A Scalable LDPC Coding Scheme for Adaptive HARQ Techniques

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***Abstract*—In wireless communication systems, the achievable data rates are always dependent on the characteristics of the wireless channel. Although a system is designed for an expected channel link quality, the channel may deteriorate for certain periods of time due to the changing nature of the propagation environment. A weak channel may lead to transmission errors, which can be corrected by employing Hybrid Automatic Repeat Requests (HARQ). In particular, type II HARQ protocols have become an appealing option since they only involve the transmission of additional parity bits, instead of repeating the original transmission. These additional parity bits can be obtained by puncturing a low rate channel code. However, finding an optimum pattern of puncturing for a specific code is non-trivial. In this work, we propose a retransmission scheme that considers various Low-Density Parity-Check (LDPC) codes of different rates that can be concatenated to perform the encoding and decoding operations, resulting in a lower encoding overhead for successful transmissions, while maintaining efficient decoding. We show that this technique can maintain high throughput and goodput, and a low outage probability, even when the received signal power is suddenly reduced due to the loss of the Line-of-sight (LOS) component.**

***Index Terms*—Multi-Input Multi-Output (MIMO), Low Density Parity Check Code (LDPC), Hybrid Automatic Repeat Request (HARQ), Orthogonal Frequency Division Multiplexing (OFDM).**

I. INTRODUCTION

The evolution of wireless communications has enabled the development of more sophisticated and specialized communication systems. The deployment of Multi-Input Multi-Output (MIMO) wireless systems [1], [2] has enabled new applications and/or increased quality of service (QoS), namely by taking advantage of the high array and beamforming gains. Even for small form factor devices, the advances in mm-Wave transceivers allow for a greater miniaturization of the devices, while maintaining high data rates.

There is a wide plethora of new services, which range from low-latency aerial video streaming to smart vehicles with real time information sharing [3]. These applications are supported by Machine-to-Machine (M2M) communications [4], which reduce the need for centralized processing nodes and relays. Also, developments regarding the amplification techniques of Orthogonal Frequency Division Multiplexing (OFDM) [5] have enabled the implementation of low-power high data rate transceivers with increased battery lifetimes.

Although these systems can achieve excellent performances, they must still contend with the unreliable nature of the physical wireless channel. In fact, the quality of the wireless link is always a determining factor in the real world performance of the system. In that sense, wireless systems try to compensate such effects by employing diverse strategies such as powerful Forward Error Correction (FEC) schemes, techniques for transmission power adaptation [6], and retransmission mechanisms [7], such as in Hybrid Automatic Repeat Request (HARQ) protocols [8]. In fact, HARQ schemes are very appealing since they do not introduce significant overhead unless a transmission fails due to a sudden drop in the received power. However, each retransmission increases the delay of the communication, which, depending on the application, limits the maximum number of retransmisions allowed.

In this work, we consider a MIMO-OFDM wireless system that operates in a propagation environment with a dominant Line-of-sight (LOS) path between the transmitter and the receiver, and several non-LOS (NLOS) weaker paths, which can be modelled using a Rician distribution. However, we admit the existence of sudden and temporary losses of the LOS, which lead to a degradation of the received signal power that can negatively impact the system's throughput. To combat these temporary channel degradation, we propose the use of a FEC scheme based on a Low Density Parity Check (LDPC) code, combined with an HARQ scheme where the transmitter re-encodes failed transmissions with a larger parity matrix, and transmits only the parity bits of the resulting codeword. It is shown that this technique allows for a significant performance improvement at the receiver, while the decoding operation can be done efficiently by combining all parity matrices at the decoder.

Notation Aspects: In this paper, we utilize bold symbols to denote matrices, and the $(\cdot)^H$ to denote the Hermitian operator.

Paper Organization: The paper is structured as follows: Section II describes the modelled channel environment and characterizes the wireless system; Section III details how the wireless system handles erroneous OFDM symbols and characterizes the proposed solution; Section IV presents the simulated results for the proposed system; Lastly, section V concludes this paper.

II. SYSTEM CHARACTERIZATION

In this work, we consider a point-to-point MIMO system where the receiver is equipped with R antennas and the transmitter has T antennas. The maximum number of independent data streams is therefore $C = \min(R, T)$. We define the channel frequency response matrix at the kth subcarrier, $\mathbf{H}_k$, as

$$\mathbf{H}_k = \begin{bmatrix} H_k^{(1,1)} & H_k^{(1,2)} & \cdots & H_k^{(1,T)} \\ H_k^{(2,1)} & H_k^{(2,2)} & \cdots & H_k^{(2,T)} \\ \vdots & \vdots & \ddots & \vdots \\ H_k^{(R,1)} & H_k^{(R,2)} & \cdots & H_k^{(R,T)} \end{bmatrix}. \quad (1)$$

The data is transmitted using OFDM with N subcarriers, each employing M-ary Quadrature Amplitude Modulation (M-QAM). Each OFDM block is denoted by

$$\mathbf{S}^{(c)} = [S_1, S_2, ...S_N], \tag{2}$$

where $\mathbf{S}^{(c)}$ denotes the OFDM block of the cth stream. We express the symbol vector at the kth subcarrier as $\mathbf{S}_k$.

The separation of the different data streams is performed through a combination of precoding and decoding operations generated by the singular value decomposition (SVD) of the channel matrix [9], which is defined as

$$\mathbf{H}_k = \mathbf{U}_k \mathbf{\Lambda}_k \mathbf{V}_k^H, \tag{3}$$

where $\mathbf{U}_k^H$ is the decoding matrix, $\mathbf{V}_k$ is the precoding matrix, and $\mathbf{\Lambda}_k$ is the diagonal matrix that contains the C singular values, with the cth singular value referred to as $\lambda_{k,c}$.

A. Transmission

Before transmitting the data symbols $\mathbf{S}_k$, the transmitter must perform the precoding operation according to the SVD of the channel. We define the precoded data symbols at the kth subcarrier as

$$\mathbf{X}_k = \mathbf{V}_k \mathbf{S}_k. \tag{4}$$

B. Reception

The received signal is denoted by $\mathbf{Y}_k$ and expressed as

$$\mathbf{Y}_k = \mathbf{H}_k \mathbf{X}_k + \mathbf{N}_k, \tag{5}$$

where $\mathbf{N}_k$ is the frequency-domain version of the additive white Gaussian noise (AWGN) at the kth subcarrier. Before the equalization process, the SVD is completed with the decoding operation. The decoded signal is given by

$$\tilde{\mathbf{W}}_k = \mathbf{U}_k^H \mathbf{Y}_k, \tag{6}$$

The equalization is based on the Zero Forcing (ZF) criterion, wich means that the equalized signal associated of the kth subcarrier of the cth stream is

$$\tilde{S}_{k,c} = \frac{\tilde{W}_{k,c}}{\lambda_{k,c}}. \tag{7}$$

We assume that all singular values exist and are nonzero, which can be ensured with sufficient antenna spacing at the transmitter and receiver. The bth bit Log-Likelihood Ratio (LLR) of the M-QAM symbol of the kth subcarrier associated to the cth branch is computed using the approximate LLR calculation defined as

$$\begin{aligned} L_{k,b} = &\frac{\lambda_{k,c}^2}{\sigma_N^2}(\min(\mathrm{Re}(\mathbf{S}_{k,b}^0)^2 + \mathrm{Im}(\mathbf{S}_{k,b}^0)^2) \\ &- \min(\mathrm{Re}(\mathbf{S}_{k,b}^1)^2 + \mathrm{Im}(\mathbf{S}_{k,b}^1)^2)), \end{aligned} \tag{8}$$

where σ_N^2 is the noise variance, and $\mathbf{S}_{k,b}^0$ and $\mathbf{S}_{k,b}^1$ are complex vectors computed by

$$\mathbf{S}_{k,b}^0 = \tilde{S}_k - S_0, S_0 \in \mathbf{S}_0 \tag{9}$$

and

$$\mathbf{S}_{k,b}^1 = \tilde{S}_k - S_1, S_1 \in \mathbf{S}_1, \tag{10}$$

where $\mathbf{S}_0$ and $\mathbf{S}_1$ are the sets of symbols where the bth bit is 0 and 1, respectively.

C. LDPC Encoding

In order to achieve high data rates in a wireless system, channel coding is essential for the system to function at an acceptable power constraint. The family of Low Density Parity Check codes have become an appealing option for modern systems [10], due to their high performance that approaches capacity.

Since LDPC codes use large code words, the coded bits can be split amongst the different wireless streams, which has the added benefit of allowing each full block to be affected by all independent channels. The sparse parity matrix for this LDPC code is composed of K_p parity check nodes and N_v variable nodes, with a code rate given by $\frac{N_v - K_p}{N_v}$. In order to lower the complexity of the encoding process, we can apply an offline preprocessing step to the parity matrix [11]. The preprocessing consists in using Gaussian elimination to transform the parity matrix into the Approximate Upper Triangular (ALT) form, which reduces the encoding complexity to approximately linear, for small gap values. We refer to the preprocessed parity matrix as $\mathbf{H}_p$.

The codeword to be transmitted is computed as

$$\mathbf{p} = \mathbf{bG}, \tag{11}$$

where $\mathbf{b}$ is the vector containing $N_v - K_p$ information bits and $\mathbf{G}$ is the $(N_v - K_p) \times N_v$ generator matrix obtained from the parity matrix. In each OFDM symbol there will be at least 1 coded block, although for large numbers of subcarriers it is possible to fit more coded blocks in a single symbol.

D. LDPC Decoding

The decoding is done using the Sum-Product Message Passing (MP) algorithm [12], which is an iterative decoding algorithm that computes up to I iterations, or less if the parity check is satisfied. In each iteration the decoder computes the parity check operation for the qth connected check node, which is in turn connected to the dth variable node, defined for the ith iteration as

$$R_j^{(i)}(d) = 2 \text{ atanh}\left(\prod_{d' \neq d} \frac{\tanh(V_{d'}^{(i-1)})}{2}\right), \tag{12}$$

where V_d' is the variable node operation of the d'th node, which must be computed for all variable nodes connected to the qth check node, except for the dth node. This operation is defined as

$$V_d^{(i)}(q) = L_d + \sum_{q' \neq q} R_{q'}^{(i)}, \tag{13}$$

where L_d is the LLR of the dth coded bit, and $R_{j'}^{(i)}$ are the parity checks of the connected check nodes, other than the jth one. For the first iteration we set $V_l^{(0)} = L_d$. The final step in the algorithm is to compute the output LLR values for this iteration, defined as

$$L_d^{(i)} = L_d + \sum_{q \in J_d} R_q^{(i)}(d), \tag{14}$$

where $L_d^{(i)}$ is the LLR of the dth coded bit at the ith iteration, and J_d contains the indices of the check nodes that are connected to the dth variable node. At each iteration the hard decision of the decoded bits is computed, and the parity check is evaluated. If it is satisfied, the decoder does not require more iterations. If the parity check is not correct, then the decoder will continue to the next iteration, up to a maximum of I iterations.

III. Retransmission Schemes

In this work we consider a Rician fading channel that is defined by two terms: a LOS term that does not have fading and a multipath composed by N_{ray} multipath components. We consider that the transmit power P is enough to guarantee the required SNR for the underlying application. However, we admit the existence of blockages of the LOS component that lead to a significant degradation of the received signal power, which decreases to $P(1-p_{LOS})$. We consider that for the first channel realization, there is a probability γ_{LOS} of having the LOS component. We also admit that if there is a blockage and the LOS component is lost, that loss spans over n_{down} channel realizations. After n_{down} realizations, the next n_{up} realizations will have the LOS component. After n_{up} realizations, the next channel realizations will again be dependent on the value of γ_{LOS}. Also, we consider that the transmitter cannot anticipate the change in the channel or its duration, and can only be made aware of the wireless link degradation after transmitting at least one block during the NLOS state.

The average channel power across all subcarriers for the channel between the rth and tth antennas, $\mathbb{E}(|H_k^{(r,t)}|^2)$, is shown in Fig. 1 for 1000 channel realizations, considering $P_{LOS} = 0.75$, $\gamma_{LOS} = 0.05$ and $n_{down} = n_{up} = 20$. Fig. 2 presents the probability density function (PDF) of the absolute value of $H_k^{(r,t)}$ was obtained with 10000 channel realizations, for the same parameters as the previous figure.

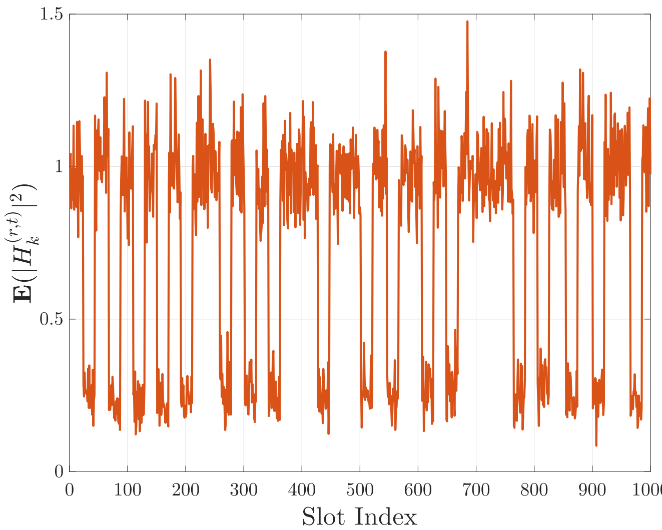

Fig. 1. Average channel power considering 1000 channel realizations with $P_{LOS} = 0.75$, $\gamma_{LOS} = 0.05$ and $N_{down} = n_{up} = 20$.

It should be noted that when a block is transmitted in the NLOS scenario, there is a high probability that the received reconstructed codeword cannot be decoded by the MP algorithm. In this case, the iterative decoder will perform the maximum number of allowed iterations, and if the final codeword does not pass the parity check, the receiver will request retransmission using a Negative Acknowledgement (NAK). The transmitter can then use one of the types of HARQ in order to achieve a successful transmission. We consider the scenario where the retransmissions are performed within the

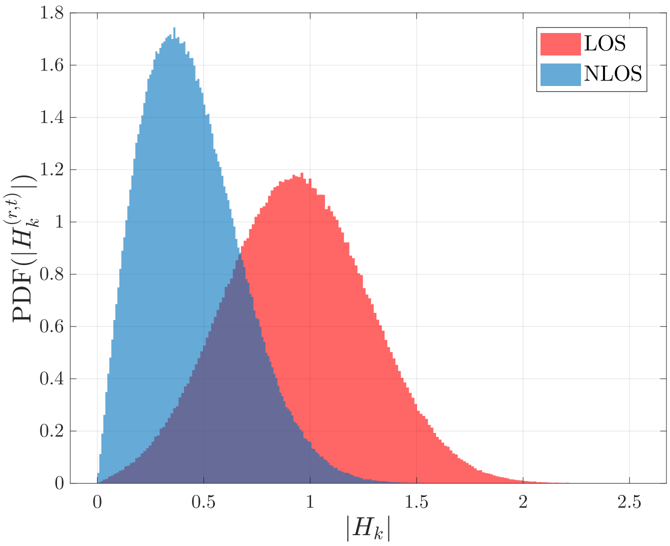

Fig. 2. PDF of $|H_k^{(r,t)}|$ over 10000 channel realizations, with $P_{LOS} = 0.75$, $\gamma_{LOS} = 0.05$ and $n_{down} = n_{up} = 20$.

coherence time interval of the channel, which means that the channel does not recover during the retransmission phase.

The simplest technique is known as Type I HARQ, or HARQ-I, and it requires transmitting the same block once more, which is then received and decoded at the receiver. If the decoding fails again, then a third retransmission is requested, continuing until the block is correctly decoded, or the maximum number of retransmissions is reached. The transmitter can also increase the transmission power in each retransmission, however, this creates a power constraint at the transmitter. Alternatively, the reliability of each retransmission can be greatly increased by combining it with the information contained in the previously transmitted OFDM blocks, known as Packet Combining (PC).

The combination of r received blocks can be performed by adding the LLR of each transmission at each bit. The resulting LLR is defined as

$$L_d^r = \sum_{r'=0}^{r} L_d^{r'}, \tag{15}$$

where L_d^r is the log likelihood of the dth bit, associated with the rth retransmission, and $L_d^0 = L_d$. This technique has the natural advantage of averaging the Gaussian noise in the received signal. We define the equivalent noise associated to the rth combined block as

$$\mathbf{N}_k^r = \frac{1}{r}\sum_{r'=0}^{r} \mathbf{N}_k^{(r')}, \tag{16}$$

where $\mathbf{N}_k^{(r')}$ is the noise term of the r'th retransmission attempt. The noise power of $\mathbf{N}_k^r$ is defined as

$$\sigma_{N,r_{tr}}^2 = \frac{\sigma_N^2}{r_{tr}}. \tag{17}$$

The retransmissions continue until the block is correctly received, or the maximum number of transmissions, r_{tr}, is reached.

A. Conventional Coded Retransmission

Instead of simply re-transmitting the block that was not correctly received, the transmitter can decrease the rate of the channel coding in use by transmitting additional parity bits, known as Type II HARQ, or HARQ-II. This can be done by using a low code rate base coding scheme and then puncturing the parity bits [13], and then sending the punctured bits in the following retransmissions. However, finding an optimum puncturing rule for a specific code is non-trivial. In [14], a solution is proposed using Generalized Root Protograph (GRP) LDPC codes, which uses feedback from the receiver to determine the codewords that were more impaired within the block. These codewords are then punctured from the original block of codewords, and the GRP LDPC matrix is extended by one block. The resulting matrix is then used to encode a new codeword that is sent to the receiver. In [15], a granular retransmission scheme is proposed, where the initial codeword is split into subpackets that can be individually requested by the receiver for retransmission.

B. Proposed Coded Retransmission

In this work, we propose an alternative adaptive rate encoding system where each retransmission contains the parity bits obtained by encoding the previously transmitted codeword with a larger LDPC code, which must satisfy:

$$N_v^{r-1} = N_v^r - K_p^r, \tag{18}$$

where N_v^r and K_p^r represent the number of variable and check nodes of the rth parity matrix, respectively, and $N_v^0 = N_v$. In this case, the code rates for the successive codes can be fine tuned for the desired use-case. The larger parity matrices can be obtained using random generation [16], and can be defined using the same parity check distributions as the first LDPC parity matrix. We refer to the pre-processed parity matrix, [11], associated with the rth coded retransmission as $\mathbf{H}_p^r$, while the rth generator matrix is referred to as $\mathbf{G}^r$. If the maximum number of retransmissions, r_{tr}, is larger than the number of different retransmission LDPC codes, r_p, then the transmitter can loop around and transmit the original coded block, followed by, if necessary, the parity bits of the successive codes.

The codeword for the rth retransmission is defined as

$$\mathbf{p}^r = \begin{cases} \mathbf{p}^{r-1}\mathbf{G}^r, & r \le r_p \\ \mathbf{p}^{r \mod r_p}, & r > r_p \end{cases} \tag{19}$$

where $\mathbf{p}^r$ is the rth codeword that will be transmitted, and p^0 is the codeword of the first transmission. The additional encoding is only performed when a retransmission is requested, which results in no overhead processing for successful transmissions, requiring only additional memory to store the generator matrices. The concatenation of the LDPC codes results in an equivalent code rate that is the product of all individual code rates, which for the rth transmission is given by

$$R_{code}^r = \prod_{r'=0}^{r} \frac{(N_v^{r'} - K_p^{r'})}{N_v^{r'}}. \tag{20}$$

At the receiver, the equalization is performed separately for each re-transmitted block, while the BP decoding can be computed using $r+1$ LDPC decoders, one for the initial code and r for the retransmission codes. In the first step, the decoder for the larger parity matrix decodes the soft decided coded bits, $\mathbf{L}$, and the output soft decoded bits are then used as the input for the decoder associated with the second largest parity matrix, repeating until it decodes the initial code. In order to maximise the performance of this approach, the resulting soft decoded bits would have to be re-encoded using the $r+1$ LDPC codes, and then decoded using the same process. This is to enable the sharing of information between all decoders.

Alternatively, we can construct a parity matrix that represents the relationship between all codes and has a code rate equal to the product of all code rates. The parity matrix can be defined using the following expression

$$\mathbf{H}_p^{de} = \begin{bmatrix} \mathbf{H}_p^{(0,0)} & \dots & \mathbf{H}_p^{(0,K_p)} & \dots & 0 \\ & & \dots & & \\ \mathbf{H}_p^{(N_v,0)} & \dots & \mathbf{H}_p^{(0,K_p)} & \dots & 0 \\ \mathbf{H}_p^{1,(0,0)} & \dots & \mathbf{H}_p^{1,(0,\frac{K_p^1}{2})} & \dots & 0 \\ & & \dots & & \\ \mathbf{H}_p^{r_p,(N_v^{r_p},0)} & & \dots & & \mathbf{H}_p^{r_p,(N_v^{r_p},K_p^{r_p})} \end{bmatrix}, \tag{21}$$

where $\mathbf{H}_p^r$ is the parity matrix for the rth re-encoded transmission, with size $K_p^r \times N_v^r$. This decoding process also allows us to scale the coding scheme further by using increasingly larger LDPC codes without significantly changing the decoding process at the receiver. Since the decoding parity matrix can be constructed by combining the matrices of each individual code, the total required storage capacity at the receiver will be lower than if using a different code for each retransmission.

C. Throughput and Goodput

In order to evaluate the effectiveness of the proposed system we use the throughput, T_{tp}, and goodput, T_{gp}, metrics, both defined in bits per second (bps). The throughput of a system refers to the rate of correctly received bits, being defined in this scenario as

$$T_{tp} = \frac{\log_2(M)\min(T,R)}{t_{tx}}, \tag{22}$$

where t_{tx} is the time necessary for a symbol to be correctly received, which will depend on the additional retransmissions. The goodput of the system refers to the correctly received information bits, and is defined as

$$T_{gp} = \frac{\log_2(M)\min(T,R)R_{code}^0}{t_{tx}}, \tag{23}$$

since it depends on the rate of the LDPC code selected. It should be noted that each retransmission will reduce the value of both of these metrics.

IV. Performance Results

In this section, we present a set of performance results obtained with Monte Carlo simulations of the proposed retransmission scheme. Unless otherwise stated, the average channel gain between two antennas with LOS is unitary, and the number of channel realizations is $N_{slot} = 1000$. The bandwidth in use is 20 MHz, and the number of antennas is $T = R = 4$. The number of data subcarriers is $N = 240$ and the Fast Fourier Transform (FFT) size is 256. The data symbols are selected from a 256-QAM constellation, which means that each transmitted block can contain up to $C_{bits} = 1920$ coded bits, with the amount of information bits being dependent on the chosen code rate. The LDPC coder is set to potentially perform up to $I = 30$ decoding iterations.

We also assume that the first transmission is always encoded by an LDPC with rate $R_{code}^0 = \frac{1}{2}$, $N_v = 960$ and $K_p = 480$, while the re-encoded transmission only scales up once, $r_p = 1$, using a code with $N_v^1 = 1920$ and $K_p^1 = 960$. The maximum number of retransmissions is set at $r_{tr} = 3$.

Let us first analyze the throughput of different retransmission techniques under stable channel conditions, i.e., considering a scenario with $\gamma_{LOS} = 1$. In Fig. 3 we present the evolution of the throughput considering different values of r_{tr}.

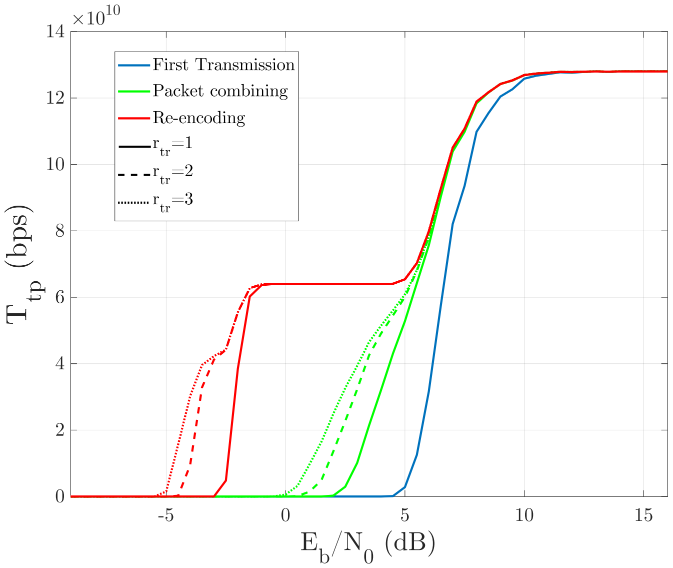

Fig. 3. Throughput of the system as a function of the E_b/N_0, for various numbers of retransmissions.

From the figure, it can be concluded that for a system with a target throughput of $T_{tp} = 12.8$ Gbps, an E_b/N_0 of 10 dB is enough. In this scenario, the use of retransmissions raises the throughput at lower values of E_b/N_0. In fact, it can be observed that the first retransmission with PC allows us to increase the throughput to around $T_{tp} = 5$ Gbps, at 5 dB. However, additional performance gains associated to the execution of more retransmissions (i.e., $r_{tr} > 1$) are marginal, since these retransmissions are performed within the channel coherence time and the transmission power is kept constant. In the same conditions, the proposed re-encoding retransmission scheme can achieve a throughput of $T_{tp} = 6.4$ Gbps with just one retransmission. This is the case even at 0 dB, resulting in a much more energy efficient transmitter.

Fig. 4 shows the Bit Error Rate (BER) results for the initial transmission, as well as the PC and proposed LDPC re-encoding techniques. In this scenario, we considered that the LOS component has $p_{LOS} = 75\%$ of the total channel power, along with a channel state that spans over $n_{down} = n_{up} = 20$ OFDM symbols, and $\gamma_{LOS} = 0.05$. From the figure, it can be noted that employing retransmissions greatly decreases the BER. Once again, it can be observed that the maximum performance gain is obtained in the first retransmission, thanks to the fact that the next successive retransmissions have the same length and are transmitted with the same power as the first transmission, their potential performance gain is lower. Moreover, it can be observed that the proposed re-encoding scheme can achieve a significantly lower BER than the PC scheme, even after $r_{tr} = 3$ retransmissions.

Let us now consider a wireless channel where the LOS component has $p_{LOS} = 90\%$ of the total channel power, and

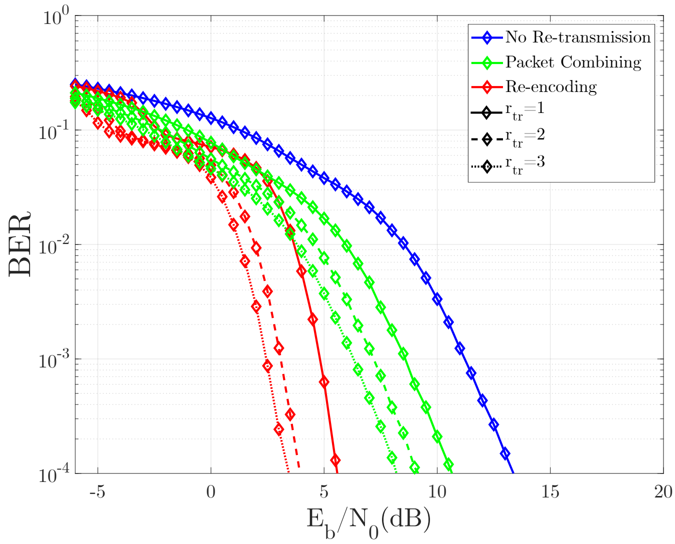

Fig. 4. Simulated BER of the proposed system considering different retransmission strategies.

a channel downtime of $n_{down} = n_{up} = 20$ OFDM symbols. Fig. 5 shows the achievable goodput in that situation.

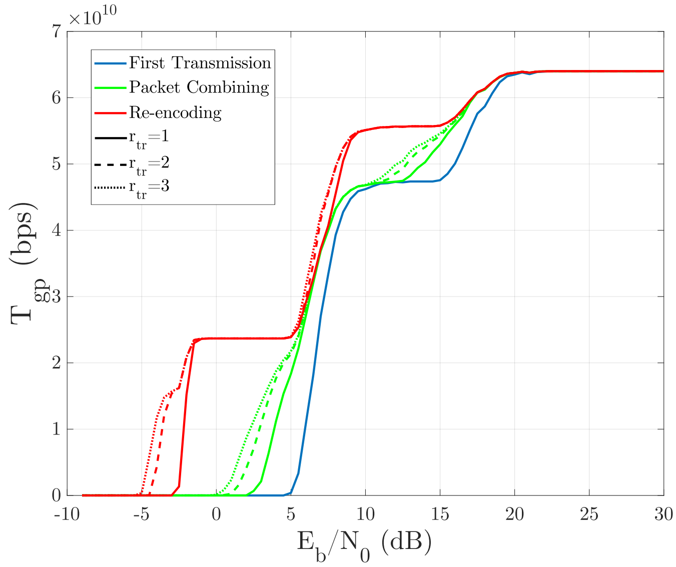

Fig. 5. Goodput of the system as a function of the E_b/N_0, for various numbers of retransmissions, with an LOS component power of 0.9.

From the figure, it is clear that the system with no retransmissions requires a significantly higher E_b/N_0 in order to cope with the channel downtimes. In the figure, two different thresholds can be observed for the initial transmission, the first between approximately 10 dB and 15 dB, and the second beyond 20 dB. The first threshold occurs when the E_b/N_0 is high enough to recover all the transmissions during the LOS channel, while the second threshold occurs when the E_b/N_0 is high enough to also recover the transmissions when there is no LOS. Each PC retransmission decreases the required E_b/N_0 to reach this threshold, while the re-encoded retransmissions increase the goodput at the first threshold, as well as create a new threshold at a lower E_b/N_0, namely between -2 dB and 5 dB, which is due to the greater error correcting capabilities of the second LDPC code.

Fig. 6 shows the simulated outage probability for the proposed system with $r_{tr} = 3$. In this figure three scenarios are shown, namely: (i) a scenario where $\gamma_{LOS} = 1$, (ii) a scenario where $\gamma_{LOS} = 0$ and lastly, (iii) a scenario where the channel randomly loses LOS with $p_{LOS} = 0.9$ and $n_{down} = n_{up} = 20$.

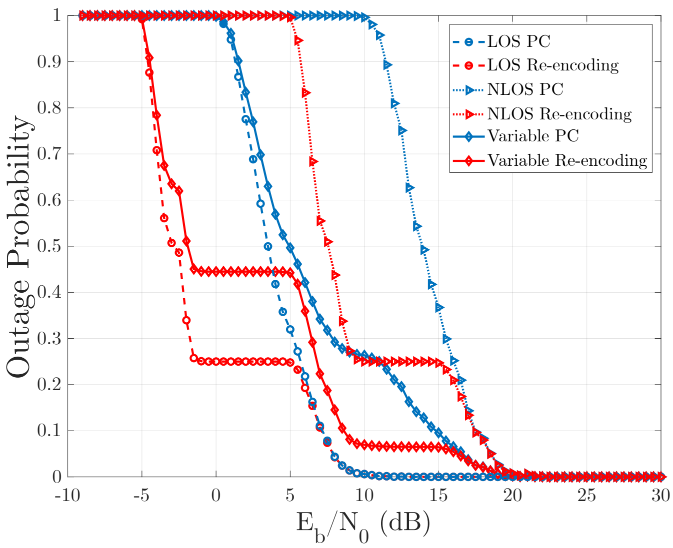

Fig. 6. Outage probability of the proposed scheme considering $r_{tr} = 3$ and a LOS component with $p_{LOS} = 0.9$.

It can be concluded that re-encoding scheme can significantly reduce the outage probability of the system, allowing the system to operate at a much lower E_b/N_0 for this scenario, with low outage probability.

V. Conclusions

In this paper we proposed a scalable LDPC based HARQ strategy for coping with wireless channels that can suffer from sudden and relatively long-lasting deep fades due lost of the LoS path. We showed that the proposed system can achieve excellent performance in both BER and goodput, and can be efficiently scaled as needed for worse propagation scenarios where the power of the received signal can be substantially lower than the one expected in normal conditions.

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