulation is performed on the blocks of symbols $\mathbf{S}_{j=1}$ and $\mathbf{S}_{j=2}$ during the first signaling interval and $\mathbf{S}'_{j=1}$ and $\mathbf{S}'_{j=2}$ during the second signaling interval. Based on (1), blocks $\mathbf{s}_{j=1}=\mathbf{s}$ and $\mathbf{s}_{j=2}=\mathbf{s}'$ are generated and transmitted simultaneously during the first signaling interval, whereas $\mathbf{s}'_{j=1}=-(\mathbf{s}')^*$ and $\mathbf{s}'_{j=2}=(\mathbf{s})^*$ are generated and transmitted during the second signaling interval. Before transmission over the channel, a cyclic prefix is appended to each block to prevent ISI and ICI.

At the receiver, we assume that N_R=1 antenna is deployed in order to simplify the analysis. Expressions can be easily extended to cover the case for N_R>1. The cyclic prefix of each OFDM block is removed, at each signaling interval. Assume that $\mathbf{r}$ and $\mathbf{r}'$ are two consecutive blocks of receive symbols of length N each, received during two consecutive signaling intervals. The relationship between the blocks of receive symbols $\mathbf{r}$ and $\mathbf{r}'$ and the blocks of transmit symbols $\mathbf{s}$ and $\mathbf{s}'$ can be expressed as:

$$\begin{aligned} \mathbf{r} &= \mathbf{H}_1\mathbf{s} + \mathbf{H}_2\mathbf{s}' + \mathbf{n} \\ \mathbf{r}' &= -\mathbf{H}_1(\mathbf{s}')^* + \mathbf{H}_2(\mathbf{s})^* + \mathbf{n}' \end{aligned} \tag{8}$$

which is equivalent to:

$$\begin{bmatrix} \mathbf{r} \\ (\mathbf{r}')^* \end{bmatrix} = \begin{bmatrix} \mathbf{H}_1 & \mathbf{H}_2 \\ \mathbf{H}_2^* & -(\mathbf{H}_1)^* \end{bmatrix} \cdot \begin{bmatrix} \mathbf{s} \\ \mathbf{s}' \end{bmatrix} + \begin{bmatrix} \mathbf{n} \\ \mathbf{n}' \end{bmatrix} \tag{9}$$

where $\mathbf{H}_j$ is an $N\times N$ circulant matrix that contains the channel impulse response from transmit antenna j to the receive antenna, and $\mathbf{n}$ and $\mathbf{n}'$ are noise vectors during two consecutive signaling intervals. Based on (3) and (4), we can derive the relation between the receive modulation symbols $\mathbf{R}$ and $\mathbf{R}'$ and the transmit modulation symbols $\mathbf{S}$ and $\mathbf{S}'$:

$$\begin{bmatrix} \mathbf{R} \\ (\mathbf{R}')^* \end{bmatrix} = \begin{bmatrix} \mathbf{\Lambda}_1 & \mathbf{\Lambda}_2 \\ \mathbf{\Lambda}_2^* & -(\mathbf{\Lambda}_1)^* \end{bmatrix} \cdot \begin{bmatrix} \mathbf{S} \\ \mathbf{S}' \end{bmatrix} + \begin{bmatrix} \mathbf{N} \\ \mathbf{N}' \end{bmatrix} \tag{10}$$

where $\mathbf{\Lambda}_j$ is an $N\times N$ diagonal matrix whose elements $\lambda_j(m)$, $m=1,\ldots,N$, correspond to the DFT of the channel impulse response from transmit antenna j to the receive antenna and $\mathbf{N}$ and $\mathbf{N}'$ correspond to the DFT of the noise vectors $\mathbf{n}$ and $\mathbf{n}'$.

As in the SISO model, the soft demapper calculates the a-posteriori LLRs of the received bits. Since $\mathbf{R}$ and $\mathbf{R}'$ are blocks of modulation symbols received in two consecutive time intervals, we use the notation $b(\ell,m)$ to express the ℓ-th bit conveyed by the m-th OFDM sub-carrier during the first time interval and $b'(\ell,m)$ to express the ℓ-th bit conveyed by the m-th OFDM sub-carrier during the second time interval. Based on (5) and (6), the a-posteriori LLR expressions of $b(\ell,m)$ and $b'(\ell,m)$, given vectors $\mathbf{R}$ and $\mathbf{R}'$, for QPSK modulation are given by:

$$\begin{aligned} L_D(b(1,m)|\mathbf{R},\mathbf{R}') &= 4\cdot\mathrm{SNR}\cdot\mathrm{Re}\{(\lambda_1(m))^*\cdot\mathbf{R}_m + \lambda_2(m)\cdot(\mathbf{R}'_m)^*\} \\ L_D(b(2,m)|\mathbf{R},\mathbf{R}') &= 4\cdot\mathrm{SNR}\cdot\mathrm{Im}\{(\lambda_1(m))^*\cdot\mathbf{R}_m + \lambda_2(m)\cdot(\mathbf{R}'_m)^*\} \end{aligned} \tag{11}$$

$$\begin{aligned} L_D(b'(1,m)|\mathbf{R},\mathbf{R}') &= 4\cdot\mathrm{SNR}\cdot\mathrm{Re}\{(\lambda_2(m))^*\cdot\mathbf{R}_m - \lambda_1(m)\cdot(\mathbf{R}'_m)^*\} \\ L_D(b'(2,m)|\mathbf{R},\mathbf{R}') &= 4\cdot\mathrm{SNR}\cdot\mathrm{Im}\{(\lambda_2(m))^*\cdot\mathbf{R}_m - \lambda_1(m)\cdot(\mathbf{R}'_m)^*\} \end{aligned} \tag{12}$$

Finally, the LLR values are de-interleaved and used by the channel decoder to produce estimates of the transmitted bits.

III. Complexity of the Decoding Algorithms

The decoding algorithms considered for turbo decoding are log-MAP, max-log-MAP and SOVA. The Viterbi algorithm is used for convolutional decoding. A complexity analysis was presented in [14] but in order to simplify the comparison it was assumed that logical and mathematical operations have similar complexity. A more thorough investigation was performed in [15], where each operation is quantified as a number of equivalent additions. In our analysis, the complexity expressions were re-derived since the complexity estimations for SOVA and for log-MAP in [15] were rather pessimistic. More specifically, it is assumed in [15] that SOVA operates in the trace-back mode [16] and that the table look-up in the log-MAP algorithm is equivalent to 6 additions. In contrast, in our analysis SOVA operates in the less complex register exchange mode, as described in [10], at the expense of additional storage requirements. A look-up operation in log-MAP is considered to be equivalent to 3 comparisons, resulting in a complexity of 3 equivalent additions, since only 8 values need to be stored in a look-up table [14]. In Table I the number of equivalent additions for the various operations is shown. Table II-IV list the complexity of the decoding algorithms, expressed in terms of the number of equivalent additions, for a code rate of 1/2, as a function of the encoder memory order M. The additional complexity of the branch metric calculations due to a-priori information exploited by the turbo decoder has also been taken into account.

TABLE I: NUMBER OF EQUIVALENT ADDITIONS PER OPERATION

Operations	Number of Equivalent Additions
Addition, Subtraction	1
Multiplication, Division	1
Comparison	1
Maximum, Minimum	2
Look-up Table	3

TABLE II: COMPLEXITY OF VITERBI ALGORITHM

Process	Number of Equivalent Additions
Branch Metric Calculations	$6 \cdot 2^M$
Path Metric Calculations	$4 \cdot 2^M$
Hard Decision	3
Overall Complexity	$\mathbf{10 \cdot 2^M + 3}$

TABLE III: COMPLEXITY OF SOVA

Process	Number of Equivalent Additions
Branch Metric Calculations	$12 \cdot 2^M$
Path Metric Calculations	$5 \cdot 2^M$
Hard Decision	3
Trace-back Procedure	$4 \cdot (5M) + 1$
Overall Complexity	$\mathbf{17 \cdot 2^M + 4 \cdot (5M) + 4}$

TABLE IV: COMPLEXITY OF LOG-MAP ALGORITHM

Process	Number of Equivalent Additions
Branch Metric Calculations	$12 \cdot 2^M$
Path Metrics (Forward proc.)	$9 \cdot 2^M$
Path Metrics (Backward proc.)	$9 \cdot 2^M$
Soft Decision	$18 \cdot 2^M - 13$
Overall Complexity	$\mathbf{48 \cdot 2^M - 13}$

TABLE V: COMPLEXITY OF MAX-LOG-MAP ALGORITHM

Process	Number of Equivalent Additions
Branch Metric Calculations	$12 \cdot 2^M$
Path Metrics (Forward proc.)	$4 \cdot 2^M$
Path Metrics (Backward proc.)	$4 \cdot 2^M$
Soft Decision	$8 \cdot 2^M - 3$
Overall Complexity	$\mathbf{28 \cdot 2^M - 3}$

The complexity of SOVA corresponds to the worst-case scenario, according to which the truncation depth of the trace-back procedure is always $5M$ and all the decoded bits across each diverging path differ from the corresponding bits of the survivor path. In practice, the actual complexity of SOVA is lower than the predicted complexity for small memory orders. As the memory order increases, the exponential term in the complexity expression dominates and the expression converges to the actual complexity, as shown in Fig. 3. For this reason, SOVA is not considered when the memory order of the turbo decoder is low and a comparison with the MAP-based algorithms would not be fair.

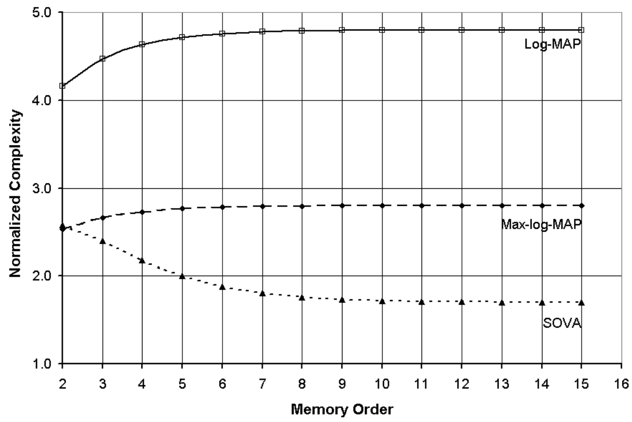

Fig. 3. Complexity of the iterative decoding algorithms normalized by the complexity of the conventional Viterbi algorithm

Based on our calculations, which are presented graphically in Fig. 4, the complexity of a turbo decoder with memory order M=2, which applies log-MAP with 7 iterations, is comparable to that of a similar turbo decoder that applies max-log-MAP with 11 iterations or a convolutional decoder with a memory order of M=8 that applies the conventional Viterbi algorithm.

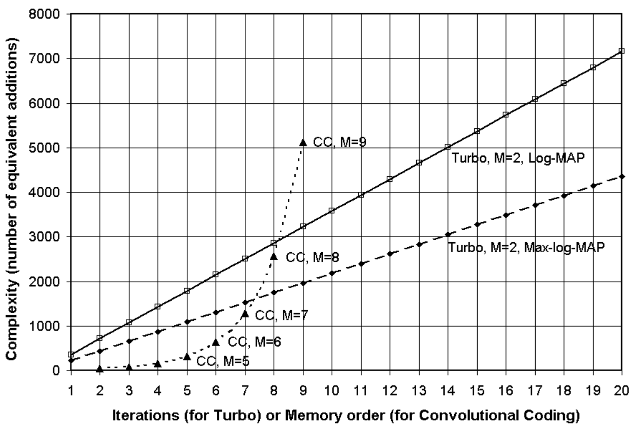

Fig. 4. Complexity comparison between turbo decoding and convolutional decoding

IV. SIMULATION RESULTS

In our simulations, the turbo encoder consists of a random interleaver with size L=1000 bits and two identical terminated RSC codes with rate 1/2, octal generator polynomial (1,5/7) and memory order M=2. The parity bits of the constituent RSC codes are punctured alternately so as to achieve an overall rate of 1/2. The convolutional encoder uses an

RSC(1,753/561) code with rate 1/2 and memory order M=8. An OFDM/16-QAM system with N=256 sub-carriers is used.

In Fig. 5 the performance of the turbo-coded OFDM system is compared to that of the convolutionally coded OFDM system. For comparison purposes the performance of these systems in the AWGN channel has been also included. We note that even though turbo codes outperform convolutional codes in the AWGN channel, the performance of both codes is equivalent in the SISO SUI3 channel (i.e., antenna diversity is not exploited). In the MIMO SUI3 channel (i.e., antenna diversity is exploited) turbo codes perform better than convolutional codes as the number of antennas increases. This is due to the fact that the underlying channel approaches a non-fading AWGN channel. However a very large number of antennas is required for turbo codes to achieve a noticeable coding gain over convolutional codes.
Furthermore, the performance of convolutional codes is better than turbo codes using the max-log-MAP algorithm with 11 iterations in SISO as well as low diversity MIMO systems (Fig. 6).

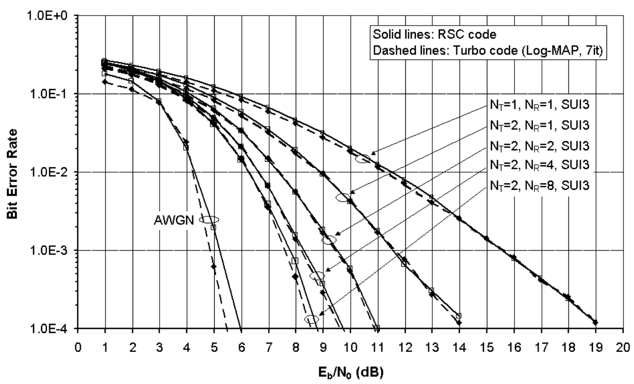

Fig. 5. BER comparison between turbo (applying the log-MAP algorithm) and convolutional codes

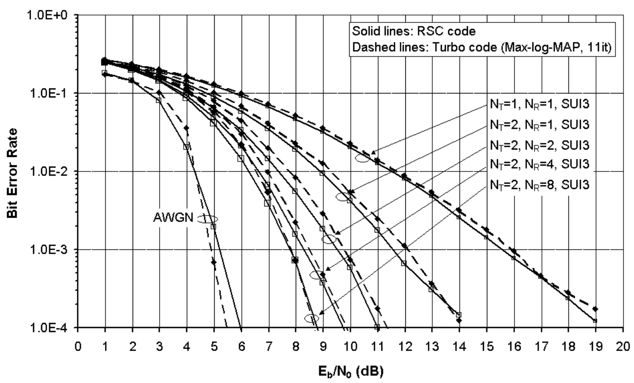

Fig. 6. BER comparison between turbo (applying the max log-MAP algorithm) and convolutional codes

V. Conclusions and Future Work

In this paper, we compare the performance of turbo codes to that of convolutional codes for broadband FWA scenarios. In particular, the various decoders are carefully configured so as to present similar complexity. It has been shown that convolutional codes perform similarly to turbo codes for BERs up to 10^{-4} in SISO systems, while turbo codes eventually outperform convolutional codes in MIMO systems using a large number of antennas. Consequently, turbo codes do not offer an advantage to broadband FWA systems, which use a limited number of antennas. These results are of practical interest for the deployment and design of high performance and broadband FWA systems. Future work will involve comparisons to convolutional codes specifically designed for block fading channels [17], which achieve maximum code diversity without necessarily exhibiting maximum free Hamming distance.

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