Iterative Demodulation for DVB-S2

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Abstract— **We propose the application of iterative demodulation at a DVB-S2 receiver. The demodulation is improved by feedback of extrinsic information from the decoder of the LDPC channel code. The presented simulation results show an enhancement of about 0.3 dB, which is notable taking the gap to the Shannon limit into account. The additionally required computational demand is negligible and the receiver only needs to be slightly modified. With the unmodified transmitter the proposed scheme is completely standard compliant.**

I. Introduction

The second-generation specification for satellite broadcasting of the Digital Video Broadcasting Project (DVB-S2) [1] adopted low-density parity check (LDPC) codes as its main channel coding technique [2],[3]. LDPC codes were originally developed by Gallager in the 1960s [4] and recently rediscovered [5], as today the necessary iterative decoding becomes practicable in terms of complexity. The LDPC codes for DVB-S2 exhibit a performance up to only 0.6 dB from the Shannon limit [3]. Furthermore, to accommodate the demand for high data rates higher order modulations such as 16APSK (amplitude phase shift keying) are specified for DVB-S2 [1].

The Turbo principle, which has been originally applied to concatenated channels codes [6],[7], can be extended to iteratively improve the demodulation at the receiver by using feedback of *extrinsic* information from the adjacent channel decoder. This technique for *iterative demodulation* is known as bit-interleaved coded modulation with iterative decoding (BICM-ID) [8],[9],[10]. Usually convolutional codes are studied in the context of *iterative demodulation*, but the only constraint for the channel code consists of the channel decoder being able to supply *extrinsic* information to the demodulator. Thus, also more complex channel codes, which are iterative codes themselves such as the Turbo codes of UMTS, IEEE 802.16, and DVB-RCS (return channel via satellite) or the LDPC codes of IEEE 802.16 and DVB-S2, can be used to enhance the system by *iterative demodulation*. However, since *iterative demodulation* improves the error floor performance usually at the cost of the waterfall region occurring at a higher E_b/N_0, it may be hard to apply *iterative demodulation* successfully to systems with very strong channel codes implying a very low error floor or none at all [11].

In this contribution we apply *iterative demodulation* to the DVB-S2 system. In Section II we present the resulting system model and show that only the receiver requires minor modifications. The transmitter is not modified at all. Thus, the system is still compliant with the standard. Section III deals with the signal constellation sets (SCSs) and mappings. Additionally different strategies for the optimization of the system are discussed based on an EXIT chart analysis [12]. However, it turns out that the mapping specified in the standard exhibits the best performance. The simulation results given in Section IV show an improvement of $\Delta_{E_b/N_0} \approx 0.3$ dB. Although this gain is not very large in its absolute magnitude, its relative size in relation to the remaining distance of $\approx 0.6 \ldots 1.2$ dB [3] to the Shannon limit is quite noteworthy. Furthermore, the increase in computational complexity due to *iterative demodulation* is negligible.

II. The System Model

Fig. 1 depicts the baseband model of the considered system, which resembles a BICM-ID system [8] with an LDPC code serving as channel code [11]. The outer BCH code being part of the FEC (forward error correcting) coding of DVB-S2 is omitted. The BCH code is only used to avoid error floors [2], while the error correcting performance itself is dominated by the almost vertical waterfall behavior (in terms of BER vs. E_b/N_0) of the LDPC code (see e.g. [3]).

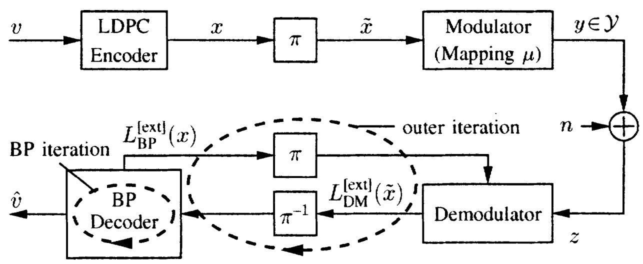

Fig. 1. Baseband model for iterative demodulation.

A frame of M random data bits v is encoded by an (N, M)-LDPC code with the generator matrix $\mathbf{G}$. To ensure easy encoding the corresponding parity check matrix $\mathbf{H}$ is of lower triangular form [3], i.e., $\mathbf{H} = [\mathbf{A}|\mathbf{B}]$, with $\mathbf{A}$ being a sparse (pseudo-)random matrix and $\mathbf{B}$ being staircase lower triangular,

$$\mathbf{B} = \begin{bmatrix} 1 & 0 & \ldots & & 0 \\ 1 & 1 & 0 & \ldots & \\ 0 & 1 & 1 & 0 & \ldots \\ \ldots & 0 & 1 & 1 & 0 \\ 0 & \ldots & 0 & 1 & 1 \end{bmatrix} . \tag{1}$$

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Details on the encoding process can be found in [1]. Two frame sizes for the encoded bits x exist in DVB-S2, regular frames with $N=64800$ and short frames with $N=16200$. The rates $r=M/N$ of the specified LDPC codes vary between $r=1/4$ and $r=9/10$.

Each frame of encoded bits x is permuted by a block interleaver π with I columns, column-wise writing, and row-wise reading. I is the number of bits per modulated channel symbol y, e.g, $I=2$ for QPSK. Thus, in the respective deinterleaver at the receiver, the data of the I bits belonging to a (possibly distorted) received channel symbol z is spread very far apart before being used in the decoder. Furthermore, the (pseudo)-random part $\mathbf{A}$ of the parity check matrix $\mathbf{H}$ has an additional interleaving effect.

In the modulator the permuted bits $\tilde{x}$ are grouped consecutively into bit patterns $\underline{\tilde{x}}_t=[\tilde{x}_t^{(1)},\dots\tilde{x}_t^{(I)}]$, where $\tilde{x}_t^{(i)}$ denotes the i^{th} bit in the bit pattern at (time) index t, $t=1,\dots T$ and $T=N/I$. Each bit pattern $\underline{\tilde{x}}_t$ is mapped according to a mapping rule μ to a complex channel symbol y_t out of the signal constellation set (SCS) $\mathcal{Y}$

$$y_t=\mu(\underline{\tilde{x}}_t) \quad ,y_t\in\mathbb{C} \quad . \tag{2}$$

The respective inverse relation is denoted by μ^{-1}, with

$$\underline{\tilde{x}}_t=\mu^{-1}(y_t)=[\mu^{-1}(y_t)^{(1)},\dots\mu^{-1}(y_t)^{(I)}] \quad . \tag{3}$$

The channel symbols are normalized to an average energy of $E_s=E\{\|y_t\|^2\}=1$.

Similar to [1],[3] we use in this contribution additive white Gaussian noise (AWGN) as channel model for the performance analysis. Thus, complex zero-mean white Gaussian noise $n_t=n_t'+jn_t''$ with a known power spectral density of $\sigma_n^2=N_0$ ($\sigma_{n'}^2=\sigma_{n''}^2=N_0/2$) is added, and the received channel symbols z_t can be written as $z_t=y_t+n_t$.

At the receiver the demodulator (DM) computes *extrinsic* L-values [7] $L_{\text{DM}}^{[\text{ext}]}(\tilde{x})$ for each bit $\tilde{x}_t^{(i)}$ according to [8],[10]

$$L_{\text{DM}}^{[\text{ext}]}(\tilde{x}_t^{(i)})=\log\frac{\sum\limits_{\hat{y}\in\mathcal{Y}_0^i}P(z_t|\hat{y})\cdot P_{\text{BP}}^{[\text{ext,i}]}(\hat{y})}{\sum\limits_{\hat{y}\in\mathcal{Y}_1^i}P(z_t|\hat{y})\cdot P_{\text{BP}}^{[\text{ext,i}]}(\hat{y})} \quad \text{, with} \tag{4}$$

$$P_{\text{BP}}^{[\text{ext,i}]}(\hat{y})\sim\prod_{j=1,j\neq i}^{I}\exp\left(-\mu^{-1}(\hat{y})^{(j)}\cdot L_{\text{BP}}^{[\text{ext}]}(\tilde{x}_t^{(j)}=\mu^{-1}(\hat{y})^{(j)})\right). \tag{5}$$

The numerator and denominator each consist of the sum over all possible channel symbols $\hat{y}$ for which the i^{th} bit of the corresponding bit pattern $\underline{\tilde{x}}=\mu^{-1}(\hat{y})$ is $b\in\{0,1\}$. These channel symbols form the subset $\mathcal{Y}_b^i$ with $\mathcal{Y}_b^i=\{\mu([\tilde{x}^{(1)},\dots\tilde{x}^{(I)}])|\tilde{x}^{(i)}=b\}$.

In the first iteration the feedback L-values $L_{\text{BP}}^{[\text{ext}]}(\tilde{x})$ are initialized as equiprobable, i.e., $L_{\text{BP}}^{[\text{ext}]}(\tilde{x})=0$. The conditional probability density $P(z_t|\hat{y})$ describing the AWGN channel is given by

$$P(z_t|\hat{y})=\frac{1}{\pi\sigma_n^2}\exp(-\|z_t-\hat{y}\|^2/\sigma_n^2) \quad . \tag{6}$$

After appropriately deinterleaving the $L_{\text{DM}}^{[\text{ext}]}(\tilde{x})$ to $L_{\text{DM}}^{[\text{ext}]}(x)$, the $L_{\text{DM}}^{[\text{ext}]}(x)$ are fed back to the belief propagation (BP) decoder for the LDPC code. The BP decoder performs a certain number of "BP iterations" Ξ_{BP} and is modified to compute *extrinsic* information $L_{\text{BP}}^{[\text{ext}]}(x_t^{(i)})$ for the encoded bits in addition to the preliminary estimated decoded data bits $\hat{v}$. For details on possible implementations for the BP decoder using L-values, i.e., the sum-product algorithm and its approximations, we refer to the literature, e.g., [13].

For the next iteration between BP decoder and demodulator the $L_{\text{BP}}^{[\text{ext}]}(x)$ are interleaved again to $L_{\text{BP}}^{[\text{ext}]}(\tilde{x})$ in order to be fed into the demodulator. We denote this iteration as "outer iteration" Ξ_{out}. Thus, in each outer iteration Ξ_{out} the BP decoder executes a certain number of BP iterations Ξ_{BP}. Note, we still use the terminating condition of the BP decoder. Nevertheless, the internal values of the BP decoder are not resetted but saved for the next outer iteration Ξ_{out}. Only the BP decoder's input $L_{\text{DM}}^{[\text{ext}]}(x)$ is updated in each outer iteration.

Thus, the algorithmic modifications necessary for *iterative demodulation* concern only the receiver, which is not specified in the standard anyway. These modifications are also very minor. The $L_{\text{BP}}^{[\text{ext}]}(\tilde{x})$ are computed in the BP decoder in any case and need just to be fed back to the demodulator. The demodulator has to be adapted to accept *a priori* knowledge in terms of L-values. The $L_{\text{BP}}^{[\text{ext}]}(\tilde{x})$ could be additionally used in other adequately adapted components of the receiver, e.g., the equalizer, to further enhance the performance.

III. Signal Constellation Sets (SCS) and Mappings

In the DVB-S2 standard [1] SCSs for QPSK, 8PSK 16APSK, and 32APSK are specified. For the two latter ones the ratio $\gamma=R_2/R_1$ of the radii of the outer and the inner circle of the channel symbols additionally depends on the code rate r. A Gray mapping is used for all SCSs. In [9] it was shown that (at least asymptotically, i.e., considering the error floor) PSK modulation is better suited for *iterative demodulation* than QAM. However, if the channel code is very strong, e.g., a well designed iterative code such as an LDPC code, the error floor might be so low that it is of no interest [11]. Thus, the APSK modulation, which can be seen as an in-between of PSK and QAM, might be a good basis for *iterative demodulation*. Therefore, we concentrate in the following on the 16APSK SCS. In the left half of Fig. 2 the SCS and the mapping defined in the DVB-S2 standard [1] are depicted. We denote this modulation by 16APSK-DVB.

In the following we will discuss different strategies to find mappings suitable for *iterative demodulation*, which usually diverge from the mappings for non-iterative case. For different scenarios, e.g., a strong or weak channel code, the optimum mappings differ. For the given case of DVB-S2 the only modification we allow is a change of the mapping rule μ, since this requires only a new lookup-table and might be an acceptable modification of the standard. It will turn out that the standardized 16APSK-DVB mapping is already a very good, in fact the best, choice due to the strong channel code. Thus, no modification at all is required.

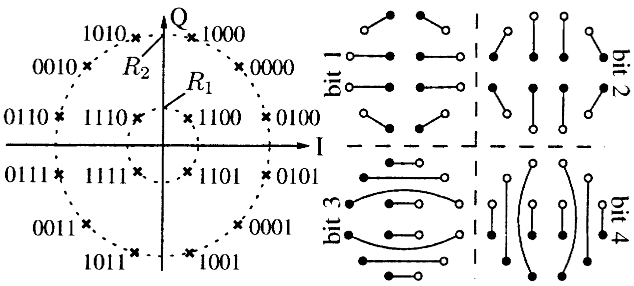

Fig. 2. 16APSK-DVB [1] with EFF decision distances.

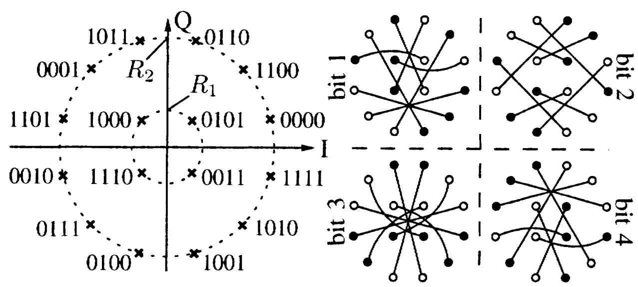

Fig. 3. 16APSK-ASYM with EFF decision distances.

For a theoretical analysis of the asymptotic behavior of the mappings it is assumed that the channel is sufficiently good and enough outer iterations are performed. Then, the feedback values $L_{\text{BP}}^{[\text{ext}]}(\tilde{x})$ can be such reliable that they may be considered as error-free feedback (EFF) [8]. With this EFF, the demodulation degenerates to a simple BPSK decision for each bit. For example, assuming the 16APSK-DVB mapping depicted in Fig. 2 and the EFF is $\tilde{x}^{(2)} = 1$, $\tilde{x}^{(3)} = 1$, and $\tilde{x}^{(4)} = 1$, the demodulator makes a soft decision for bit $\tilde{x}^{(1)}$ only between $\underline{\tilde{x}} = 1111$ and $\underline{\tilde{x}} = 0111$. These decision distances are depicted for the different bit positions on the right side of Fig. 2, where a black dot denotes $\tilde{x}^{(i)} = 1$ and a white dot $\tilde{x}^{(i)} = 0$. Obviously, the larger the distances are, the more reliable the decision will be. Gray mapping, which is the optimum mapping for the non-iterative BICM case [14], usually allows none or only a small gain by *iterative demodulation*. For the iterative BICM-ID other mappings may be required. The optimum mapping regarding the asymptotic performance, i.e., the error floor, for the 16APSK SCS is the 16APSK-ASYM mapping depicted in Fig. 3. This mapping was found by an exhaustive search as described in [9]. As visible, the decision distances for EFF are much larger than for 16APSK-DVB in Fig. 2. For a detailed analysis of asymptotically good mappings for BICM-ID we refer to the literature, e.g., [8],[10],[9].

A. EXIT Chart Analysis

To predict the performance of the different mappings and find a good mapping for the considered problem, i.e., *iterative demodulation* using the DVB-S2 LDPC code, we use extrinsic information transfer (EXIT) charts [12]. Since the mapping shall be used in conjunction with the LDPC code, we first determine the EXIT characteristic of BP decoding the LDPC code (using the sum-product algorithm). The EXIT characteristic for the $r=2/3$ LDPC code with $N=64800$ is depicted in Fig. 4. The *a priori* mutual information $\mathcal{I}_{\text{BP}}^{[\text{apri}]}$ is for $L_{\text{BP}}^{[\text{apri}]} = L_{\text{DM}}^{[\text{ext}]}$ and the *extrinsic* mutual information $\mathcal{I}_{\text{BP}}^{[\text{ext}]}$ for $L_{\text{BP}}^{[\text{ext}]}$. Note, these EXIT characteristics are different from the ones used for analyzing and optimizing the LDPC code itself. In that case two separate EXIT characteristics are determined for the check nodes and variable nodes of the LDPC code, see e.g. [15]. However, in our case we consider the BP decoder as a "black box" and are not interested in its internal behavior. As expected the EXIT characteristic of the whole BP decoder in Fig. 4 obviously depends on the number of BP iterations Ξ_{BP}. Additionally depicted is the EXIT characteristic of a "perfect" channel code with $r=2/3$. As visible, the real LDPC code of DVB-S2 approaches this perfect code quite well. The slope is rather small, especially for $\Xi_{\text{BP}} \geq 25$. Also, perfect *extrinsic* mutual information, i.e., $\mathcal{I}_{\text{BP}}^{[\text{ext}]} = 1$ bit, is reached very fast.

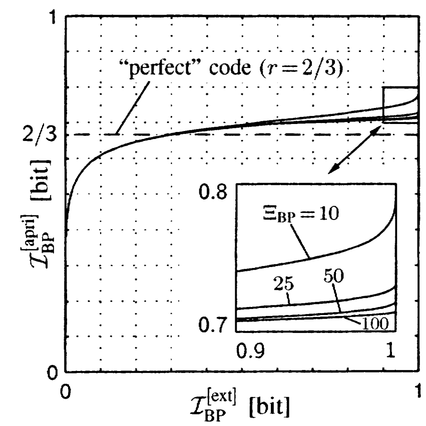

Fig. 4. EXIT characteristics for the $r=2/3$ LDPC code, $N=64800$

In Fig. 5 the EXIT characteristics of the considered 16APSK mappings are compared to the EXIT characteristic for $\Xi_{\text{BP}}=25$ and $\Xi_{\text{BP}}=100$ of the $r=2/3$ LDPC code of Fig. 4. The comparison is made at $E_b/N_0=4.3$ dB, which corresponds to $E_s/N_0=E_b/N_0+10\log(rI)\approx 8.56$ dB. The 16APSK-ASYM mapping exhibits a steep slope yielding the best performance at high $\mathcal{I}_{\text{DM}}^{[\text{apri}]}$. However, this steep slope implies a bad performance at low $\mathcal{I}_{\text{DM}}^{[\text{apri}]}$, resulting in this case of a strong channel code in early intersections at almost $\mathcal{I}_{\text{DM}}^{[\text{apri}]}=0$ bit with the EXIT characteristics of the LDPC code. In contrast, the EXIT characteristic of the 16APSK-DVB mapping with its very slight slope fits the EXIT characteristics of the LDPC already quite well. The intersections occur at $\mathcal{I}_{\text{DM}}^{[\text{apri}]}\approx 0.85$ bit for $\Xi_{\text{BP}}=25$ and $\mathcal{I}_{\text{DM}}^{[\text{apri}]}\approx 1$ bit for $\Xi_{\text{BP}}=100$.

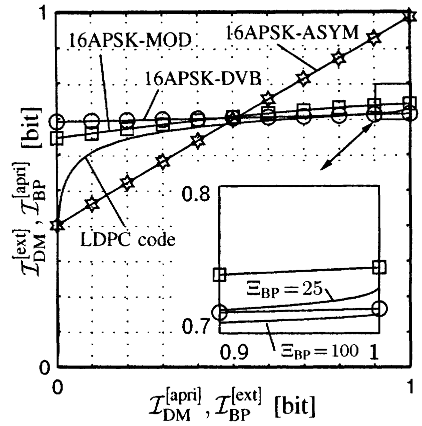

Fig. 5. EXIT characteristics for the different 16APSK mappings, $E_b/N_0 = 4.3$ dB

Thus, if any, only minor changes should be sufficient. By experiments[1] we found that the EXIT characteristic of the mapping depicted in Fig. 6 fits the slope of the $r=2/3$ LDPC code EXIT characteristic almost perfectly. We denote this mapping as 16APSK-MOD and include it in the comparison. Only two bit patterns are exchanged with respect to 16APSK-DVB. With 16APSK-MOD mapping the intersections of the EXIT characteristics in Fig. 5 occur at $\mathcal{I}_{DM}^{[apri]} = 1$ bit for both displayed Ξ_{BP}. However, the simulation results in Section IV indicate that nevertheless the 16APSK-DVB mapping exhibits a better performance.

Several other techniques are known to fit the EXIT characteristic of the mapping better to the one of the channel code. These techniques usually require significant modifications to the standardized transmitter, beyond the simple change of the lookup-table containing the mapping rule μ.

For example, a rate-1 precoder could be inserted before the modulator as proposed in [16],[17]. With this precoder the EXIT characteristic for the 16APSK-DVB mapping would reach the point $(\mathcal{I}_{DM}^{[apri]}, \mathcal{I}_{DM}^{[ext]}) = (1, 1)$. However, this method is not a suitable solution if the transmitter shall remain unmodified to be still compliant with the standard. Obviously, a precoder would require significant modifications to the transmitter as well as the receiver, which seems not feasible and would yield additional computational complexity. Furthermore, the strong LDPC code does not require the EXIT characteristic of the mapping to reach (or even get close) to the upper right corner of the EXIT chart, i.e., $\mathcal{I}_{DM}^{[apri]} = 1$. But with a precoder the EXIT characteristic for, e.g., 16APSK-DVB, would deteriorate at low $\mathcal{I}_{DM}^{[apri]}$.

Another technique is to use two (or more) different mappings in one frame [16], e.g., change the mapping for a certain ratio of channel symbols from 16APSK-DVB to 16APSK-ASYM. The resulting EXIT characteristic is the weighted superposition of the individual EXIT characteristics. But with the LDPC code being irregular, a careful placement of 16APSK-ASYM modulated symbols in the frame would be required.

[1] We started with removing the shortest EFF decision distances. These distances dominate the performance in the iterative case [8],[9],[10].

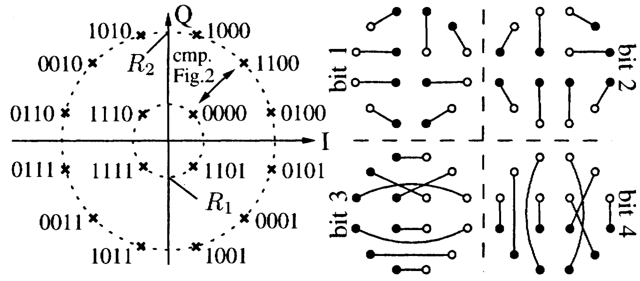

Fig. 6. 16APSK-MOD with EFF decision distances.

IV. Simulation Results

In this section we present simulation results of the BER performance of our proposed algorithm. As LDPC code the $r=2/3$ LDPC code is used with the two frames sizes $N=64800$ (Fig. 7) and $N=16200$ (Fig. 9). In combination with the considered 16APSK SCS ($\gamma=3.15$) the Shannon limit is at $E_b/N_0 \approx 3.64$ dB. For the proposed *iterative demodulation* scheme $\Xi_{BP}=25$ BP iterations are performed in each outer iteration. With $\Xi_{out}=\{2,4,8\}$ the total number of iteration $\Xi_{tot}=\Xi_{out}\cdot\Xi_{BP}$ amounts to $\Xi_{tot}=\{50,100,200\}$. The conventional system without *iterative demodulation* executes a similar number of total iterations, i.e., $\Xi_{out}=1$ and $\Xi_{BP}=\{50,100,200\}$. The non-approximated sum-product algorithm is used for BP decoding. The 16APSK-ASYM mapping is not considered any further due to the comparison of EXIT characteristics in Fig. 5 indicating, that this mapping is not suitable for the used LDPC code.

For the BER with regular sized frames depicted in Fig. 7 the 16APSK-DVB mapping yields gain of up to $\Delta_{E_b/N_0} \approx 0.3$ dB by *iterative demodulation*. The additionally required computational effort is extremely low. The simple update of the $L_{DM}^{[ext]}$ in (4) in every outer iteration, i.e., only once every $\Xi_{BP}=25$ BP iterations, seems negligible compared to numerous calculations in each BP iteration. In the DVB-S2 standard [1] $E_s/N_0 \approx 8.97$ dB is given as performance requirement for the quasi error-free (QEF) case with 50 iterations of the BP decoder. Despite its EXIT characteristic in Fig. 5 suggesting a slightly better performance, the 16APSK-MOD mapping exhibits a worse BER performance than 16APSK-DVB. The reasons for this behavior require further research. Nevertheless, with the 16APSK-DVB mapping showing the superior results, the transmitter requires no modification at all for producing the best performance with *iterative demodulation*.

In Fig. 8 the EXIT chart including the trajectory is depicted for the 16APSK-DVB mapping at $E_b/N_0 = 4.3$ dB. Without *iterative demodulation* the trajectory would stop at $\mathcal{I}_{DM}^{[apri]} \approx 0.56$ bit for all Ξ_{tot}. However, using *iterative demodulation* the trajectory reaches $\mathcal{I}_{DM}^{[apri]} \approx 1$ bit after approximately 4 to 5 outer iterations. Note, for each outer iteration Ξ_{out} the respective BP decoder EXIT characteristic for $\Xi_{out}\cdot\Xi_{BP}$ should be used, since the internal BP decoder values are not resetted. The influence of the changed $L_{DM}^{[ext]}$ should be negligible compared to the respective internal L-values of the BP decoder.

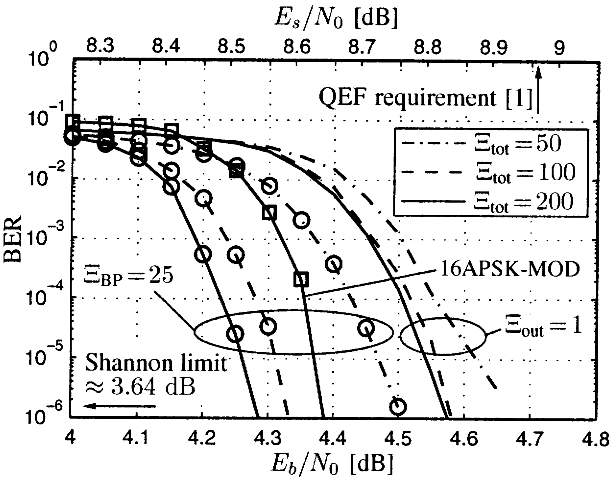

Fig. 7. BER for regular frames $N = 64800$, 16APSK-DVB, $r = 2/3$.

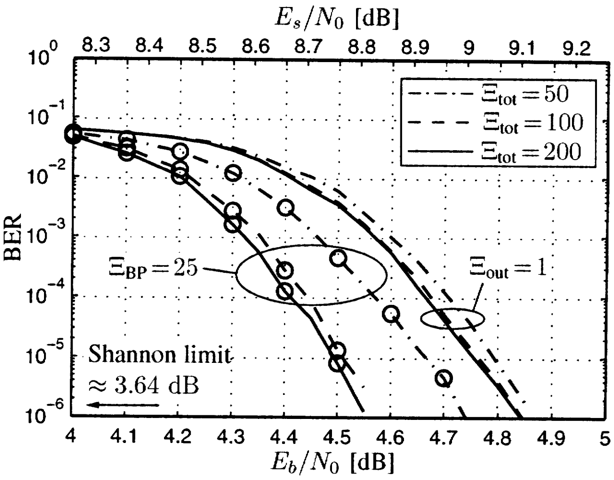

Fig. 9. BER for short frames $N = 16200$, 16APSK-DVB, $r = 2/3$.

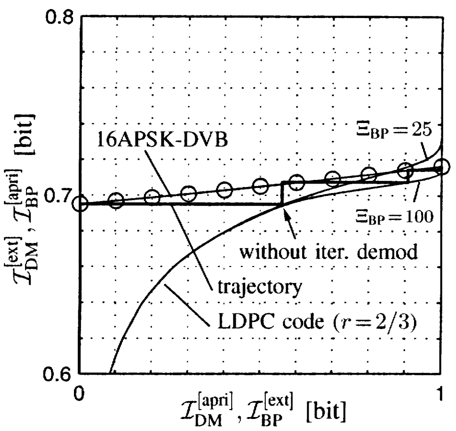

Fig. 8. EXIT Chart for 16APSK-DVB mapping and the $r = 2/3$ LDPC code, frame size $N = 64800$, $E_b/N_0 = 4.3$ dB, $\Xi_{BP} = 25$, $\Xi_{out} = 8$.

The BER simulation results in Fig. 9 show that also for short frames with $N = 16200$ similar gains of up to $\Delta_{E_b/N_0} \approx 0.25$ dB can be achieved. Comparing Figs. 9 and 7 an impact on the performance of ≈ 0.2 dB due to the shorter frame size can be observed (cmp. [1]).

V. Conclusion

In this paper we propose the enhancement of the channel decoding of DVB-S2 by using *iterative demodulation* in conjunction with the LDPC code. Simulation results show a possible gain of $\Delta_{E_b/N_0} \approx 0.3$ dB, which is a noteworthy improvement in relation to the small remaining gap of $\approx 0.6 \ldots 1.2$ dB to the Shannon limit. Of the investigated mappings, the 16APSK-DVB mapping defined in the standard yields the best results. Thus, the transmitter remains completely unmodified. The receiver requires only very minor modifications. The resulting increase in computational complexity is negligible.

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