EXPLOITING CYCLIC PREFIX REDUNDANCY IN OFDM TO IMPROVE DECODING OF LDPC CODE

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ABSTRACT

Redundancy, which is added in transmitter in standard OFDM in form of Cyclic Prefix (CP) is usually discarded in the receiver. However, it can be interpreted as a repetition code, which repeats only part of the symbols. Therefore it is a rather weak error control code. Nevertheless the receiver could be modified to implement a signal processing technique using the CP redundancy, which allows increasing of error correcting capability of an outer LDPC code. The details on such a modification and corresponding simulations results are presented in this paper.

1. INTRODUCTION

All practical Coded OFDM (COFDM) systems contain redundancy not only in the error control code (e. g. Turbo-code or LDPC), but also in the form of a cyclic prefix (CP), added by the OFDM transmitter to eliminate the inter-block interference (IBI) caused by multipath propagation. As shown in later sections this redundancy turns the linear channel convolution, represented by the channel impulse response ***h***, into cyclic convolution, thus allowing a computationally feasible frequency-domain equalization (FDE) also described in [1]. The added redundancy is usually discarded in the receiver.

As described in [2] and [3], the CP insertion can be understood as a short repetition-like code over complex numbers in time domain with only part of the symbols repeated. The high coderate and simplicity give such code a very weak error correcting capability. However, if some conditions are met, the inverse Fourier transform in a COFDM transmitter can approximate the operation of an interleaver and such transmitter can be understood as a system with serial concatenation of codes, as defined in [4]. The two concatenated codes – outer ECC defined by a communication standard and inner repetition can be then decoded iteratively according to the well known turbo-decoding principle defined in [5]. Even a weak inner repetition code can improve the decoded error ratio of the present powerful outer error correcting code. In theory such a modification is easy to implement in an SDR system. In reality the situation is more complicated. Because the decoding of a powerful outer code, such as Turbo-code, is an iterative process, the computational complexity of resulting super-iterative decoder in the receiver can be prohibitively high. Other problems arise from the fact that the inner code operates on complex numbers in the time domain, while the outer code is a binary code (GF(2^p) based) and operates in frequency domain – after DFT and frequency domain equalization. In turbo decoding, the partial decoders must share extrinsic information in the form of Log-Likelihood Ratio (LLR), which is defined only for binary random variables. For the inner complex-field code (CFC), there is no compatible LLR definition. This problem is addressed in [3], where it is suggested, that the decoding of the inner code must be transformed to an equivalent decoding operation in frequency domain that operates over binary codes and can use LLR values.

Even with the problem of cooperation of decoders solved, other problems remain. In a multipath channel, the inter-block interference always affects OFDM transmission. The redundancy in CP allows the elimination of this interference. This is implemented usually by dropping the affected prefix samples in the receiver. On the other hand, if the information from the CP is to be used in the decoding, samples of the prefix cannot be dropped. In [2] a new method for eliminating IBI from OFDM transmission, by applying additive and subtractive corrections in the receiver under the assumption of perfect channel state information (CSI) knowledge in the receiver is presented along with a suboptimal method for improving the decoding of Turbo-codes by exploiting the redundancy in the OFDM cyclic prefix. The presented method proposed addition of a transmitted block reconstruction in the receiver. This reconstruction block involves also a turbo-code encoding process that uses the already decoded data bits to estimate the transmitted OFDM block and remove interference. Unfortunately, errors in the decoded data stream multiply in the subsequent turbo-encoding operation and the resulting system outperforms the original system by a mere 0.1 dB while the computational complexity of the modified receiver is more than double.

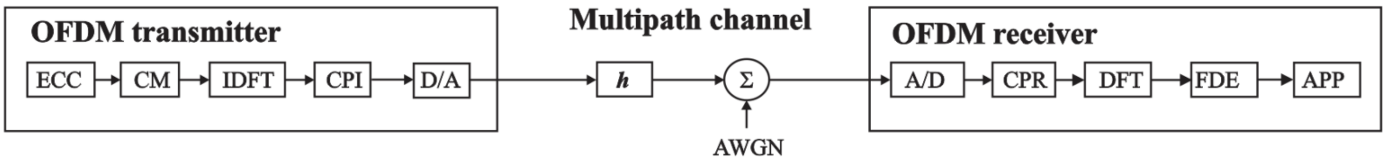

Fig. 1: OFDM transmission in a multipath channel. The ECC block represents a powerful Error Correcting Code specified in a communication standard, Constellation Mapping (CM) block maps bits to complex numbers, Inverse Discrete Fourier Transform block maps vectors of complex samples in frequency domain to time domain. Cyclic Prefix Insertion (CPI) block introduces redundancy necessary to eliminate IBI, which is then removed in the Cyclic Prefix Removal (CPR) block in the receiver. FDE block implements a simple Frequency Domain Equalization. The ECC code is then decoded by utilizing an A-Posteriori Probability (APP) decoding algorithm.

This paper examines the effect of the prefix samples extraction method presented in [2] when a more suitable, LDPC ECC along with a simple Min-Sum (MS) decoding algorithm is used. The simulation results confirm the theoretical predictions, that such a decoder outperforms Turbo-code based system in terms of both bit error ratio and computational complexity. In the next section a proper, more formal definition of the prefix information samples extraction, compared to somehow intuitive definition in [2] is presented. The third section then describes several new alternatives of receiver design modification using Low-Density-Parity Check code as an ECC. The simulation results in the last section evaluate the error performance of various configurations of the modified system.

2. MATRIX MODELS

In this section, formal matrix models for the most important system blocks are presented. Some of them are well know, other are original. The motivation for matrix modeling is the desire to express the whole system operation as a simple matrix-vector multiplication.

The first well known model is the discrete Fourier transform matrix $\boldsymbol{W}$ defined in [6]:

$$\boldsymbol{W} = \frac{1}{\sqrt{N}} \begin{bmatrix} 1 & 1 & 1 & \dots & 1 \\ 1 & \omega & \omega^2 & & \omega^{N-1} \\ 1 & \omega^2 & \omega^4 & & \omega^{2(N-1)} \\ \dots & & & & \\ 1 & \omega^{N-1} & \omega^{2(N-1)} & \dots & \omega^{(N-1)(N-1)} \end{bmatrix}$$

where ω depends on the transform size:

$$\omega = e^{-\frac{2\pi j}{N}}$$

Since $\boldsymbol{W}$ is a unitary matrix, the IDFT matrix is easily obtained by means of Hermitian transpose:

$$\boldsymbol{W}^{-1} = \boldsymbol{W}^{\mathcal{H}}$$

The next well known matrix models are the models of channel convolution. The linear channel convolution of the transmitted block $\boldsymbol{t}$ with channel's impulse response $\boldsymbol{h}$ can be expressed as multiplication by a convolution matrix $\boldsymbol{H_L}$:

$$\boldsymbol{H_L}^{((B+\nu-1)\times B)} = \begin{bmatrix} h_1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ h_2 & h_1 & 0 & 0 & 0 & 0 & 0 & 0 \\ \vdots & h_2 & \ddots & 0 & 0 & 0 & 0 & 0 \\ h_\nu & \vdots & & \ddots & 0 & 0 & 0 & 0 \\ 0 & h_\nu & & & & 0 & 0 & 0 \\ 0 & 0 & \ddots & & & \ddots & 0 & 0 \\ 0 & 0 & 0 & \ddots & & & h_1 & 0 \\ 0 & 0 & 0 & 0 & & & h_2 & h_1 \\ 0 & 0 & 0 & 0 & 0 & \ddots & \vdots & h_2 \\ 0 & 0 & 0 & 0 & 0 & 0 & h_\nu & \vdots \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & h_\nu \end{bmatrix}$$

The size of $\boldsymbol{H_L}$ depends on the input vector size B and size of the impulse response ν. The linear convolution prolongs the transmitted block by ν - 1 samples. Circular convolution on

$$\boldsymbol{H_c}^{(N\times N)} = \begin{bmatrix} h_1 & 0 & 0 & 0 & 0 & h_\nu & \dots & h_2 \\ h_2 & h_1 & 0 & 0 & 0 & 0 & h_\nu & \dots \\ \dots & h_2 & h_1 & 0 & 0 & 0 & 0 & h_\nu \\ h_\nu & \dots & h_2 & h_1 & 0 & 0 & 0 & 0 \\ 0 & h_\nu & \dots & h_2 & h_1 & 0 & 0 & 0 \\ 0 & 0 & h_\nu & \dots & h_2 & h_1 & 0 & 0 \\ 0 & 0 & 0 & h_\nu & \dots & h_2 & h_1 & 0 \\ 0 & 0 & 0 & 0 & h_\nu & \dots & h_2 & h_1 \end{bmatrix}$$

the other hand preserves the transformed vector size [7]: $\boldsymbol{H_c}$ is a square circulant matrix , that can be diagonalized by multiplication with the Fourier transform matrices [6,7]:

$$\boldsymbol{D_h} = \boldsymbol{W} \times \boldsymbol{H_c} \times \boldsymbol{W}^{-1}$$

Where $\boldsymbol{D_h}$ is a diagonal matrix with nonzero entries equal to samples of an N-point DFT of the channel impulse response:

$$diag(\boldsymbol{D_h}) = [H(0), H(e^{-2\pi j/N}) \dots H(e^{-2\pi j(N-1)/N})]$$

The key point in cyclic prefix insertion and removal is the transformation of actual channel linear convolution to cyclic convolution. This can be easily shown if proper formal models of prefix operations are at hand. Despite their simplicity, the following models are usually missing in literature:

$$\boldsymbol{\Omega}^{((N+O)\times N)} = \begin{bmatrix} \boldsymbol{0}^{O\times N-O} & \boldsymbol{I}_O \\ & \boldsymbol{I}_N \end{bmatrix}$$

$\boldsymbol{\Omega}$ is the cyclic prefix insertion matrix. The multiplication of a vector of size N by $\boldsymbol{\Omega}$ implements the operation of inserting a cyclic prefix of size O, resulting in a longer vector containing $N + O$ samples. $\boldsymbol{\Omega}$ is very simple – it consists of identity and zero matrices of proper sizes. Similarly, the cyclic prefix removal matrix $\boldsymbol{\Psi}$ implements corresponding operation in the receiver:

$$\boldsymbol{\Psi}^{(N\times(N+O+\nu-1))} = \begin{bmatrix} \boldsymbol{0}^{N\times O} & \boldsymbol{I}_N & \boldsymbol{0}^{N\times\nu-1} \end{bmatrix}$$

Under the condition that $\nu \leq O+1$ the conversion of $\boldsymbol{H_L}$ to $\boldsymbol{H_c}$ is trivial:

$$\boldsymbol{H_c}^{(N\times N)} = \boldsymbol{\Psi}\times\boldsymbol{H_L}^{((N+O+\nu-1)\times(N+O))}\times\boldsymbol{\Omega}$$

After the channel convolution matrix can be considered circulant, a simple per-component frequency equalization takes place in most modern OFDM receivers:

$$\boldsymbol{S}' = \boldsymbol{D_h}^{-1}\times\boldsymbol{Z} = \boldsymbol{S} + \boldsymbol{D_h}^{-1}\times\boldsymbol{N}$$

Where $\boldsymbol{S'}$ is an estimate of the transmitted frequency domain samples $\boldsymbol{S}$ and $\boldsymbol{N}$ is a Fourier transform of channel noise vector (and $\boldsymbol{Z}$ are unequalized frequency domain samples).
The method for extracting the prefix redundancy presented in [2] is based on identification of a second circulant submatrix in the linear channel convolution matrix as demonstrated in Fig.2:

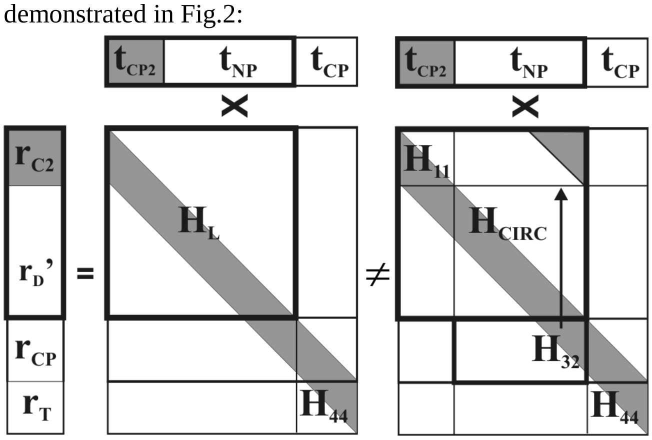

Fig. 2: Second circulant submatrix in linear channel convolution matrix after a correction is applied in receiver.

However this graphical intuitive description deserves a more formal equivalent. It is presented in the following section.

2.1. Matrix segmentation

Let $\boldsymbol{A} = (a_{ij})$ be a matrix of complex numbers: $a_{ij} \in C$, of size $R \times S$:

$$\boldsymbol{A} = \begin{bmatrix} a_{11} & a_{12} & \dots & a_{1S} \\ a_{21} & & & a_{2S} \\ \vdots & & & \\ a_{R1} & a_{R2} & \dots & a_{RS} \end{bmatrix}$$

It is possible to define two sets M_b and N_b of positive numbers:
$M_b = \{ r_1, r_2, \dots , r_m\}$, where $\forall\, r_i > 0$, $\sum r_i = R$ and $|M_b| = m$
$N_b = \{ s_1, s_2, \dots , s_n\}$, where $\forall\, s_i > 0$, $\sum s_j = S$ and $|N_b| = n$

M_b is the set of row groups' dimensions and N_b denotes the set of column group's dimensions. m is the number of the row groups and n is the number of column groups. These groups of rows and columns partition the matrix $\boldsymbol{A}$ into a grid of submatrices:

$$\boldsymbol{A} = \begin{bmatrix} \boldsymbol{A}_{11} & \boldsymbol{A}_{12} & \dots & \boldsymbol{A}_{1n} \\ \boldsymbol{A}_{21} & & & \boldsymbol{A}_{2n} \\ \vdots & & & \\ \boldsymbol{A}_{m1} & \boldsymbol{A}_{m2} & \dots & \boldsymbol{A}_{mn} \end{bmatrix}$$

Where $\boldsymbol{A}_{ij}$ is the submatrix of $\boldsymbol{A}$ with r_i rows and s_j columns. This partitioning will be called *segmentation of matrix* $\boldsymbol{A}$. Using this partitioning, a usual matrix-vector multiplication operation:

$$\boldsymbol{y} = \boldsymbol{A}\times\boldsymbol{x}$$

can be expressed in terms of subvectors of $\boldsymbol{x}$ and submatrices of $\boldsymbol{A}$ defined by the partitioning above:

$$\boldsymbol{y}_i = \sum_{j=1}^{n} \boldsymbol{A}_{ij}\times\boldsymbol{x}_j \quad i = 1, 2 \dots m\,.$$

where vector $\boldsymbol{x}$ is partitioned into subvectors in a similar manner using the set of subvector sizes N_b.

2.2. Segmentation of channel convolution matrix

The segmentation of the linear channel convolution matrix is defined by the segmentation of the transmitted and received vector $\boldsymbol{t}$ and $\boldsymbol{r}$. The definition of the sets M_b and N_b is given by the analysis of the operation of an OFDM system in multipath channel conditions:

$N_b = \{O, N - O, O\}$
$M_b = \{O, N - O, O, \nu - 1\}$

This definition of subvector sizes defines the partitioning of transmitted vector $\boldsymbol{t}$ ($\|$ denotes vector concatenation):

$$\boldsymbol{t}^{((N+O)\times 1)} = \left[\boldsymbol{t}_{CP2}{}^{(O\times 1)} \,\|\, \boldsymbol{t}_{NP}{}^{((N-O)\times 1)} \,\|\, \boldsymbol{t}_{CP}{}^{(O\times 1)} \right]$$

It is necessary to define two subvectors of the transmitter output:

$$\boldsymbol{s}^{(N\times 1)} = \left[\boldsymbol{t}_{NP} \,\|\, \boldsymbol{t}_{CP} \right]$$

$$\boldsymbol{s}_2{}^{(N\times 1)} = \left[\boldsymbol{t}_{CP2} \,\|\, \boldsymbol{t}_{NP} \right]$$

Vector $\boldsymbol{s}$ is the original data vector, while vector $\boldsymbol{s}_2$ is the rotated version of $\boldsymbol{s}$ containing the redundant cyclic prefix samples. Similarly M_b defines the partitioning of the received vector $\boldsymbol{r}$:

$$\boldsymbol{r} = \left[\boldsymbol{r}_{CP}{}^{(O\times 1)} \,\|\, \boldsymbol{r}_D{}'^{((N-O)\times 1)} \,\|\, \boldsymbol{r}_D{}''^{(O\times 1)} \,\|\, \boldsymbol{r}_T{}^{((\nu-1)\times 1)} \right]$$

Again, two subvectors are particularly interesting for our purposes :

$$\boldsymbol{r}_D{}^{(N\times 1)} = \left[\boldsymbol{r}_D{}' \,\|\, \boldsymbol{r}_D{}'' \right]$$

$$\boldsymbol{r}_{D2}{}^{(N\times 1)} = \left[\boldsymbol{r}_{CP} \,\|\, \boldsymbol{r}_D{}' \right]$$

$\boldsymbol{r}_D$ is the received subvector processed in current ODFM systems and $\boldsymbol{r}_{D2}$ is the second received subvector, processed in the modified receiver in the added processing branch.

Vector $\boldsymbol{r}_T$ denotes the prolonging of the transmitted block and since it overlaps with $\boldsymbol{r}_{CP}$ from the next transmitted OFDM block, it is discarded in current systems. This inter-block interference can be described by introducing block index n :

$$\boldsymbol{r}_{IBI(n)} = \boldsymbol{r}_{CP(n)} + \boldsymbol{r}_{T(n-1)}$$

If the samples of $\boldsymbol{r}_{CP}$ are to be processed further, a simple subtractive correction has to be applied to $\boldsymbol{r}_{IBI}$ to eliminate IBI:

$$\boldsymbol{r}_{cor1(n)} = \boldsymbol{r}_{T(n-1)} = \boldsymbol{H}_{43(n-1)} \times \boldsymbol{t}_{CP(n-1)}$$

This correction depends on presence of the proposed transmitted block reconstruction block in the receiver.

Vectors $\boldsymbol{r}_D$ and $\boldsymbol{r}_{D2}$ are not completely independent, they contain a common subset of samples $\boldsymbol{r}_D$**'**. The information from these samples is taken into account in two branches of the modified receiver proposed in further sections and therefore affects the resulting decoding twice. This is an obvious drawback that renders this design suboptimal.

Finally, the sets M_b and N_b define the segmentation of the channel convolution matrix:

$$\boldsymbol{H}_L = \begin{bmatrix} \boldsymbol{H}_{11}{}^{(O\times O)} & \boldsymbol{H}_{12}{}^{(O\times(N-O))} & \boldsymbol{H}_{13}{}^{(O\times O)} \\ \boldsymbol{H}_{21}{}^{((N-O)\times O)} & \boldsymbol{H}_{22}{}^{((N-O)\times(N-O))} & \boldsymbol{H}_{23}{}^{((N-O)\times O)} \\ \boldsymbol{H}_{31}{}^{(O\times O)} & \boldsymbol{H}_{32}{}^{(O\times(N-O))} & \boldsymbol{H}_{33}{}^{(O\times O)} \\ \boldsymbol{H}_{41}{}^{((\nu-1)\times O)} & \boldsymbol{H}_{42}{}^{((\nu-1)\times(N-O))} & \boldsymbol{H}_{43}{}^{((\nu-1)\times O)} \end{bmatrix}$$

It is obvious that received vector $\boldsymbol{r}_D$ is related to transmitted vector $\boldsymbol{s}$ by a multiplication with circulant matrix $\boldsymbol{H}_c$:

$$\boldsymbol{H}_C{}^{(N\times N)} = \begin{bmatrix} \boldsymbol{H}_{22} & \boldsymbol{H}_{21} + \boldsymbol{H}_{23} \\ \boldsymbol{H}_{32} & \boldsymbol{H}_{31} + \boldsymbol{H}_{33} \end{bmatrix}$$

Similarly the second received vector $\boldsymbol{r}_{D2}$ can be expressed as a result of multiplying $\boldsymbol{s}_2$ with matrix $\boldsymbol{H}_L$**'**, which is not circulant :

$$\boldsymbol{H}_L{}' = \begin{bmatrix} \boldsymbol{H}_{11} & \boldsymbol{H}_{12} \\ \boldsymbol{H}_{21} & \boldsymbol{H}_{22} \end{bmatrix}$$

Therefore the desired circular dependence between $\boldsymbol{r}_{D2}$ and $\boldsymbol{s}_2$ must be created artificially by introduction of an additive correction :

$$\boldsymbol{r}_{cor2} = \boldsymbol{H}_{32} \times \boldsymbol{t}_{NP}$$

This correction must be applied to the proper subvector of the received block :

If we redefine $\boldsymbol{r}_{D2}$ to use this corrected values, vectors $\boldsymbol{r}_{D2}$

$$\boldsymbol{r}_{CP}{}' = \boldsymbol{r}_{CP} + \boldsymbol{r}_{Cor2} = \boldsymbol{H}_{11} \times \boldsymbol{t}_{CP2} + \boldsymbol{H}_{32} \times \boldsymbol{t}_{NP}$$

and $\boldsymbol{s}_2$ become bound by a circulant matrix $\boldsymbol{H}_c$, same as in standard processing branch :

$$\boldsymbol{H}_c = \begin{bmatrix} \boldsymbol{H}_{11} & \boldsymbol{H}_{12} + \boldsymbol{H}_{32} \\ \boldsymbol{H}_{21} & \boldsymbol{H}_{22} \end{bmatrix}$$

Fig. 3 summarizes all the definitions in one example of linear channel convolution matrix segmentation:

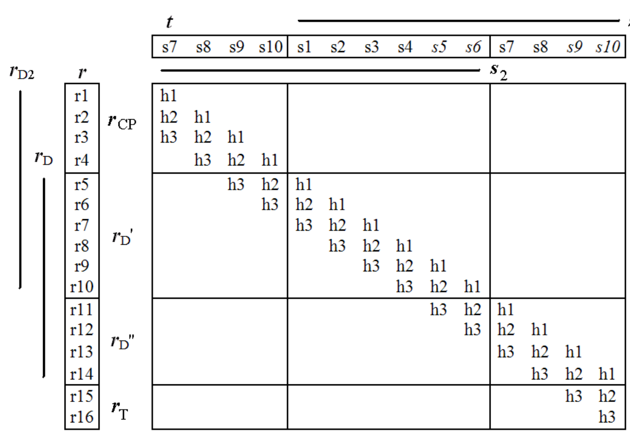

Fig. 3: Example channel matrix segmentation along with the segmentation of transmitted and received time domain vectors for N=10, O=4 and $\nu = 3$.

Because the redundant samples in subblock $\boldsymbol{s}_2$ of the transmitted block undergo exactly the same transformation as payload samples, the received subvector that originates from them can be equalized in exactly the same way using a simple per-component frequency domain equalization.

Furthermore, the equalization of these redundant samples uses exactly the same channel frequency estimate values. Since vector $\boldsymbol{r}_{D2}$ is a rotated version of $\boldsymbol{r}_D$, a constant phase correction must be applied to its equalized form in frequency domain:

$$S_2{}''_{(k)} = S_2{}'''_{(k)} \times e^{j2\pi \frac{O}{N}(k-1)} \quad ; \; k = 1, \ldots, N$$

3. PROPOSED RECEIVER MODIFICATION

The previous section formally described the process of extraction of two transmitted vector estimates $\boldsymbol{S}_2$**'** and $\boldsymbol{S}_2$**''** from one received OFDM samples block. Although the additive noise was omitted from the formulas for the sake of brevity, these vectors are vectors of complex samples in the frequency domain affected by complex AWGN noise. As shown in Fig. 4, in current OFDM systems vector $\boldsymbol{S}_2$**'** enters the LLR computation block followed by a soft-input soft-output ECC decoder. After several decoder iterations, a hard decision (HD) is made based on the posterior LLR estimates of received bits and the decoded data is fed to upper layers. The purpose of the modification shown in Fig. 4 is to improve the BER of the decoded data stream by utilizing the redundancy in cyclic prefix samples while preserving compatibility with present communication standard – without imposing any additional requirements for the transmitter.

Unlike the modification described in [2], the new system utilizes the decoded data in a very effective way: the decoded posterior LLR estimates from decoder #1 are stored in memory (M) while the transmitter output is reconstructed and a second set of LLR is computed in the added processing branch. The Min-Sum (MS) decoding algorithm, used for decoding of LDPC codes delivers posterior LLR values for all the bits in the codeword, not just for the data bits as is common in turbo-codes decoding. This greatly improves the overall decoding process, since no encoding operation is necessary in the transmitter output estimation block, thus no extra errors are introduced there.

The second processing branch consists of proper subblock selection (PBS – Prefix Block Selection) that selects the $\boldsymbol{r}_{D2}$ subvector of the received block. In Additive Correction (AC) block both $\boldsymbol{r}_{cor1}$ and $\boldsymbol{r}_{cor2}$ are applied. The DFT and FDE blocks are exactly the same as in the standard processing branch while the Spectral Shift (SS) block implements the multiplicative phase correction. The resulting samples $\boldsymbol{S}_2$**''** then enter a second LDPC decoder (#2) that performs the same number of iterations *nIter* as decoder #1. The posterior LLR estimates from both branches then enter the repetition code decoder. Since soft decoding of a repetition code is just a summation of LLR values [2,3] this block brings almost no additional complexity. The resulting improved LLRs then enter a third LDPC decoder, where several extra iterations are performed. The resulting data stream should contain fewer errors than in a standard receiver implementation.

5. SIMULATION RESULTS

The simulation parameters were set to be as close to real world values as possible. The OFDM system parameters values were taken from the WiMax standard IEEE 802.16-2009 [9]. The used LDPC code defined in this standard has a coderate R = 1/2 with codeword size n = 1152 bits. The OFDM symbol consisted of N = 1152 BPSK mapped samples. The size of the cyclic prefix was set to one eighth of the useful data samples.

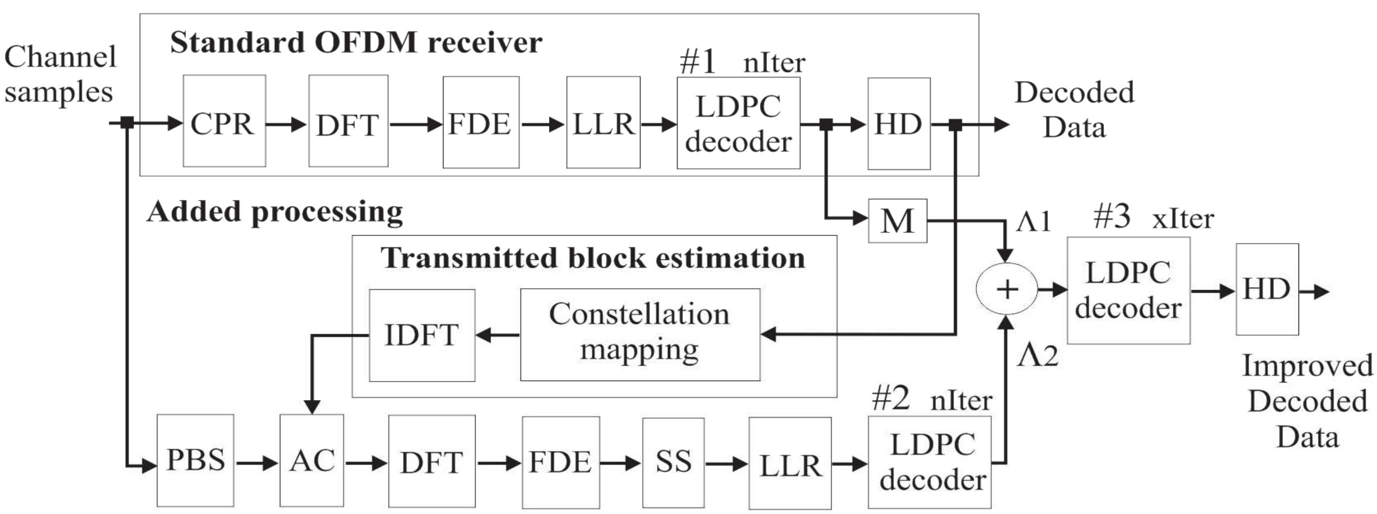

Fig. 4: Modification of an OFDM system. Added transmitter output estimator serves as a basis for computing corrections necessary for prefix redundancy extraction. The decoded LLRs from both processing branches are then combined.

The channel was modeled as a multipath quasi-static Rayleigh fading channel with Doppler shift parameter set to zero. The number of the multipath components was set to 18 with path gains distributed according to [10].

Various configuration parameters of the modified receiver from Fig. 4 were tested and compared to a standard system implementation that used 8 decoder iterations. The first configuration of the modified design used *nIter* = 8 in both processing branches and zero extra iterations *xIter* = 0 of the third LDPC decoder, therefore it is denoted [8 + 0]. The second configuration used *nIter* = 7 and *xIter* = 1 and is denoted [7 + 1]. The last configuration was [5 + 3]

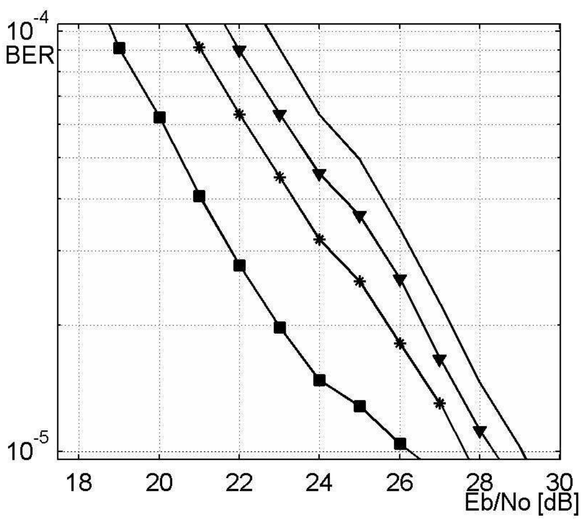

Fig. 5: Various receiver modifications error performance. Standard system with 8 iterations (no marker). [8 + 0] setup (triangular marker), [7 + 1] setup (star) and [5 + 3] (square).

As shown in Fig. 5, the last configuration's ([5 + 3]) performance is best among all simulated configurations. Especially in the range of error ratios between BER = 10^{-4} and BER = 10^{-5} it outperforms the standard receiver by about 4dB. The price for this improvement is the approximate 65 % increase in receiver computational complexity (total 13 LDPC decoder iteration compared to original 8).

6. CONCLUSION

This paper has shown how it is possible to improve the decoding of LDPC codes used in modern communication standard such as IEEE 802.16-2009 in systems using OFDM modulation with cyclic prefix. The presented modification doesn't impose any restriction to the communication standard and it reuses as many existing receiver functional blocks as possible, making it ideal for an SDR implementation, eventually enabled and disabled adaptively on the fly. Simulation results confirm that depending on the desired BER, the proposed suboptimal modification improves decoding in multipath channel with AWGN by 1 to 4dB if a prefix redundancy of one eighth of the useful payload samples is used and perfect channel state information is present in the receiver. The modified receiver computational complexity is approximately 65 % higher than the complexity of the original system.

8. ACKNOWLEDGMENT

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