Novel Algorithms and Techniques in Telecommunications and Networking
Tarek Sobh · Khaled Elleithy · Ausif Mahmood
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Springer
15. Nonlinear Congestion Control Scheme for Time Delayed Differentiated-Services Networks .......... 93
   R. Vahidnia et al.

   Dina Darwish et al.

17. Improving BGP Convergence Time via MRAI Timer ................................................................. 105
   Abdelshakour Abuzneid and Brandon J. Stark

18. Error Reduction Using TCP with Selective Acknowledgement and HTTP with Page Response Time over Wireless Link ............................................................................................... 111
   Adelshakour Abuzneid, Kotadiya Krunalkumar

   Nicolae Crisan, Ligia Chira Cremene

    Abuhelaleh, Mohammed et al.

21. Intrusion Detection and Classification of Attacks in High-Level Network Protocols Using Recurrent Neural Networks ....................................................................................................... 129
    Vicente Alarcon-Aquino et al.

22. Automatic Construction and Optimization of Layered Network Attack Graph ................................ 135
    Yonggang Wang et al.

23. Parallel Data Transmission: A Proposed Multilayered Reference Model ...................................... 139
    Thomas Chowdhury, Rashed Mustafa

24. Besides Tracking – Simulation of RFID Marketing and Beyond .................................................. 143
    Zeeshan-ul-Hassan Usmani et al.

25. Light Path Provisioning Using Connection Holding Time and Flexible Window ................................ 149
    Fatima Yousaf et al.

    Dongkyun Kim et al.

27. Performance of the Duo-Binary Turbo Codes in WiMAX Systems .............................................. 161
    Teodor B. Iliev et al.

    Faisal Bashir Hussain, Yalcin Cebi

29. A Low Computational Complexity Multiple Description Image Coding Algorithm Based on JPEG Standard ................................................................................................................. 173
    Ying-ying Shan, Xuan Wang

    B. Y. Bedzhev and M. P. Iliev
Abstract – In this paper the broadband wireless access system, provided by the IEEE 802.16 wireless MAN air interface with its amendment to mobile users (IEEE 802.16e), is being analyzed. We provide performance results for the most important forward error correcting (FEC) schemes intended for IEEE 802.16e – convolutional turbo codes (CTC) and Low Density Parity Check (LDPC) codes.

Keywords – WiMax, Forward error correction, Turbo codes, Convolutional codes

I. INTRODUCTION

The IEEE 802.16 telecommunications standard [1] envisions broadband wireless access technology as a means of providing wireless “last mile” broadband access in a metropolitan area network (MAN). The performance and services should be comparable or better than those provided by traditional DSL, cable or T1/E1 leased lines. Especially in areas beyond the reach of DSL and cable, IEEE 802.16 could offer a cost-effective broadband access solution. The term WiMax (worldwide interoperability for microwave access) has become synonymous with IEEE 802.16, promoting and certifying compatibility and interoperability of broadband wireless products. In its original release 802.16 focused on line-of-sight (LOS) applications in the licensed 10 to 66 GHz frequency range based on single carrier (SC) transmission (WirelessMAN-SC). In the first amendment of the standard were covered non-line-of-sight (NLOS) applications in licensed and unlicensed bands in the 2 to 11 GHz frequency range (WirelessMAN-SCa). To meet the requirements of a low cost solution in a multipath environment, orthogonal frequency division multiplexing (OFDM) was chosen as physical layer transmission technique (WirelessMAN-OFDM). To deliver optimum broadband wireless access performance, the concept of scalable OFDMA was adopted. The architecture is based on a scalable subchannel bandwidth using a variable sized FFT according to the channel bandwidth. Within task group E (IEEE 802.16e, [1]) there is an ongoing evolution of IEEE 802.16 addressing mobile applications thus enabling broadband access directly to portable devices like smart phones, PDAs, notebooks and laptop computers.

Our investigations are focused on the uplink of the WirelessMAN-OFDMA physical layer of IEEE 802.16 together with its amendments for mobile applications addressed in IEEE 802.16e. For the purpose of forward error correction (FEC) within the IEEE 802.16 WirelessMAN-OFDMA standard, there is a mandatory convolutional code (CC) and an optional block turbo code (BTC) and convolutional turbo code (CTC). In the amendment for mobility (IEEE 802.16e) a Low-Density Parity-Check code (LDPC) was added.

The 8-state family has already been adopted in the digital video broadcasting (DVB) standards for return channel via satellite (DVB-RCS) [2] and the terrestrial distribution system (DVB-RCT) [3], and also in the 802.16a standard for local and metropolitan area networks [1]. Combined with the powerful technique of circular trellises, those duo-binary turbo codes offer good performance and versatility for encoding blocks with various sizes and rates, while keeping reasonable decoding complexity. The replacement of the 8-state component encoder by a 16-state encoder will provide better performance at low error rates, at the price of a doubled decoding complexity. The minimum Hamming distances are increased by 30%–50%, with regard to 8-state TCs, and allow frame-error rate (FER) curves to decrease below $10^{-7}$ without any noticeable change in the slope (the so-called flattening effect).

II. SYSTEM OVERVIEW

In IEEE 802.16e-2005, the channel coding stage consists of the following steps: (1) data randomization, (2) channel coding, (3) rate matching, (4) HARQ, if used, (5) and interleaving. Data randomization is performed in the uplink and the downlink, using the output of a maximum length shift-register sequence that is initialized at the beginning of every FEC block. This shift register sequence is modulo 2, added with the data sequence to create the randomized data. The purpose of the randomization stage is to provide layer 1 encryption and to prevent a rogue receiver from decoding the data. When HARQ is used, the initial speed of the shift-register sequence for each HARQ transmission is kept constant in order to enable joint decoding of the same FEC block over multiple transmissions.

Channel coding is performed on each FEC block, which consists of an integer number of subchannels. A subchannel is the basic unit of resource allocation in the PHY layer and comprises several data and pilot subcarriers. The exact number of data and pilot subcarriers in a subchannel depends on the subcarrier permutation scheme. The maximum number of subchannels in an FEC block depends on the channel coding scheme and the modulation constellation.
If the number of subchannels required for the FEC block is larger than this maximum limit, the block is first segmented into multiple FEC subblocks.

These subblocks are encoded and rate matched separately and then concatenated sequentially, as shown in Figure 1, to form a single coded data block. Code block segmentation is performed for larger FEC blocks in order to prevent excessive complexity and memory requirement of the decoding algorithm at the receiver. The channel is assumed to be a time-variant multipath channel for modeling mobile users in an NLOS scenario. The receiver noise is modeled by an additive white Gaussian noise (AWGN) process added to the received signal.

Assuming perfect synchronization, the receiver extracts the useful symbol time and therefore removes the cyclic prefix. After the computation of the frequency domain signal via the FFT, the receiver extracts the user specific information (Data Extraction) and feeds it to the channel estimator and the 1-tap equalizer. The channel estimator computes:
\[ \hat{H}_i = H_i X_i + N_i, \quad i = 0, \ldots, N_{awd} - 1 \]

where \( X_i \) is the transmitted symbol, \( H_i \) is the complex valued sample of the channel transfer function and \( N_i \) is the complex valued noise sample in subcarrier \( i \). The channel estimator computes \( \hat{H}_i \) which are estimates of the real channel factor \( H_i \). The topic of channel estimation for IEEE 802.16 is addressed in [4]. In the following we assume perfect knowledge of the channel transfer function and the 1-tap equalizer. With the assumption that the delay spread of the channel is smaller than the cyclic prefix and the time variance of the channel during one OFDM symbol is negligible, the received symbols in frequency domain \( R_i \) are given by:

\[ R_i = H_i X_i + N_i, \quad i = 0, \ldots, N_{awd} - 1 \]

These equalized symbols are fed into the soft output demodulator computing log-like ratios (LLRs) for bits. After deinterleaving the LLRs are fed into the FEC decoders using this soft input for decoding.

### III. Constituent RCS Codes

**A. Circular RCS Codes**

Among the different techniques aiming at transforming a convolutional code into a block code, the best way is to use any state of the encoder as the initial state, and to encode the sequence so that the final state of the encoder is equal to the initial state. The code trellis can then be viewed as a circle, without any state discontinuity. This termination technique, called tailbiting [5], [6] or circular, presents three advantages in comparison with the classical trellis-termination technique using tail bits to drive the encoder to the all-zero state. First, no extra bits have to be added and transmitted; thus, there is no rate loss, and the spectral efficiency of the transmission is not reduced. Next, when classical trellis termination is applied for TCs, a few codewords with input Hamming weight of one may appear at the end of the block (in both coding dimensions), and can be the cause of a marked decrease in the minimum Hamming distance of the composite code. With tailbiting RSC codes, only codewords with minimum input weight of two have to be considered. In other words, tailbiting encoding avoids any side effects

**B. Permutation**

Among the numerous permutation models that have been suggested up to now, the apparently most promising ones, in terms of minimum Hamming distances, are based on regular permutation calling for circular shifting [7] or the co-prime [8] principle. After writing the data in a linear memory, with address \( i (0 \leq i \leq N-1) \), the information block is likened to a circle, with both extremities of the block \( i = 0 \) and \( i = N-1 \) being contiguous. The data is read out, such that the \( j \)th data read was written at the position \( i \), given by:

\[ i = \Pi(j) = P j + i_0, \]

where the skip value \( P \) is an integer, relatively prime with \( N \), and \( i_0 \) is the starting index. This permutation does not require the block to be seen as rectangular, that is, \( N \) may be any integer.

In [9] and [10], two very similar modifications of (3) were proposed, which generalize the permutation principle adopted in the DVB-RCS or IEEE802.16a TCs. In the following, we will consider the almost regular permutation (ARP) model detailed in [10], which changes relation (3) into:

\[ i = \Pi(j) = P j + Q(j) + i_0 \mod N \]

where \( Q(j) \) is an integer, whose value is taken in a limited set \( \{0, Q_1, Q_2, \ldots, Q_{C-1}\} \), in a cyclic way. \( C \), called the cycle of the permutation, must be a divider of \( N \) and has a typical value of four or eight. For instance, if \( C=4 \), the permutation law is defined by:

\[ j = 0 \mod 4, \quad i = \Pi(j) = P j + 0 + i_0 \mod N \]

**Fig.1 Functional stages of WiMAX PHY**
if \( j = 1 \mod 4 \), \( i = \Pi(j) = P_j + Q_j + i_0 \mod N \)

if \( j = 2 \mod 4 \), \( i = \Pi(j) = P_j + Q_j + i_0 \mod N \)

if \( j = 3 \mod 4 \), \( i = \Pi(j) = P_j + Q_j + i_0 \mod N \)  

(5)

and \( N \) must be a multiple of four, which is not a very restricting condition, with respect to flexibility.

In order to ensure the bijection property of \( \Pi \), the \( Q \) values are not just any values. A straightforward way to satisfy the bijection condition is to choose all \( Q \)'s as multiples of \( C \).

IV. PERFORMANCE OF DUO-BINARY TURBO CODES

Several optional channel coding schemes such as block turbo codes, convolutional turbo codes, and low density parity check (LDPC) codes are defined in IEEE 802.16e-2005. Of these optional channel coding modes, the convolutional turbo codes (CTC) are worth describing because of their superior performance and high popularity in other broadband wireless systems, such as HSDPA and WCDMA. As shown in Figure 2, WiMAX uses duo-binary turbo codes with a constituent recursive encoder of constraint length 4. In duo-binary turbo codes two consecutive bits from the uncoded bit sequence are sent to the encoder simultaneously.

Duo-binary turbo codes are a special case of nonbinary turbo codes, which have many advantages over conventional binary turbo codes [1]:

- **Better convergence**: The better convergence of the bi-dimensional iterative process is explained by a lower density of the erroneous paths in each dimension, reducing the correlation effects between the component decoders.

- **Larger minimum distances**: The nonbinary nature of the code adds one more degree of freedom in the design of permutations (interleaver)-intrasymbol permutation, which results in a larger minimum distance between codewords.

- **Less sensitivity to puncturing patterns**: In order to achieve code rates higher than 1/3 less redundancy, bits need to be punctured for nonbinary turbo codes, thus resulting in better performance of punctured codes.

A. 8-state Duo-Binary Turbo Code

The parameters of the component codes are:

\[
G = \begin{bmatrix} 1 & 0 & 1 \\ 1 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix}, \quad C = \begin{bmatrix} 1 & 1 \\ 0 & 1 \\ 0 & 1 \end{bmatrix}
\]

(6)

\[
R_1 = [1 \ 1 \ 0]\quad R_2 = [0 \ 0 \ 1]\]

(7)

The diagram of the encoder is described in Fig. 3. Redundancy vector \( R_2 \) is only used for coding rates less than 1/2. For coding rates higher than 1/2, puncturing is performed on redundancy bits in a regular periodical way, following the patterns that are described in [5]. These patterns are identical for both constituent encoders.

The permutation function \( i = \Pi(j) \) is performed in two steps. For \( j = 0, \ldots, N-1 \), we have the following:

**Step 1**: inversion of \( d_{j,1} \) and \( d_{j,2} \) in the data couple, if \( j \mod 2 = 0 \);

**Step 2**: this permutation step is described by a particular form of (5).

\[
i = (P_j + Q(j) + 1) \mod N \quad \text{with} \quad Q(j) = 0 \quad \text{if} \quad j \mod 4 = 0
\]

\[
Q(j) = \frac{N}{2} + P_1 \quad \text{if} \quad j \mod 4 = 1
\]

\[
Q(j) = P_2 \quad \text{if} \quad j \mod 4 = 2
\]

\[
Q(j) = \frac{N}{2} + P_3 \quad \text{if} \quad j \mod 4 = 3
\]

(8)

Value \( i_0 = 1 \) is added to the incremental relation in order to comply with the odd–even rule [11]. The disorder is instilled in the permutation function, according to the ARP principle, in two ways.

- A shift by \( N/2 \) is added for odd values of \( j \). This is done because the lowest subperiod of the code generator is one (see Fig. 3). The role of this additional increment is thus to spread to the full the possible errors associated with the shortest error patterns.

- \( P_1 \), \( P_2 \), and \( P_3 \) act as local additional pseudorandom fluctuations.

Fig.2 Duo-binary Turbo code

Fig.3 Structure of the 8-state encoder
B. 16-state Duo-Binary Turbo Code

The extension of the 8 state coding scheme to 16 states enables minimum distances to be increased by 50% on average. The parameters of the component code are:

\[
G = \begin{bmatrix}
0 & 0 & 1 & 1 \\
1 & 0 & 0 & 0 \\
0 & 1 & 0 & 0 \\
0 & 0 & 1 & 0
\end{bmatrix}, \quad C = \begin{bmatrix}
1 & 1 \\
0 & 1 \\
0 & 0 \\
0 & 0
\end{bmatrix}, \quad R = \begin{bmatrix}
1 & 1 & 1 & 0
\end{bmatrix}
\] (9)

The diagram of the encoder is described in Fig. 4. Puncturing is performed by redundancy in a periodic way, with identical patterns for both constituent encoders. It is usually regular, except when the puncturing period is a divisor of the LFSR period. For example, for coding rate 3/4, the puncturing period is chosen equal to six, with puncturing pattern [101000].

For this code, the permutation parameters have been carefully chosen, following the procedure described in [10], in order to guarantee a large minimum Hamming distance, even for high rates. The level-1 permutation is identical to the intrapermutation of the 8-state code. The level-2 intersymbol permutation is given by:

For \( j = 0, \ldots, N - 1 \)

\[
i = (Pj + Q(j) + 3) \mod N \quad \text{with}
\]

\[
Q(j) = 0 \quad \text{if } j \mod 4 = 0
\]

\[
Q(j) = Q_1 \quad \text{if } j \mod 4 = 1
\]

\[
Q(j) = 4Q_0 + Q_2 \quad \text{if } j \mod 4 = 2
\]

\[
Q(j) = 4Q_0 + Q_3 \quad \text{if } j \mod 4 = 3
\] (11)

The spirit in which this permutation was designed is the same as that already explained for the 8-state TC. The only difference is that the lowest subperiod of the 16-state generator is two, instead of one. That is why the additional shift (by 4Q_0) is applied consecutively, twice every four values of \( j \).

V. BLOCK LDPC CODES OF Wimax

The block irregular LDPC codes have competitive performance and provide flexibility and low encoding/decoding complexity [12]. The entire \( H \) matrix is composed of the same style of blocks with different cyclic shifts, which allows structured decoding and reduces decoder implementation complexity. Each base \( H \) matrix in block LDPC codes has 24 columns, simplifying the implementation. Having the same number of columns between code rates minimizes the number of different expansion factors that have to be supported. There are four rates supported: 1/2, 2/3, 3/4, and 5/6, and the base \( H \) matrices for these code rates are defined by systematic fundamental LDPC code of \( M_b \) by \( N_b \), where \( M_b \) is the number of rows in the base matrix and \( N_b \) is the number of columns in the base matrix. The following base matrices are specified: 12×24, 8×24, 6×24, and 4×24. The base model matrix is defined for the largest code length (\( N=2304 \)) of each code rate. The set of shifts in the base model matrix are used to determine the shift sizes for all other code lengths of the same code rate. Each base model matrix has \( N_b=24 \) block columns and \( M_b \) block rows. The expansion factor \( z \) is equal to \( N_b/24 \) for code length \( N \). The expansion factor varies from 24 to 96 in the increments of 4, yielding codes of different length. For instance, the code with length \( N=2304 \) has the expansion factor \( z=96 \). Thus, each LDPC code in the set of WiMax LDPC codes is defined by a matrix \( H \) as:

\[
H = \begin{bmatrix}
P_{1,1} & P_{1,2} & \ldots & P_{1,N_b} \\
P_{2,1} & P_{2,2} & \ldots & P_{2,N_b} \\
\vdots & \vdots & \ddots & \vdots \\
P_{M_b,1} & P_{M_b,2} & \ldots & P_{M_b,N_b}
\end{bmatrix}
\] (12)

where \( P_{ij} \) is one of a set of \( z \)-by-\( z \) cyclically right shifted identity matrices or a \( z \)-by-\( z \) zero matrix [12]. Each 1 in the base matrix \( H_b \) is replaced by a permuted identity matrix while each 0 in \( H_b \) is replaced by a negative value to denote a \( z \)-by-\( z \) zero matrix. The codeword length can be calculated by \( N=24z \) and ranges from \( N=576 \) to \( N=2304 \) bit with a granularity of 96 bit.

VI. SIMULATION AND RESULTS

We have simulated and compared LDPC codes and convolutional turbo codes intended for the WiMAX (IEEE 802.16e) forward error correcting schemes. For the CTC, iterative decoding was stopped after 10 iterations. Concerning the LDPC decoder, the maximum number of iterations of belief propagation decoding was limited to 100. The simulations were carried out for different code rates, lengths and modulation schemes in additive white Gaussian noise (AWGN) channel. Simulations were run to determine the performance of CTC and LDPC in AWGN channel with BPSK modulation. For each simulation, a curve showing the bit-error rate (BER) versus \( E_b/N_0 \) was computed.
In Fig. 5 we depicted the bit error rate (BER) versus $E_b/N_0$ for CTC with code length $N=576$ bits, code rate $R=1/2$ and different modulation schemes were computed. Fig. 6 shows the bit error performance of convolutional turbo code for various code rates, an input frame size of 288 bits and 64QAM used in 802.16e. Fig. 7 shows the comparison between LDPC codes and CTC codes with code rate of $R=1/2$, two modulation schemes (QPSK and 16QAM) and $N=576$ bits.

The performance of the CTC and the LDPC code is quite similar, whereby there is a ‘tenth of a dB advantage’ for the CTC. To reach a BER of $10^{-4}$ the LDPC code needs round 0.4 dB more for QPSK and round 0.6 dB more for 16QAM compared to CTC.

VI. CONCLUSIONS

The contribution of this paper is a study of WiMAX forward error correcting codes. It presents a validation and a discussion of these types of codes. Secondly, this paper presents an implementation of convolutional turbo codes and LDPC codes developed in Matlab. The performance gain using advanced coding techniques like CTC and LDPC is quite small for rate 1/2 codes. One reason for this is that the standard only provides short to moderate code lengths ($N \leq 2304$), which is the most crucial parameter for this class of codes. The performance of CTC and LDPC is about the same and by changing some decoding parameters the small advantage of one of them can be interchanged. Nevertheless, LDPC decoding is less complex than CTC decoding.

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