Buletinul Ştiinţific al Universităţii "Politehnica" din Timişoara

Seria ELECTRONICĂ şi TELECOMUNICAŢII
TRANSACTIONS on ELECTRONICS and COMMUNICATIONS

Tom 49(63), Fascicola 2, 2004

Performances of LDPC-Coded QAM Constellations Employing Non-Coded Bits

Vasile Bota, Mihaly Varga, Zsolt Polgar [1]

Abstract – The paper deals with the BER and Throughput vs. SNR performances of the LDPC-coded QAM signal constellations modulated on OFDM-type multicarrier transmissions. It analyzes, by means of computer simulation, the influences of the LDPC-code parameters (codeword length and rate) and the effects of mapping various ratios of coded and non-coded bits on QAM symbols, upon the SNR performances of these transmissions. Comparisons to the Shannon limits are presented as well.

Keywords: LDPC codes, Bit Error Rate, Throughput

I. INTRODUCTION

The LDPC codes are powerful block error correcting codes that provide high coding gains, comparable to the ones provided by the turbocodes for the same coding rate, at the expense of a simple encoding and moderate decoding complexity.

A. *Parameters and Types of LDPC Codes*

The Low Density Parity Check codes are block codes, basically defined by three parameters p, j, and k (integer numbers) that observe the following conditions [1]:

$$p \text{ is a prime integer;} \quad j < k \le p; \tag{1}$$

The codeword length N, the number of control bits C, the number of information bits J and the code ratio R_C are defined, respectively, by [1]:

$$N = k \cdot p; \quad C = j \cdot p; \quad J = (k-j) \cdot p; \quad R_C = (k-j)/k; \tag{2}$$

The control matrix, H, of a LDPC code is a 4-cycle free sparse matrix that might take three forms, which define the three types of LDPC codes: randomly generated by computer search [2], [3], by complete array-code control matrices [3] and by triangular-shaped array-code control matrices [4].

This paper considers only the codes generated by a triangular-shaped array-code control matrix H (jp x kp).

The generic form of the triangular-shaped matrix H_T is generated, see (3), by using an elementary matrix α, pxp, and the unity and null pxp matrices, I and 0:

$$H_T = \begin{bmatrix} I & I & I & \cdots & I & I & \cdots & I \\ 0 & I & \alpha & \cdots & \alpha^{j-2} & \alpha^{j-1} & \cdots & \alpha^{k-2} \\ 0 & 0 & I & \cdots & \alpha^{2(j-3)} & \alpha^{2(j-2)} & \cdots & \alpha^{2(k-3)} \\ \vdots & & & & \vdots & \vdots & & \vdots \\ 0 & 0 & 0 & \cdots & I & \alpha^{j-1} & \cdots & \alpha^{(j-1)(k-j)} \end{bmatrix} \tag{3}$$

Two different types of elementary matrices α that may be employed; they can be generated starting from the unity pxp matrix, either by shifting its rows downwards or upwards with one position.

The code parameter j equals the number of control equations a codeword bit is involved in, while code parameter k shows the number of bits involved in a control equation.

B. *Generation of the Control Matrix H_T*

The regular structure of the triangular-shaped matrix allows a systematic generation, starting from the code parameters j, k, p. Using the property of the matrix α, $\alpha^p = I$ (pxp), and the rule that gives the power of α inside the H_T (3), an algorithm to compute the indexes (row, column) defining the positions that take the logical value "1", in terms of the parameters j, k and p, can be determined. So, the binary matrix H_T is generated by filling with "1" only on the positions given by that algorithm.

C. *Shortening the LDPC Codes*

In order to adapt the number of information bits of the codeword to the information source or to adapt the codeword length, to the transmitted symbol-packet, the LDPC codes can be shortened [1].

The shortened code rate R' is smaller than the rate of the parent code R:

$$R' = J'/(J'+C) < R = J/(J+C); \tag{4}$$

D. *Encoding the LDPC Codes*

If the codeword is $v = [c_0, \ldots, c_{jp-1}, i_0, \ldots, i_{(k-j)p-1}]$, the control bits c_m are computed in terms of the information bits i_l by solving the C equations system:

---

[1] Technical University Cluj-Napoca, Faculty of Electronics and Telecommunications, Communications Department, str. G. Baritiu Nr.25, 400427, Cluj-Napoca Romania; e-mail: Vasile.Bota@com.utcluj.ro

$$H \cdot v^t = 0 \quad (5)$$

This approach has two major shortcomings:

- for great values of parameters j and/or p, C becomes large implying a significant computational load that increases the processing time and/or the hardware required by the implementation;
- it requires all information bits i_l, $l = 0, \ldots, (k-j)p-1$ at the same time; this requirement induces a one-codeword additional latency in the system.

These shortcomings may be avoided by a simpler and faster encoding method, described in [5].

E. *Bit-Mapping on the QAM Signal Constellation*

When the LDPC-coded bits are to be modulated on a b-bit/symbol QAM constellation, the b-tuple is mapped on the I and Q coordinates of the QAM vector, by splitting the b-tuple into two groups of b/2 bits, each group being assigned to one axis. The bits that are assigned to an axis are mapped to the amplitude levels of that axis according to a Gray encoding [1]. Since the transmission bit-loading might involve non-coded information bits, they are also mapped according to a separate Gray encoding, in order to maximize the distance between levels having the same non-coded bits. Therefore, the multibit assigned to an axis of the QAM constellation, coded and non-coded bits, is mapped according to a 2-level Gray encoding described in [3]. Fig. 1 presents an example of mapping b/2 = 4 bits on one axis (I or Q) of a QAM constellation.

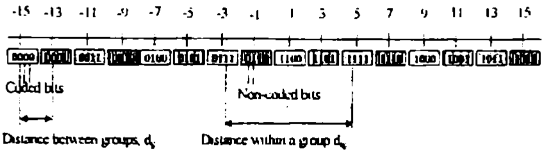

Fig. 1. Bit mapping on a QAM-constellation axis using the 2-level Gray encoding

The amplitude levels employed on each axis belong to the set A defined by:

$$A = \{A_l = 2l-(L_b-1);\ l = 0,1,\ldots L_b-1;\};\quad L_b = 2^{b/2};\quad (6)$$

This bit-mapping method, which allows only for the employment of the square QAM constellations, is simpler because the mapping is identical on both coordinates.

F. *Decoding the LDPC Codes*

The decoding of the LDPC codes is accomplished by using the message-passing algorithm (MP), as presented in [2], which will not be described here. This algorithm, based on the Bayes criterion, requires the previous computation of the *a posteriori* probabilities for every bit of a codeword, $F_n^0(r/0)$ and $F_n^1(r/1)$, where r denotes the received vector, n is the bit index and 0/1 denote the bit logical value.

Basically, the algorithm performs the decoding of a codeword by using the *a posteriori* probabilities of a group of N bits v' that is checked by means of syndrome-computation; if the syndrome equals zero, the algorithm considers v' to be the correct codeword; otherwise it performs another iteration adjusting the values of the a posteriori probabilities by using some internal values computed in the previous iteration. The maximum number of iterations allowed, B, is a parameter of the algorithm. The values of the *a posteriori* probabilities are previously extracted from the OFDM-demodulated coordinates I and Q of each QAM-symbol, by means of the soft-demapping procedure.

This algorithm does not search for the closest codeword compared to the received sequence, but tries to correct every bit. Due to this property, the number of error bits after the decoding is always smaller than the one of error bits prior the decoding, when the algorithm is convergent. Extensive simulation performed by the authors confirmed this property, which might lead to the decrease of error-packet length that should be corrected by the RS code that follows the LDPC or convolutional codes in many applications.

G. *Soft-Demapping*

Because the MP algorithm requires the *a posteriori* probabilities of each bit and the received vector carries more bits, a soft-demapping [3] is required in order to provide the $F_n^0(0/r)$ and $F_n^1(1/r)$ probabilities of each bit mapped the received vector.

For multibit/symbol modulations, the two probabilities of each bit are extracted, from the received level on the I or Q branches, by using (7) that gives the probability of bit b_j to be „1" when the demodulated level on a branch equals r and the channel is AWGN [1]:

$$F_j^1 = \frac{\sum_{l=1}^{2^{b/2}} \exp\left(-\frac{(r-L(l))^2}{2\sigma^2}\right)\cdot b_{lj}}{\sum_{l=1}^{2^{b/2}} \exp\left(-\frac{(r-L(l))^2}{2\sigma^2}\right)};\ j = 0,\ldots,b/2-1; \quad (7)$$

In (7) b_{lj} denotes the logical value of j-th bit of the l-th modulating level of the I or Q branch of the demodulated vector. A similar expression is derived for F_j^0 and the two values are normalized to their sum. The soft-demapping requires a previous estimation of the noise variance σ; computer simulations run by the authors showed that estimation errors of less than 2 dB, between the actual channel noise variance and the one stored in the soft-demapper, lead to insignificant decreases of the decoder performances.

H. *Soft Decision of the Non-Coded Information Bits*

The information non-coded bits mapped on a QAM symbol can be decided by two methods, namely:

- **hard decision**, applying the Bayes criterion to the probabilities provided by the soft-demapping; this method does not employ the information provided by decoding the coded bits placed on the same tone during the same symbol period.

- **soft decision**, that considers the information provided by the decoding of the coded bits mapped on the same QAM symbol and tone.

Basically, the optimal decision memorizes the received level r and, using the decoded bits provided by the LDPC decoder, selects the closest (in the d_E sense) level that was mapped with the same decoded bits, see fig. 2. This method provides lower BER of the non-coded bits, as resulted from simulations performed by the authors, but may error the non-coded bits if the corresponding coded bits were wrongly decoded.

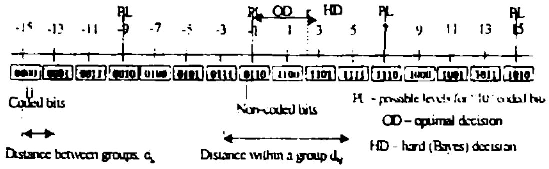

Fig 2 Optimal decision of non-coded bits for "10" coded bits

I. *Bit-loading on Coded-QAM OFDM Employing Non-Coded Bits*

Denoting by T the number of available sub-carriers, by D_{OFDM} the OFDM symbolrate, by N_{ci} and N_{ni} the numbers of coded and non-coded bits on the i-th subcarrier and by R_C the LDPC code rate, the nominal payload D_n of the OFDM transmission would be (8). The number of bits actually carried by each QAM-symbol equals $N_c + N_{ni}$.

$$D_n = D_{OFDM} T(N_{ci} R_C + N_{ni}); \qquad (8)$$

The rate of the coded QAM-modulation is com uted by:

$$R_{CM} = \frac{N_{ci} \cdot R_C + N_{ni}}{N_{ci} + N_{ni}}; \qquad (9)$$

II. BER PERFORMANCES OF THE LDPC-CODED QAM CONSTELLATIONS

Due to the complexity of the theoretical evaluation of the BER provided by the LDPC coded QAM constellations that are also loaded with non-coded bits, the authors developed a computer simulation environment that was employed to derive the results shown below.

A *Simulation Environment and Parameters*

The simulation program that implements the LDPC-coded multi-carrier transmissions, allows the following parameters to be set: LDPC code parameters (k, j, p), number of sub-carrier groups G, number of sub-carriers within a group T_i, bit-loading for each group (number of coded bits - N_{ci}, number of non-coded bits - N_{ni}), maximum number B of iterations/codeword of the LDPC decoder, range and step of SNR, Rayleigh channel model, test length.

It displays the BER values and the BER vs. SNR characteristic for the selected SNR range, the number of coded bits error after the decoding of each codeword, and the number of non-coded bits decided by soft-demapping. It also displays the throughput of the transmission for a defined packet dimension.

The simulations were performed on a test of 10^6 information bits and the maximum number of B = 15 iterations/codeword for the decoding algorithm.

B. *Effects of the Coding Rate upon the BER Performances of LDPC-Coded QAM Constellations*

As shown in (2) the coding rate might be changed, without changing the codeword length, by changing the parameter j. Considering k=14 and p = 31, so that a short codeword N = 434 bits (see (2)) is used, a family F_1 of LDPC codes with R_C ranging from 0.78 to 0.21 is displayed in table 1. The family includes the non-coded configuration for comparison.

Table 1. Parameters of F_1 LDPC codes; k = 14, p = 31

F_1	j	N	C	R
C_{11}	Non-coded			1
C_{12}	3	434	93	0.78
C_{13}	7	434	217	0.50
C_{14}	9	434	279	0.35
C_{15}	11	434	341	0.21

In order to evaluate the correction capability of these codes, the BER vs. SNR performances were evaluated employing a 2-PSK constellation. And are shown in fig. 3.

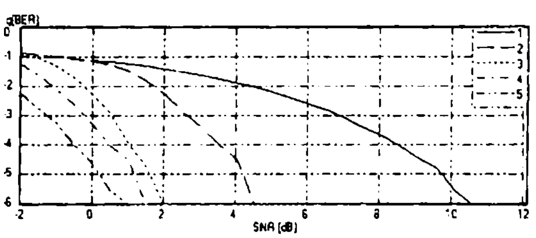

Fig. 3. BER vs SNR of 2-PSK coded with the LDPC codes of family F1

The coding gains provided by the F_1 codes, at BER = $1 \cdot 10^{-6}$, and their rates are displayed in table 2.

Table 2. Coding gains provided by the F_1 codes

Code	C_{11}	C_{12}	C_{13}	C_{14}	C_{15}
CG (dB)	0	6	8.5	9	9.5
R_C	1	0.78	0.5	0.35	0.21

As expected, the coding gains provided increases with the decrease of the coding rate.

The coding gain provided by the "parent code (R_C = 0.78) is about 6.5 dB, comparable to the ones of convolutional codes of R = ½ and K = 5 – 7.

The R = ½ LDPC code provides a coding gain of 8.5 dB, larger than the one the convolutional codes.

Taking into account the fact that the MP decoder has about the same implementation complexity as the 64-state Viterbi decoder we may say that the LDPC codes provide about the same coding gain as the convolutional code, at a higher coding rate.
Comparing the coding gains with the ones of the turbo codes, the R_C = ½ LDPC code ensures BER =10-6 at a SNR = 2 dB, about 1 dB higher than the turbo-codes [6], but it requires only one MAP decoder instead of two decoders (MAP + MLD) and a de-interleaver.
By decreasing the code rate by changing the code parameter j additional coding gains of up to 3.5 dB can be obtained. The code C_{15} (R_C = 0.21) requires a SNR = 1dB to provide a BER = 10^{-6}.
In order to evaluate how close these codes are to the Shannon limit, we recall that the maximum spectral efficiency β_{wMi}, which could be provided by a code of rate R_{ci} in an error-free transmission across an f_s bandwidth, is:

$$\beta_{wMi} = \frac{1}{2} lg_2(1 + R_{ci} \frac{S}{N}) \; [\; ps/Hz]; \qquad (10)$$

Assuming that the error-free transmission is accomplished for SNRs > SNR_{0i}, where:

$$BER\,(SNR_{0i}) = 1 \cdot 10^{-5}; \qquad (11)$$

we may compute the maximum spectral efficiency, β_{wM0i}, that could be accomplished by a code with rate R_{ci}, by using (10) and SNR_{0i} obtained by simulation (see fig.3). This value, compared to the actual spectral efficiency β_{wi}, provided by the code (which equals R_{ci}) indicates how far is the code from the theoretical limit.
Another evaluation can be performed by computing the minimum signal/noise ratio, SNR_{mi}, for which β_{wi} could be obtained. The difference $\Delta SNR_i = |SNR_{mi} - SNR_{0i}|$ shows the "quality" of the code in terms of the SNR.
The values of these parameters for the codes of table 1 are shown in table 3

Table 3. Ideal and actual performances for codes of family F_1

Code C_i	β_{wi} [bps/Hz]	β_{wMi} [bps/Hz]	SNR_{0i} [dB]	SNR_{mi} [dB]	ΔSNR_i [dB]
C_{12}	0.785	1.59	4.1	-0.35	4.45
C_{13}	0.5	0.80	1.7	-0.82	2.52
C_{14}	0.357	0.546	1.1	-1.04	2.14
C_{15}	0.214	0.29	0.1	-1.26	1.36

As expected, codes of a certain length come closer to the theoretical limits as their rate decreases. Results of tables 2 and 3 show that rather short codes, easy to implement, are close enough to the theoretical maximum performances.
The performances of the LDPC codes might be improved by increasing the maximum number of iterations/codeword of the MP decoder; simulations showed that increasing B = 25, leads to extra coding gains of 0.5-1 dB, at the expense of a longer processing time required.

C. *Effects of the Codeword Length upon the BER Performances of LDPC-Coded QAM Constellations*

In order to evaluate the effects of the codeword length upon its performances we consider a family of codes F_2, see table 4, with R_C = 0.214 (as code C_{15} of F_1). The code word length N is modified by means of parameter p, ranging from 23 to 73.

Table 4. Codes of rate R_c = 0.214 and various codeword length N – family F_2

Code	k	j	p	N [bits]
C_{151}	14	11	23	322
C_{151}	14	11	31	434
C_{153}	14	11	43	602
C_{154}	14	11	53	742
C_{155}	14	11	73	1022

The BER vs. SNR performances of these codes are displayed in fig.4.

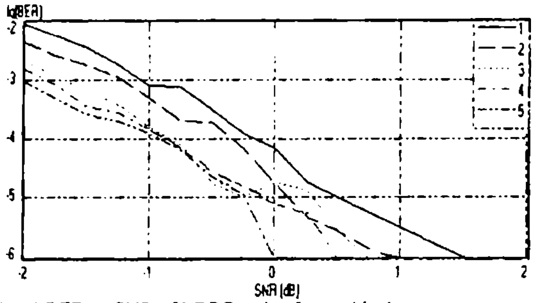

Fig. 4 BER vs SNR of LDPC codes from table 4

The results in fig.3 indicate that by increasing the codeword length, extra coding gain, can be provided at the same coding rate, but at the expense of a more difficult implementation. The extra coding gain may be as high as 1.5 dB, see codes C_{152} and C_{155} in fig.4 on one hand and code C_{15} in fig.3 on the other. The code C_{155} provides a total 11 dB coding gain.
Increasing the codeword length would also bring the code performances closer to the Shannon limit. Computing the parameters defined in table 3 for the codes of family F_2 we get the values of table 5.

Table 5. Ideal and actual performances for codes of family F_2

Code	B_{wi} [bps/Hz]	β_{wMi} [bps/Hz]	SNR_{0i} [dB]	SNR_{mi} [dB]	ΔSNR_i [dB]
C_{151}	0.214	0.352	0.5	-1.26	1.76
C_{152}	0.214	0.291	0.2	-1.26	1.46
C_{153}	0.214	0.280	0.0	-1.26	1.26
C_{154}	0.214	0.274	-0.1	-1.26	1.16
C_{155}	0.214	0.268	-0.25	-1.26	1.01

The longest code of family F_2, C_{155}, is about 1 dB away from the Shannon limit, for BER < $1 \cdot 10^{-5}$.
Simulations performed by the authors showed that the coding gains provided by the LDPC-coded QAM constellations are about the same, compared to the same non-coded QAM constellations, regardless the constellation employed.

III. LDPC-CODED QAM CONFIGURATIONS EMPLOYING NON-CODED BITS

The employment of error-correcting codes leads to a decrease of throughput provided by the coded QAM constellation. This occurs due to the control bits inserted by the code, and the throughput decrease is higher for low-rate codes that secure a good correction capability.

In order to reach a reasonable trade-off between the correction capability and the coding rate, which affects significantly the throughput, each QAM symbol is loaded with n_{ci} coded bits and with n_{ni} non-coded bits.

Considering a LDPC code with an R_C coding rate the coding rate of the configuration employing coded and non-coded bits is expressed by (9).

The coded and non-code bits are mapped on the I and Q coordinates (6) of the QAM symbol using the 2-level Gray mapping, see fig.1, on each axis.

A. *BER Performances of the LDPC-Coded QAM Configurations Employing Non-Coded Bits*

To evaluate the effects of employing non-coded bits, a family F_3 of possible LDPC-coded configurations using non-coded bits based on the 256-QAM (N_{ci} + N_{ni} = 8 bits) are presented in table 6 together with their coding rates R_{CM} and coding gains C_G. The LDPC code employed is (k = 14, j = 3, p = 31; R_C = 0.78), which provided a 6.5 dB coding gain on a 2-PSK modulation (code C_{12} in table 1 and fig. 3). The BER vs. SNR performances of this family are presented in fig. 5.

Table 6. Coded QAM Configurations of Family F_3

Cfg.	N_{ci}	N_{ni}	R_{CM}	C_G
1	0	8	1	-
2	2	6	0.945	5
3	4	4	0.890	6
4	6	2	0.835	6.5
5	8	0	0.780	7

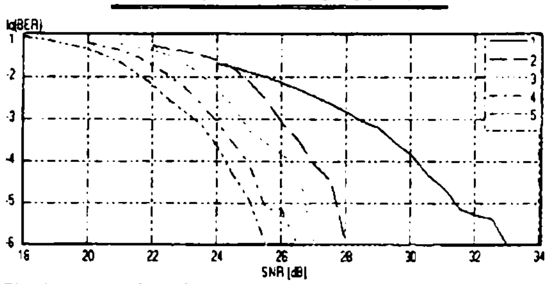

Fig. 5. BER vs. SNR for configurations of F_3 defined in table 6

As shown in table 6, the employment of non-coded bits leads to a significantly increase of the coding rate, at the expense of a coding gain decrease of about 1-2 dB. The coding rate increases starting from 0.78 up to 0.945, in terms of the proportion of non-coded bits within the 8 bits loaded on a QAM symbol.

The relatively small decrease of the coding gain could be explained by the "protection" of the non-coded provided by the 2-level Gray mapping and by their soft decoding. Fig. 6 shows the BER vs. SNR curves of the coded and non-coded bits before the MP decoding (for the coded bits) and before the soft decision (for the non-coded bits), and after these decoders. The BER prior to decoding was obtained by using a hard Bayes decision that employs the *a posteriori* probabilities provided by the soft-demapping. The curves correspond to configuration no.3 from table 6.

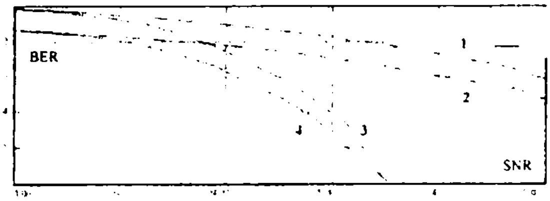

Fig.6 BER vs. SNR of the coded and non-coded bits of configuration 3 from table 6, line 1 - coded bits Bayes decision, line 2 - non-coded bits Bayes decision; line 3 - coded bits MP decision; line 4 - non-coded bits soft decision

As shown in figure 6, the non-coded bits have lower BER than the coded bits loaded on the same QAM symbols, both before their decoding (line 1 vs. line 2) and after it (line 3 vs. line 4). This is due to the 2-level Gray mapping of the coded and non-coded bits.

Fig. 6 shows that the number of error bits is always smaller after the decoding process, than before it. This, combined with additional simulations performed by the authors indicate that the two decoders might require some smaller outer codes (small RS or even BCH) in FEC schemes employing concatenated codes.

The spectral efficiencies of the configurations employing non-coded bits are higher than the ones of the proximity to the Shannon limit of the configurations from family F_3 are displayed in table 7.

Table 7. Ideal and actual performances of coded configurations from family F_3

Cfg	R_{CMi}	β_{wi} [bps/Hz]	β_{wMi} [bps/Hz]	SNR_{0i} [dB]	SNR_{mi} [dB]	ΔSNR_i [dB]
1	1	8	10.46	31.5	24	7.5
2	0.945	7.52	9.05	27.5	22.9	4.6
3	0.890	7.12	8.64	26.5	21.9	4.6
4	0.835	6.68	8.21	25.5	20.9	4.6
5	0.780	6.24	8.31	25.0	19.8	5.2

The spectral efficiencies of the coded configurations carrying non-coded bits (lines 2, 3, 4 in table 7) are higher than the ones of the coded configuration with no non-coded bits (line 5). Also they are closer, in the SNR sense, to the Shannon limit.

C. *Throughput Performances of the LDPC-Coded QAM Configurations Employing Non-Coded Bits*

The employment of the non-coded bits decreases the coding gain leading to a higher BER at a given SNR, on one hand, and increases the coding rate leading to more information bits transmitted, on the other. The effects of these two factors upon the throughput provided by such configurations are shown below.

For throughput evaluation we considered an OFDM transmission that is based on a user-bin of T_i tones and F OFDM symbol periods, thus containing $T_{si} = T_i \times F$ QAM symbols (packet length), out of which only A_{si} are "active" symbols being used for the payload. The cyclic prefix is denoted by G and represents a fraction of the symbol period, the number of bits per QAM symbol is n_i (it defines the QAM constellation employed), the bin rate is D_b and the CRC (required for channel estimation-prediction) is t bits long.

Using (8) the nominal payload for a non-coded constellation i (n_i), i.e. the maximum value for the payload when SNR is very high, is:

$$D_{ni} = \frac{1}{1+G} \cdot \frac{A_s}{T_s} \cdot D_b \cdot T_s \cdot n_i \cdot \left(1 - \frac{t}{A_s \cdot n_i}\right); \quad (12)$$

As for the coded configurations, their nominal payload is computed using (8) and is given by (13). There should be noted that the number of control bits of a codeword jp should be divided by a constant that indicates the number of bins on which a codeword is loaded.

$$D_{ci} = \frac{1}{1+G} \cdot \frac{A_s}{T_s} \cdot D_b \cdot T_s \cdot n_i \cdot \left(1 - \frac{jp}{k_i \cdot A_s \cdot n_i}\right); \quad (13)$$

Considering an adapted version of the values proposed in [7], namely D_b = 1500 bins/sec, T_s = 120 symbols, A_s = 108 symbols, G = 0.11, t = 8 bits and n_i = 8 the nominal payloads (12, 13) of the configurations of table 7 are listed in table 8. The constants k_i are respectively 2, 1, 2/3 and ½.

Table 8. Nominal payload of configurations from table 6

Cfg	1	2	3	4	5
D_{ni} (kbps)	1156	1104	1041	979	916

The non-coded throughput Θ_{ni} is computed considering only the correctly received bins given by the error-bin probability $B_{in}ER_{ni}$, and is:

$$\theta_{ni} = D_{ni}(1 - BinER_{ni}) \quad (14)$$

The throughput Θ_{ci} of the coded configurations is computed using (14), but the error-rate of the coded bins $B_{in}ER_{ci}$ is employed.

The throughput Θ_{ni} or Θ_{ci} vs. SNR curves of the transmissions employing the configurations of table 6 were obtained by simulations and are displayed in fig. 7.

Each configuration exhibits a range of SNR where it provides the best performance, out the entire family.

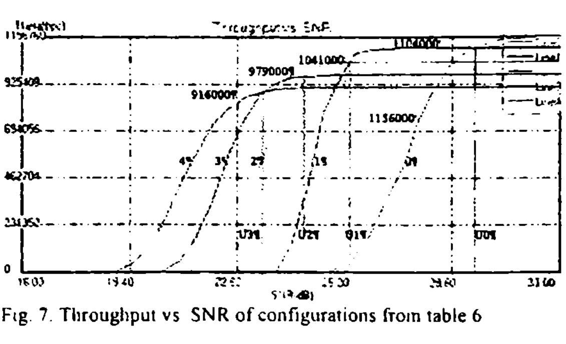

Fig. 7. Throughput vs SNR of configurations from table 6

The SNR ranges of optimum for each configuration are separated by the thresholds U_i.

The family F_3 together with the thresholds U_i may be employed into an adaptive modulation scheme that provides the best throughput according to the channel current SNR.

This scheme is very simple since it changes only the bit-loading and employs the same QAM constellation and LDPC code. Despite its simplicity, it provides a reasonable throughput over a SNR range of about 10-12 dB.

V. CONCLUSIONS

The array-based LDPC codes employed in the present paper allow for a simple encoding and a moderate complexity decoding, compared to the turbo codes. The LDPC decoder requires the a priori knowledge of the channel's noise variance.

The BER performances of the LDPC-coded QAM modulations are close to the ones provided by the similar modulations coded with turbo codes at the same rate and the same number of iterations per code word. A BER = 10^{-6} at SNR = 2 dB in an AWGN channel can be obtained by an R = 0.5 LDPC-coded 2-PSK, ensuring a coding gain of about 9 dB.

A very flexible rate changing LDPC-coded scheme can be obtained by using a bit-loading that combines coded and non-coded bits. This approach allows for significant increases of the coding rate at the expense of rather small coding gain losses.

Due to the behavior of the LDPC-decoding algorithm and to the soft-decision of the non-coded bits, the authors estimate that small and high rate RS outer codes should be employed in FEC schemes based on concatenated codes.

The LDPC-coded modulation scheme proposed in the paper ensures coding gains of about 6 dB, compared to the correspondent non-coded scheme.

It also provides an increased throughput and offers the possibility of adaptive employment according to the channel SNR.

REFERENCES (SELECTED)

[1] - ITU-T, *"LDPC codes for G.dmt.bis and G.lite.bis,"* Temporary Document CF-060.

[2] - D.J.C. McKay, *"Good error-correcting codes based on very sparse matrices,"* IEEE Trans. on Information Theory, vol. 45, No. 2, March, 1999.

[3] - R. Gallagher, *"Low-density parity-check codes"* IRE Trans. Information. Theory, vol. IT-8, January 1962.

[4] - ITU-T, *"Low-density parity-check codes for DSL transmission,"* Temporary Document BI-095.

[5] - V.Bota, Zs.Polgar, M.Varga, *"BER Performance of the QAM Modulation Coded with LDPC Codes"*, International Symposium Etc. '2002", *Buletinul Universităţii "Politehnica"*, Tom 47 (61), 2002, Fascicola 1, 2, 2002, pp. 104.

[6] - Cl. Berrou, A.Glavieux, *"Near Optimum Error Correcting Coding And Decoding: Turbo-Codes"*, IEEE Transactions on Communications, vol.44, pp. 1261-1271, October 1996

[7] - W. Wang, M.Sternad, T. Ottosson, A. Ahlen, A. Svensson, - *„Impact of Multiuser Diversity and Channel Variability on Adaptive OFDM"*, Proceedings of COST 289 *"Spectral and Power Efficient Broadband Communications"* Seminar, Budapest 2004.