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Frequency-Selective Link Adaptation using Duo-Binary Turbo Codes in OFDM Systems

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***Abstract*— This paper is devoted to a novel frequency-selective *link adaptation* (LA) strategy developed in the WINNER Integrated Project, which aims to provide a new radio interface for future 4G wireless communication systems. For a single-user adaptation of modulation and code rate in a coded OFDM system, the presented bit-loading algorithm offers performance results very close to the optimum solution of the well-known Hughes-Hartogs bit and power loading algorithm, which is verified by simulations. Efficient rate-compatible punctured *duo-binary turbo codes* (DBTC) have been used in evaluations. The adaptation is performed in two steps: first, adaptive QAM mapping is done on subcarrier groups in frequency direction, according to a short-term channel quality information. Then, an average code rate is calculated for the entire set of used subcarriers with heterogeneous modulation alphabets. This method results in moderate system complexity while offering excellent performance.**

I. Introduction

One of the main goals of next generation (4G) wireless communication systems is to provide users with a flexible bulletproof high speed connection, which can be used by the same mobile terminal in various environments. Such an ubiquitous approach is currently investigated in the WINNER (*Wireless World Initiative New Radio*) Integrated Project, within the European Union 6^{th} Framework Programme.

Link adaptation to varying radio channel conditions is nowadays one of the key technologies required for efficient use of the available spectrum. The idea of LA in the form of adaptive modulation was proposed a few decades ago by J. F. Hayes [1]. Since then, a lot of research has been done in this area. Following the famous Hughes-Hartogs bit and power loading algorithm [2], a set of sub-optimum algorithms that approach the *water-filling* concept [3] has been proposed for an *asymmetric digital subscriber line* (ADSL) wired technology [4]–[6]. These algorithms have been designed mainly for uncoded multicarrier transmission, but recently they were reused in a coded OFDM (*orthogonal frequency division multiplexing*) wireless transmission over frequency-selective fading channels [7]–[9]. However, there is one common problem with the use of these algorithms in wireless systems. The so far proposed bit-loading schemes for coded OFDM transmission assumed either a fixed code rate transmission (only modulation and power is adapted), or a joint code rate and modulation adaptation, but with quite short variable-length codewords. This prohibits the use of strong channel codes like Turbo or LDPC (*low-density parity check*) codes that would require long codeword sizes to be efficient.

This problem has been elegantly solved in the *mutual-information based adaptive coding and modulation* (MI-ACM) approach [10], which is specially designed for the WINNER frequency-selective adaptive transmission using efficient powerful coding schemes [12]. A scheme based on similar ideas has been proposed recently independently by Li and Ryan [11]. Evaluation of the MI-ACM algorithm using rate-compatible duo-binary turbo codes [13] is the main target of this paper.

The rest of this work is organised as follows. Section II describes briefly the investigated system model. The next two sections present the modulation and coding schemes used in adaptation, and the MI-ACM algorithm. Simulation results are discussed in Section V. Finally, conclusions are drawn in Section VI.

II. System Model

The investigated adaptive OFDM system scheme is depicted in Figure 1. In the transmitter, an input data flow from one user is first grouped into N_{cw} packets of K bits, which are encoded by a DBTC encoder and punctured according to an adapted code rate R. Such a group of encoded packets, each of length $N = K/R$, is then bit-interleaved and mapped onto a set of N_{ch} so-called *chunks*, which have been already assigned to the user by a scheduler. The chunk is the smallest time-frequency unit used in the frequency-adaptive transmission and consists of n_{f} adjacent subcarriers and n_{t} OFDM symbols.

Next, the used chunks are modulated according to the MI-ACM algorithm (cf. Sec. IV) and are optionally power-loaded. The modulation order, i.e. the number of bits per QAM symbol, determined by the algorithm is denoted as $\mathbf{b} = [b_1, b_2, \ldots, b_{N_{\mathrm{ch}}}]$, where $b_n \in \mathcal{B} = \{1, 2, \ldots, 8\}$. The number of coded bits at the output of the interleaver is thus

$$N_{\mathrm{cb}} = N_{\mathrm{q}} \sum_{n=1}^{N_{\mathrm{ch}}} b_n, \qquad (1)$$

where $N_{\mathrm{q}} = n_{\mathrm{f}} n_{\mathrm{t}}$ stands for the number of QAM symbols per chunk. This leads to $N_{\mathrm{cw}} = \lfloor N_{\mathrm{cb}}/N \rfloor$ codewords. The missing

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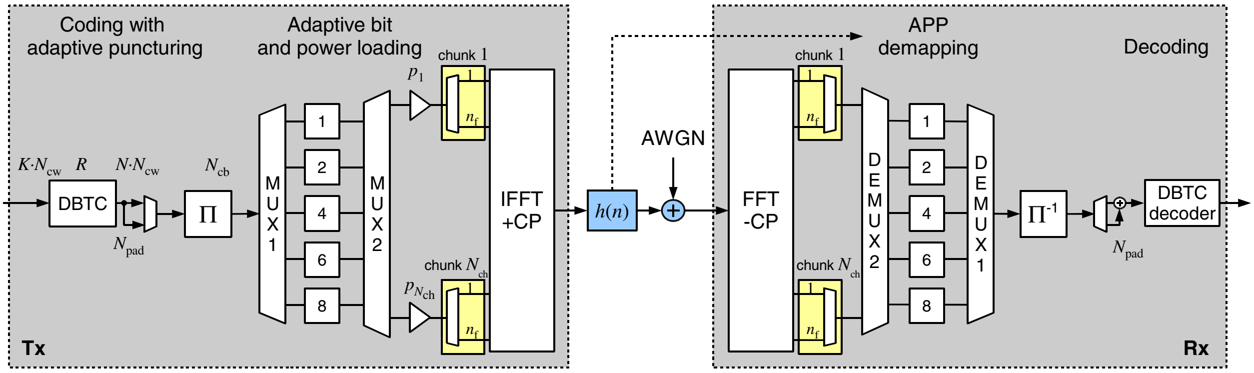

Fig. 1. Simplified scheme of the simulated adaptive system

$N_{\text{pad}} = N_{\text{cb}} - N_{\text{cw}}N$ bits are inserted before interleaving as a cyclic repetition of the first N_{pad} bits of the FEC (*forward error correction*) block.

Next, n_{t} OFDM symbols are created from the modulated chunks with the use of IFFT. Finally, these OFDM symbols are cyclic-prefixed and sent through a frequency-selective fading channel.

The receiver performs an inverse processing, i.e. it removes the cyclic prefixes, demodulates OFDM symbols, performs soft demapping of each chunk, deinterleaves the obtained log-likelihood ratios, and finally divides such deinterleaved block into packets, which are decoded in the DBTC decoder. Note that the system model in Figure 1 allows for iterative demapping where the APP demapper plays the role of the inner code [9].

The link adaptation approach realised by the MI-ACM algorithm allows to adapt the modulation scheme per chunk while the same code rate is chosen for all chunks that belong to the same user. This design choice is motivated by the following characteristics of the channel and the properties of state-of-the-art FEC schemes:

- The channel gain variation between chunks belonging to the same user is much lower than the variation between different users.
- The minimum codeword length should be in the order of several hundred bits in order to make best use of modern coding schemes like DBTC or LDPCC.
- By employing one decoder for all chunks, the decoding complexity is moderate.

An alternative approach of adapting both the modulation and coding scheme per chunk has also been investigated. However, as expected, this latter approach suffers from the limited coding gain due to the shorter codewords and is thus not considered in the following.

For the choice of the codeword length, there are basically two possibilities: (1) The codeword length is determined by the link-adaptation algorithm depending on the selected modulation schemes on the available chunks. (2) The codeword size is determined by upper-layer algorithms in order to fit to the MAC frame structure.

In the following, we consider the second possibility since this results in a lower system complexity.

TABLE I
SNR THRESHOLDS $\gamma_{m,i}^{(2)}$ FOR $K = 1152$ (IN DB)

b \ R	1/3	2/5	1/2	2/3	3/4	4/5	6/7
1	-3.64	-2.70	-1.44	0.56	1.68	2.37	3.34
2	-0.63	0.31	1.58	3.57	4.69	5.38	6.35
3	3.00	4.11	5.57	7.73	8.87	9.56	10.55
4	4.21	5.35	6.92	9.40	10.79	11.60	12.78
5	6.59	7.90	9.70	12.40	13.81	14.70	15.82
6	8.05	9.38	11.35	14.45	16.11	17.09	18.39
7	10.15	11.64	13.78	17.08	18.75	19.81	21.20
8	11.37	13.12	15.51	19.24	21.19	22.37	23.89

III. Modulation and Coding Schemes

A. Performance of Modulation and Coding Schemes

The performance of the *modulation and coding schemes* (MCS) has been evaluated for all relevant combinations of code rate, codeword length and constellation sizes. The simulation setup for the generation of reference curves for the bit and *codeword error ratio* (CWER) encompassed an FEC coder, a random interleaver and a QAM mapper over an AWGN channel. The selected channel code is a DBTC [12], [14] with mother code rate $R_{\text{c}} = 1/3$, from which the following code rates are derived by rate-compatible puncturing:

$$R \in \mathcal{R} = \left\{ \frac{1}{3}, \frac{2}{5}, \frac{1}{2}, \frac{2}{3}, \frac{3}{4}, \frac{4}{5}, \frac{6}{7} \right\} \tag{2}$$

For QAM mapping, all constellation sizes from BPSK to 256-QAM have been implemented. Gray labeling is applied to the square constellations, while for non-square constellations, where this is not possible, quasi-Gray labeling is used. At the receiver side, APP soft demapping is applied, using the max-log approximation for 16-QAM and larger constellations.

The required SNR limits for all MCS and codewords with $K = 1152$ information bits, at a target CWER of 0.01, are listed in Table I. For $K = 288$, slightly higher SNR limits are required.

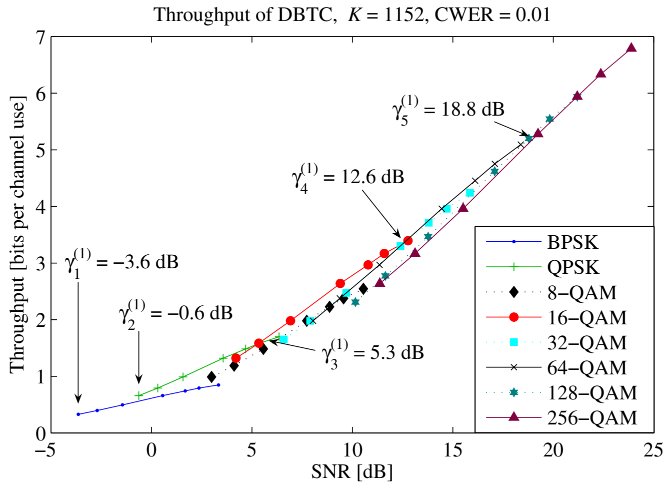

Fig. 2. Throughput for main MCS at a target CWER of 0.01. The depicted SNR thresholds indicate the switching points for the modulation format.

Based on the thresholds from Table I, the throughput of each MCS can be illustrated as a function of the SNR, as it is done in Figure 2 for square constellations. From this diagram, the thresholds $\gamma_i^{(1)}$ are derived, which determine the modulation scheme in the MI-ACM algorithm (cf. Sec. IV).

B. Addition of Non-Square Constellations

In order to obtain a finer granularity of the adaptation, non-square QAM constellations ($b \in \{3, 5, 7\}$) have been considered (cf. Tab. I). For 8-QAM the constellation with maximum distance for a given energy has been chosen, while for 32- and 128-QAM the well-known cross constellations have been employed [15].

The SNR thresholds for non-square constellations are depicted in Fig. 2 with dotted lines. It can be observed that the throughput with 8-QAM and 32-QAM is always below that of the square constellations. Only 128-QAM achieves, in the range between 18.5 dB and 21.2 dB, a slightly higher throughput than 64- and 256-QAM. This indicates that the possible gain from including 128-QAM is very small, while 8- and 32-QAM are excluded *a priori*.

IV. Bit-Loading

Since the guard interval is chosen sufficiently long for the relevant channel models (cf. Sec. V), we can describe the channel by one transfer coefficient H_n per subcarrier. Based on these coefficients, the *channel gain to noise ratio* (CNR) is defined per chunk as

$$T_n = f^{-1}\left(\frac{1}{n_\mathrm{f}}\sum_{i=1}^{n_\mathrm{f}} f\left(\frac{|H_{n_\mathrm{f}(n-1)+i}|^2}{N_0}\right)\right),\ n = 1, \ldots, N_\mathrm{ch}, \tag{3}$$

where we use the mutual information based averaging with $f(x) = \mathrm{ld}(1+x)$ to obtain *one* CNR per chunk. The channel is assumed constant for the duration of one chunk (block fading), and codewords are transmitted completely within the allocated chunks.

In the following, we analyze the performance of two bit-loading algorithms:

1) The Hughes-Hartogs algorithm [2], which is known to provide the optimum solution if power-loading is included. It is a discrete approximation to water-filling, which yields the power-loading that achieves the highest channel capacity. Although originally conceived for uncoded QAM, the Hughes-Hartogs algorithm can be straightforwardly applied to coded systems by using look-up tables with SNR thresholds. Here, we apply this algorithm separately to each column of Table I and choose the code rate which leads to the highest throughput.
2) The constant-power mutual-information based adaptive coding and modulation (MI-ACM) method developed in the WINNER project. This method applies the same transmit power to all used chunks and is based on averaging of code rates [10].

The latter algorithm is specified in detail in Figure 3. We can assume without loss of generality that chunks are ordered such that $T_1 \le T_2 \le \cdots \le T_{N_\mathrm{ch}}$, i.e. the worst chunk comes first. If the SNR of a chunk is lower than $\gamma_1^{(1)}$, then this chunk is excluded and the power is redistributed over the other chunks. Otherwise, first the constellation is chosen according to the levels in Figure 2 and then a local (virtual) code rate is computed by a linear interpolation of the values from Table I. Finally, a code rate out of $\mathcal{R}$ is selected. For a detailed justification of this algorithm please refer to [10].

For comparison, we also consider non frequency-selective adaptation. In this case, the same MCS is applied to all chunks. First, one CNR is computed by the following averaging method over all transfer coefficients [16]:

$$T = \beta \cdot f^{-1}\left(\frac{1}{N_\mathrm{c}}\sum_{n=1}^{N_\mathrm{c}} f\left(\frac{|H_n|^2}{N_0}\right)\right) + (1-\beta)\min_n \frac{|H_n|^2}{N_0} \tag{4}$$

Then, the appropriate MCS is selected according to Table I. The throughput obtained with this simpler adaptation method provides a baseline for the frequency adaptive scheme.

V. Simulation Results and Discussion

All simulations in this paper are based on the B1 NLoS channel model defined in [17] and frame structure and parameters for the TDD mode of the WINNER system concept, as specified in Table II. Perfect *channel state information* (CSI) is assumed at both the transmitter and receiver. Moreover, the channel is modelled as a block fading process, which is updated every n_t OFDM symbols.

A reference curve and an upper bound from information theory are depicted together with the achieved throughput. The reference curve refers to the ergodic capacity of a Rayleigh fading channel with inputs taken out of the 256-QAM constellation. The capacity of the Rayleigh fading channel with perfect adaptation (i.e. temporal water-filling) [18] constitutes an upper bound for the cases where the chunks are assigned randomly to the considered user. Note that these curves do not

$p_n = P/N_{\text{ch}} \quad \forall n$
for $n = 1, \ldots, N_{\text{ch}}$
  $m = \arg\max_i \{\gamma_i^{(1)} < p_n T_n\}$
  if $m = \emptyset$ then
    $p_i = \begin{cases} 0, & i = 1, \ldots, n \\ P/(N_{\text{ch}} - n), & i = n+1, \ldots, N_{\text{ch}} \end{cases}$
    continue with next n
  end if
  $b_n = \mathcal{B}(m)$
  $\hat{\imath} = \arg\max_i \{\gamma_{m,i}^{(2)} < p_n T_n\}$
  if $\hat{\imath} = |\mathcal{R}|$ then
    $r_n = \mathcal{R}(\hat{\imath})$
  else
    $r_n = \frac{(\mathcal{R}(\hat{\imath}+1) - \mathcal{R}(\hat{\imath})) p_n T_n + \gamma_{m,\hat{\imath}+1}^{(2)} \mathcal{R}(\hat{\imath}) - \gamma_{m,\hat{\imath}}^{(2)} \mathcal{R}(\hat{\imath}+1)}{\gamma_{m,\hat{\imath}+1}^{(2)} - \gamma_{m,\hat{\imath}}^{(2)}}$
  end if
end for
$R = \max_{r \in \mathcal{R}} \left\{ r \leq \frac{\sum_n b_n r_n}{\sum_n b_n} \right\}$

Fig. 3. Mutual-information based adaptive coding and modulation (MI-ACM) algorithm

TABLE II
SIMULATION PARAMETERS

parameter	value	description
K	288, 1152	input packet length
$n_{\text{f}} \times n_{\text{t}}$	8×15	chunk size (subcarriers $\times$ OFDM symbols)
N_{c}	1664	number of subcarriers
N_{ch}	5, 21	number of assigned chunks

apply to the case where the best chunks are selected by the scheduler, since then the channel seen by the LA algorithm is not a Rayleigh channel.

In Figure 4, the obtained throughput and CWERs for the considered bit-loading algorithms, codeword lengths and number of chunks are illustrated. From these results, various conclusions can be drawn:

- The achieved throughput is at about 4 dB from the reference curve while the CWER fulfils the given target of 0.01.
- For a sufficient number of chunks (Fig. 4(a)), longer codeword sizes perform better, while for a low number of chunks (d), the granularity loss reverses the situation.
- The MI-ACM approach achieves nearly the same throughput as the Hughes-Hartogs algorithm for all considered cases. This indicates that (1) the gain from power loading is insignificant, and (2) there is very little margin for improvement of the MI-ACM algorithm, since the Hughes-Hartogs algorithm provides a strict upper bound. Additionally, the achieved error ratios of the MI-ACM algorithm are lower.
- For the case that the scheduler allocates the N_{ch} best chunks to a given user, the frequency-selective LA does not provide any additional gain since, in this case, the channel gains of these chunks are almost equal. It depends on the channel characteristics and scheduling policy, if this situation is representative for more than one user.

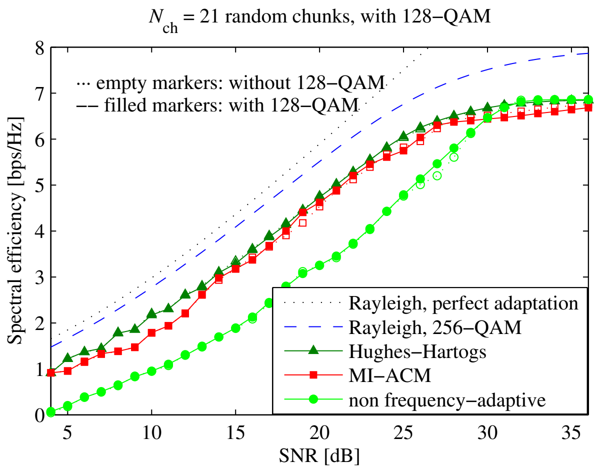

Fig. 5. Achieved spectral efficiencies including non-square constellations.

Finally, Figure 5 depicts the obtained throughput by including 128-QAM in the adapted set of modulations. The throughput with or without non-square constellations is indistinguishable and hence only square constellations are recommended to use.

VI. Conclusion

In this paper we have evaluated a new bit-loading scheme for the WINNER frequency-adaptive transmission, which employs an efficient MI-ACM algorithm with a powerful channel coding in the form of DBTC.

The presented simulation results showed that this adaptation scheme performs close to the optimum solution given by the Hughes-Hartogs bit & power loading algorithm. This is remarkable since the presented algorithm does not apply power loading. The achieved throughput found to be within 4 dB of the capacity of the input-constrained Rayleigh fading channel.

Besides, we have indicated that the use of non-square constellations for bit-loading (8-, 32- and 128-QAM) does not result in higher throughput of the transmission. We can also conclude that the adaptation gain of this scheme depends on the employed scheduling algorithm, which should be taken into account for the evaluation of system level results.

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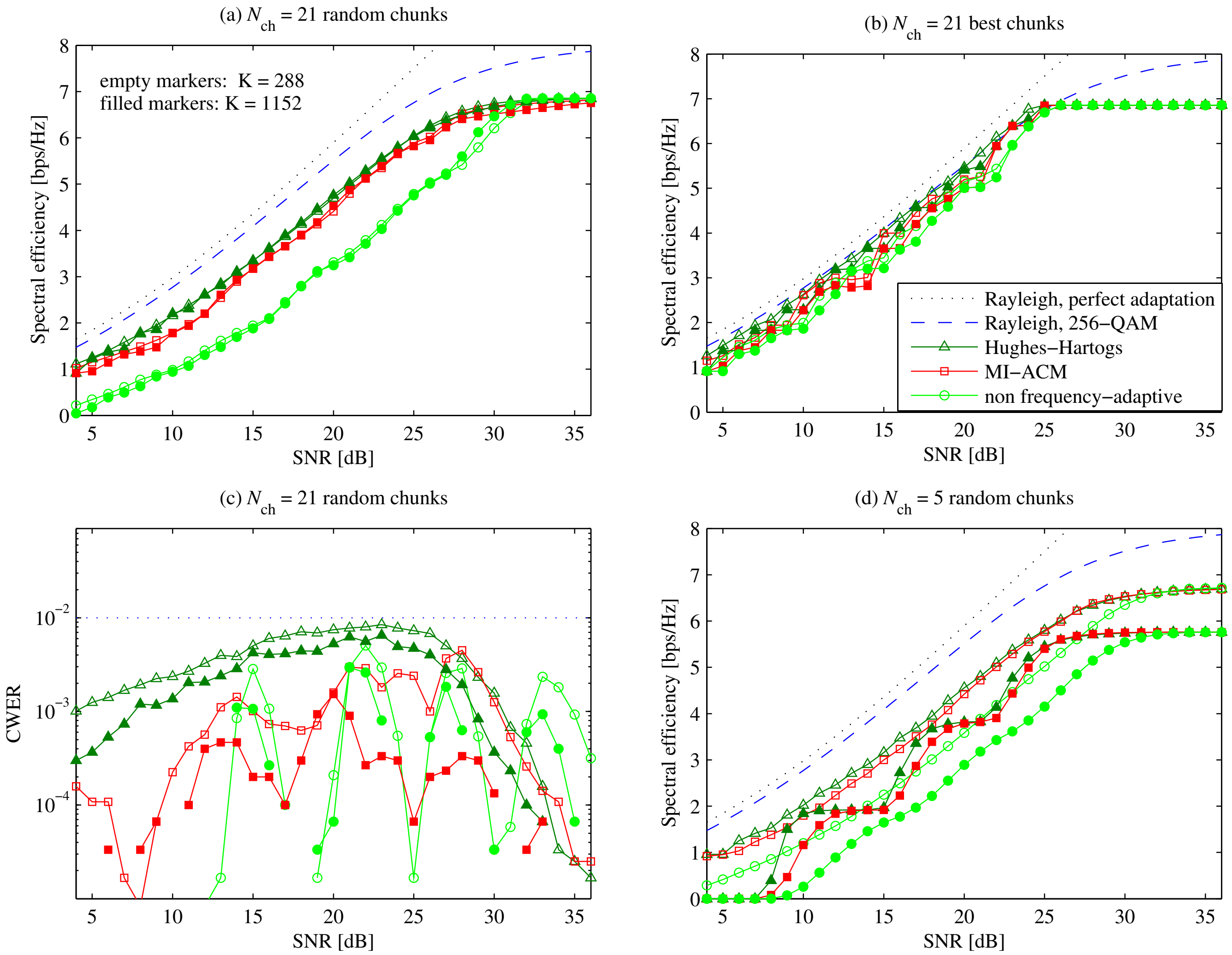

Fig. 4. Achieved spectral efficiencies and codeword error ratios.

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