An Efficient Variable-Length Code Construction for Iterative Source-Channel Decoding

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***Abstract*—We present a novel variable-length code (VLC) construction which exhibits an inherent error correcting capability due to the exclusive presence of codewords with even Hamming weight. Besides error robustness, the proposed code construction features a similar codeword length distribution as Golomb-Rice codes, and therefore, in particular for sources with exponentially distributed symbols, has good source compression properties at the same time. We show that in a source channel coding framework with outer source encoding, inner channel encoding with a recursive convolutional code, and iterative decoding the proposed VLC construction can lead to significant performance improvements compared to fixed-length source encoding with optimized mappings. In particular, simulation results for the AWGN channel verify that for Gauss-Markov sources a performance close to the theoretical limit can be achieved.**

***Index Terms*—Variable-length codes, joint source channel coding, iterative decoding, serially concatenated codes**

I. Introduction

THE robust transmission of variable-length encoded source signals over wireless channels has become an active research area during the last years. It is motivated by the increasing demand on multimedia and data services in third and fourth generation wireless networks. In such applications source compression is usually carried out using standardized techniques which, in order to achieve high compression gains, often employ variable-length codes (VLCs).

VLCs were originally designed for error-free transmission scenarios. In this form they are highly sensitive to transmission errors and suffer from error propagation; for example, single bit errors may lead to insertions and deletions of symbols, causing a loss of synchronization. In order to make VLCs applicable to the peculiarities of wireless channels, a variety of joint source-channel coding and decoding schemes addressing the error-resilient transmission of VLCs have been proposed in the literature. This includes the design of error-resilient VLCs, the joint optimization of VLCs and channel codes, the robust decoding of VLCs, and the joint decoding of VLCs and channel codes. For delay- and / or complexity-constrained transmission scenarios those combined techniques are often more advantageous than the classical separation of source and channel encoding.

A class of robust VLCs is given by reversible variable-length codes (RVLCs) [2]–[5], which are a well established technique for robust video compression. RVLCs exploit additional redundancy in the VLC code words in order to guarantee bidirectional decoding. If the number of error events is low, the impact of a synchronization loss can be mitigated. In [6] variable-length error-correcting codes (VLECs) were introduced, which combine the variable-length character of VLCs with distance constraints of channel codes at the cost of additional redundancy. VLECs are furthermore addressed in [7], [8] where efficient and improved algorithms for the code design are discussed.

A key component for a reliable transmission of VLCs is given by the VLC decoder. The first maximum *a posteriori* probability (MAP) decoder was introduced in [9], where an optimal VLC sequence estimation was carried out by the Viterbi algorithm, applied to an appropriate VLC trellis. Another trellis representation, which has become the standard trellis for VLC decoding, was introduced in [10]. It benefits from an efficient state representation and allows both a bitwise MAP and *a posteriori* probability (APP) decoding with the BCJR algorithm [11] and an optimal sequence estimation with the Viterbi algorithm. Later on, both exact and complexity-reduced approximate MAP decoding techniques for variable-length source codes were proposed by different authors, either by using a modified or list Viterbi algorithm [12]–[17] or by using sequential decoding approaches [18], [19].

Many approaches (e.g., [13]–[15]) exploit the redundancy due to the residual source correlation after source encoding for additional error protection. Generally, this allows for a less powerful forward error correction (FEC) code, resulting in a reduction of allocated bandwidth or latency for the overall transmission system. As reported in [20]–[24], a significant amount of residual source correlation, modeled as first-order Markov process, can be observed for several output parameters in state-of-the-art speech codecs and the MPEG audio compression scheme.

Also, recently, distributed source compression schemes based on Wyner-Ziv coding [25] have been proposed, e.g. for video signals (see, e.g., [26]–[28]). In these schemes, typically the majority of the encoding complexity is shifted to the decoder and therefore, they can be applied in situations where computational complexity is a critical resource at the encoder, for example in sensor networks. As a consequence, there exists a certain amount of residual source redundancy which cannot

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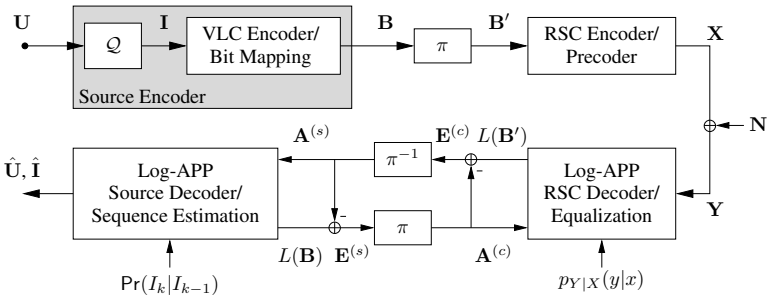

Fig. 1. Model of the transmission system containing source encoding, channel encoding, and iterative source-channel decoding for the AWGN channel.

and provide log-likelihood ratios (LLRs) [55] for the corresponding symbols at the encoder side. For channel decoding, we apply the BCJR algorithm [11] to the trellis representation of the underlying RSC code and assume perfect knowledge of the noise statistics $p_{Y|X}(y\,|\,x)$ at the decoder. Source decoding in the VLC case is carried out by the bit-based Log-APP VLC decoder proposed in [43]. By appropriately modifying the BCJR algorithm for the trellis from [10], the residual source correlation, modeled by the index-transition probabilities $\Pr(I_k|I_{k-1})$, can be exploited for additional error protection. Due to a low number of trellis states this results in an efficient decoding algorithm which is especially well-suited for decoding of large block lengths. Estimates for the source symbols U_k and the quantization indices I_k are obtained from a sequence estimation on the VLC trellis [10] during the final iteration. Note that an optimal symbol estimation is not feasible in the VLC case for large block length since its complexity increases with K^2 [43]. This is in contrast to fixed-length source coding where symbol-wise Log-APP decoding can be realized by choosing the source symbols as trellis states. The optimal minimum mean-squared error estimator and the maximum *a posteriori* estimator for estimating U_k and I_k are applicable in this case.

The iterative decoding process is based on the exchange of extrinsic information [55] between the two component decoders. Based on the observations $\mathbf{Y} = [Y_1, \ldots, Y_M]$ for the transmitted symbols X_m at the output of the communication channel and on the *a priori* LLRs $\mathbf{A}^{(c)} = [A_1^{(c)}, \ldots, A_N^{(c)}]$ for the interleaved symbols $B'_{n'}$, the inner channel decoder computes the vector $L(\mathbf{B}')$ of *a posteriori* LLRs $L(B'_{n'}\,|\,\mathbf{Y}, \mathbf{A}^{(c)})$ for the interleaved source bits $B'_{n'}$. Similarly, given the vector of *a priori* LLRs $\mathbf{A}^{(s)} = [A_1^{(s)}, \ldots, A_N^{(s)}]$ the outer source decoder computes the vector $L(\mathbf{B})$ of *a posteriori* LLRs $L(B_n\,|\,\mathbf{A}^{(s)})$ for the bits B_n at the output of the source encoder. The extrinsic LLRs $E_{n'}^{(c)}$ and $E_n^{(s)}$, which are exchanged between the two decoders, can now be obtained by subtracting the *a priori* LLRs $\mathbf{A}^{(c)}$ and $\mathbf{A}^{(s)}$ from the respective *a posteriori* LLRs $L(B'_{n'}\,|\,\mathbf{A}^{(c)})$ and $L(B_n\,|\,\mathbf{A}^{(s)})$ as shown in Fig. 1.

After (de-)interleaving, the extrinsic LLRs provided by the one decoder become the *a priori* input of the other decoder. Since no extrinsic information is available at the output of the outer decoder during the initial decoding step, the *a priori* LLRs $A_{n'}^{(c)}$ of the inner decoder are initialized with zeros. The iterative process continues until no transmission errors can be observed anymore for the given block or an appropriate stopping criterion is satisfied.

C. Code Rate of the Transmission System

Since the source correlation ameliorates the achievable error protection, the overall code rate R of the transmission system relies on both the explicit redundancy of the channel code and the amount of residual source redundancy. Therefore, we associate the source redundancy with a code rate R_S. It can be separated into the product of the code rate R_{corr}, which corresponds to the redundancy due to the intra-frame correlation, and the code rate R_{Map}, which is associated with the redundancy introduced by the bit mapping or the VLC. With $H(X)$ denoting the entropy of a random variable X, the rate R_S can be expressed as

$$R_S = \underbrace{\frac{H(I_k) + (K-1)\cdot H(I_k|I_{k-1})}{K \cdot H(I_k)}}_{=:R_{corr}} \cdot \underbrace{\frac{H(I_k)}{\bar{l}}}_{=:R_{Map}}. \qquad (1)$$

Due to the serial concatenation of source and channel encoding, the overall code rate R of the transmission system is then given by the product of the code rate R_S of the source encoder and the code rate R_{RSC} of the RSC code: $R = R_S \cdot R_{RSC} = R_{corr} \cdot R_{Map} \cdot R_{RSC}$.

III. Even-Weight Variable-Length Codes

In the following, we propose a simple VLC construction which provides good compression properties and at the same time guarantees a minimum Hamming distance of $d_{min} = 2$ between code sequences of equal length. These properties are especially useful for the design of efficient encoding schemes for iterative source-channel decoding, which will be discussed in Section IV.

A. Code Construction

In order to satisfy the distance constraint $d_{min} = 2$, we only consider codewords with even Hamming weight in the codeword set $\mathcal{C}$. Further constraints as symmetry or reversibility are neglected. We thus obtain an efficient VLEC with good error-correction properties.

The construction of the EW-VLC for a set $\mathcal{I}$ of source symbols I_k is parametrized by the minimum code word length l_{min}. In the following, l_{min} is set to the minimum word length of the Huffman code, derived for the underlying source distribution $\Pr(I_k)$. The derivation of the codeword set $\mathcal{C}$ for the EW-VLC can then be summarized as follows:

1. Initialization of
   - the current code word length $l = l_{min}$,
   - the number of assigned code words $A = 0$, and
   - the set of available length-l codewords $\underline{\mathcal{V}}_l = \{\mathbf{b} \,|\, \mathbf{b} \in \mathbb{B}^l\}$.
2. The code word candidates $\mathbf{b}$ available in the set $\underline{\mathcal{V}}_l$ are separated into the set of code words with *even* Hamming weight,
$$\underline{\mathcal{E}}_l := \{\mathbf{b} \,|\, \mathbf{b} \in \underline{\mathcal{V}}_l,\ w_H(\mathbf{b}) \text{ is even}\},$$
and into the set of code words with *odd* Hamming weight,
$$\underline{\mathcal{O}}_l := \{\mathbf{b} \,|\, \mathbf{b} \in \underline{\mathcal{V}}_l,\ w_H(\mathbf{b}) \text{ is odd}\}.$$
3. We assign a number of $\min\{|\underline{\mathcal{E}}_l|, |\mathcal{I}| - A\}$ code words $\mathbf{b} \in \underline{\mathcal{E}}_l$ with even Hamming weight to the code word set $\mathcal{C}$ of the EW-VLC.
4. The number of assigned code words, the set of available code words, and the current code word length are updated according to
$$\begin{aligned} A &\leftarrow A + \min\{|\underline{\mathcal{E}}_l|, |\mathcal{I}| - A\} \\ \underline{\mathcal{V}}_{l+1} &= \{\mathbf{b} = [b_1, \ldots, b_{l+1}] \,|\, [b_1, \ldots, b_l] \in \underline{\mathcal{O}}_l\} \\ l &\leftarrow l + 1. \end{aligned}$$
5. If all code words are assigned ($A = |\mathcal{I}|$), the code construction is completed; otherwise return to step 2.

After constructing the code word set $\mathcal{C}$, the codewords are assigned to the symbols I_k in $\mathcal{I}$ such that highly probable symbols are represented by short codewords. An example for the EW-VLC is given in Table I for the English alphabet and a minimum code-word length $l_{min} = 3$. The probabilities $\Pr(I_k)$ of the letters in Table I are taken from [5].

For the performance of other VLC construction algorithms for the English alphabet we refer to Tab. I in [5].

TABLE I
EW-VLC WITH $l_{min} = 3$ FOR THE ENGLISH ALPHABET. THE PROBABILITIES $\Pr(I_k)$ FOR THE LETTERS I_k ARE TAKEN FROM [5].

I_k	$\Pr(I_k)$	EW-VLC	I_k	$\Pr(I_k)$	EW-VLC
E	0.148786	000	F	0.024553	010 001
T	0.093541	011	M	0.023619	100 001
A	0.088337	101	C	0.020817	111 001
O	0.072458	110	W	0.018682	001 000 1
R	0.068722	001 1	G	0.015212	010 000 1
N	0.064985	010 1	Y	0.015212	100 000 1
H	0.058313	100 1	B	0.012677	111 000 1
I	0.056445	111 1	V	0.011609	001 000 01
S	0.055378	001 01	K	0.008674	010 000 01
D	0.043768	010 01	X	0.001468	100 000 01
L	0.041233	100 01	Y	0.000801	111 000 01
U	0.027622	111 01	Q	0.000801	001 000 001
P	0.025754	001 001	Z	0.000534	010 000 001

B. Properties

In the following, we prove some fundamental properties of the EW-VLC with code-word set $\mathcal{C}$.

Proposition 1. *The code in $\mathcal{C}$ is a prefix code.*

Proof: Let $\mathcal{C}_l$ denote the set of VLC codewords with the maximal length l. For $l = l_{min}$, all codewords in $\mathcal{C}_{l_{min}}$ have the same lengths. For $l \rightarrow l + 1$, a codeword of length $l + 1$ in $\mathcal{C}_{l+1}$ has a prefix of length l in $\underline{\mathcal{O}}_l$ by construction. Since $\underline{\mathcal{E}}_l \subseteq \mathcal{C}_l$ but $\underline{\mathcal{O}}_l \cap \underline{\mathcal{E}}_l = \emptyset$, $\mathcal{C}_{l+1}$ is prefix-free. ■

Proposition 2. *The code in $\mathcal{C}$ has a minimum Hamming distance of two.*

Proof: All codewords in $\mathcal{C}$ have even Hamming weight. Consider two codeword sequences $\mathbf{B}$ and $\hat{\mathbf{B}}$ of the same bit length N, which contain at least one different source symbol. Since $\mathbf{B}$ and $\hat{\mathbf{B}}$ have even Hamming weight, they must differ in at least two bit positions, leading to $d_{min} = 2$. ■

The following theorem addresses the length distribution of EW-VLCs which has some important implications on the source compression performance of these codes.

Theorem 1. *Let the length distribution of a VLC be defined by the number A_l of length-l code words. Then, the length distribution $A_l^{(\mathcal{C})}$ of the code-word set $\mathcal{C}$ is equivalent to the length distribution $A_l^{(GR)}$ of a Golomb-Rice code.*

Proof: By construction, the length distribution $A_l^{(\mathcal{C})}$ of the code-word set $\mathcal{C}$ is given by
$$A_l^{(\mathcal{C})} = \begin{cases} 0 & \text{for } l \notin [l_{min}, l_{max}], \\ 2^{l_{min}-1} & \text{for } l_{min} \le l < l_{max}, \\ |\mathcal{I}| - (l_{max} - l_{min}) \cdot 2^{l_{min}-1} & \text{for } l = l_{max}. \end{cases}$$
The length distribution $A_l^{(GR)}$ of a Golomb-Rice code with $q = \lfloor i/2^k \rfloor$, $k \in \mathbb{N}$, which encodes an integer number i to a code word composed of a length-q prefix and a length-k suffix, is given by
$$A_l^{(GR)} = \begin{cases} 0 & \text{for } l \le k, \\ 2^k & \text{for } l > k. \end{cases}$$
By applying the Golomb-Rice code to the finite set $\mathcal{I} = \{0, \ldots, 2^W - 1\}$, the equality of both length distributions becomes obvious for $k = l_{min} - 1$. ■

From Theorem 1 it follows that both the Golomb-Rice code and the EW-VLC lead to the same code rate R_{Map}. The connection between the length distributions of the proposed EW-VLC and a Golomb-Rice code becomes furthermore obvious from Table I. Since Golomb-Rice codes are known to be nearly optimal for sources with exponentially distributed symbols, we can expect from Theorem 1 that the proposed EW-VLC shows a good performance in terms of compression and efficiency. Furthermore, due to the minimum Hamming distance of two the EW-VLC construction exhibits an error-correction capability. This property becomes particularly important if this code is employed in a joint source-channel coding framework as described in Section II.

C. Performance

Table II verifies the good compression properties of the EW-VLC which we expect from Theorem 1. For the distribution of the English alphabet and for a Laplacian, a Gaussian, and a uniform distribution with $|\mathcal{I}| = 32$ the code rate R_{Map} of the EW-VLC is compared with the code rates achieved by a Huffman code and by asymmetric and symmetric reversible

variable-length codes (RVLCs) from [4], [5], respectively. It is evident from Table II that the EW-VLC outperforms the RVLCs with respect to their compression properties. Furthermore, as a comparison with Table III from [8] reveals, the performance of the EW-VLC is similar to the performance of the VLEC from [8] for the case $d_{min} = 2$. However, the VLC construction introduced in the previous section is much simpler than the iterative approach proposed in [8].

TABLE II
COMPARISON OF THE CODE RATE R_{Map} FOR A EVEN-WEIGHT VARIABLE-LENGTH CODE (EW-VLC) ($d_{min} = 2$), A HUFFMAN CODE ($d_{min} = 1$), AND REVERSIBLE VARIABLE-LENGTH CODES (RVLCS) PROPOSED IN [4] ($d_{min} = 1$) AND [5] ($d_{min} = 2$).

	Code rate R_{Map}			
	Engl. alphabet	Laplace	Gaussian	uniform
Huffman	0.9916	0.9926	0.9899	1
EW-VLC	0.9726	0.9818	0.9736	0.9091
asym. RVLC [4]	0.9612	0.9638	0.9568	1
asym. RVLC [5]	0.9466	0.9657	0.9484	0.9091
sym. RVLC [4]	0.8640	0.9122	0.8944	0.7692
sym. RVLC [5]	0.8307	0.8993	0.8852	0.7143

In particular, for Laplacian and Gaussian distributions the performance of EW-VLCs is close to the one of Huffman codes. This observation confirms our conclusion from Theorem 1. Finally, comparing Table I with Table I from [5] and Table III from [8] reveals that EW-VLCs lead to a shorter maximum codeword length than asymmetric RVLCs from [4], [5], whereas the maximum codeword length is equivalent to the one for the VLECs from [8] with $d_{min} = 2$. This is of relevance since the complexity of symbol-based decoding algorithms (e.g., [33], [38], [40]) depends on the difference between maximum and minimum codeword length.

IV. ENCODER DESIGN FOR ITERATIVE SOURCE-CHANNEL DECODING

In this section we address an optimization of source and channel encoder in Fig. 1 for iterative source-channel decoding. In particular, the properties of established bit mappings and variable-length codes with respect to the design guidelines for serial concatenated channel codes will be analyzed. We will show in the following that both conventional quantizer bit mappings and VLCs lead to several drawbacks which can be avoided by EW-VLCs. In addition, for both EW-VLCs and a fixed-length source encoding scheme with the optimized bit mappings from [45] we present optimization results for the inner channel encoder.

A. Convergence Analysis and Design Criteria

A powerful framework for the analysis of the convergence behavior of the iterative decoder is given by EXIT characteristics [50]. Based on information theoretic measures this approach allows for precise performance predictions of concatenated codes under iterative decoding without the need to fully simulate the iterative decoder.

In order to compute EXIT characteristics SISO decoders are treated as information filters. They are characterized by the amount of mutual information. $\mathrm{I}_{extr}^{(s/c)} := \mathrm{I}(E^{(s/c)}; B)$ between their extrinsic output $E^{(s/c)}$ and the corresponding quantity B at the encoder, given the corresponding mutual information $\mathrm{I}_{apri}^{(s/c)} := \mathrm{I}(A^{(s/c)}; B)$ between the *a priori* input $A^{(s/c)}$ and the underlying symbols B. With the mutual information $\mathrm{I}_c := \mathrm{I}(X_m; Y_m)$ between the received and transmitted symbols X_m and Y_m at the input and output of the channel, resp., the EXIT characteristics specifying the source and the channel decoder are defined by the mappings

$$\mathcal{X}^{(s)}: \ \mathrm{I}_{apri}^{(s)} \ \mapsto \ \mathrm{I}_{extr}^{(s)} = \mathcal{X}^{(s)}(\,\mathrm{I}_{apri}^{(s)}\,) \qquad \text{and}$$
$$\mathcal{X}^{(c)}: \ \mathrm{I}_{apri}^{(c)} \ \mapsto \ \mathrm{I}_{extr}^{(c)} = \mathcal{X}^{(c)}(\,\mathrm{I}_{apri}^{(c)} \,|\, \mathrm{I}_c\,).$$

Note that, since the inner decoder also utilizes the *a priori* input for the channel observations, the EXIT characteristic of the inner channel decoder is conditioned on I_c.

In order to avoid a convergence towards high bit error rates, the components of the encoder should be chosen in such a way that the EXIT characteristic $\mathcal{X}^{(c)}$ lies above ${\mathcal{X}^{(s)}}^{-1}$ and intersections at $(\mathrm{I}_{apri}, \mathrm{I}_{extr}) < (1,1)$ bits / channel use[2] are avoided. As a consequence, the inner encoder must be recursive [49], [50], and the minimum Hamming distance $d_{min}^{(o)}$ of the outer encoder must satisfy $d_{min}^{(o)} \geq 2$. Both recommendations confirm the design guidelines derived from the analysis of the upper bound to the bit error probability in [46].

Another design criterion follows from the area property of EXIT characteristics, shown in [56] for the binary erasure channel (BEC). It states that the area $\mathcal{A}({\mathcal{X}^{(s)}}^{-1})$ under the inverse EXIT characteristic of the outer decoder corresponds to its code rate $R^{(o)}$, while the area $\mathcal{A}(\mathcal{X}^{(c)})$ under the inner EXIT characteristic $\mathcal{X}^{(c)}$ is lower or equal to the ratio of the mutual information I_c of the communication channel and the inner code rate $R^{(i)}$:

$$\mathcal{A}({\mathcal{X}^{(s)}}^{-1}) = R^{(o)} \quad \text{and} \quad \mathcal{A}(\mathcal{X}^{(c)}) \leq \mathrm{I}_c / R^{(i)}.$$

Since equality holds for $R^{(i)} = 1$ it is reasonable to choose inner rate-1 codes. Otherwise, the transmission system may suffer from a capacity loss [56] which may increase the decoding threshold towards higher channel signal-to-noise ratios (SNRs).

B. Inverse EXIT Characteristics for the Inner Source Decoder

For both standard and optimized bit mappings, different types of RVLCs, and the EW-VLC introduced in Section III inverse EXIT characteristics for the inner source decoder [43] are depicted in Fig. 2(a). Here, the source signal is modeled by a first-order autoregressive process which is realized by filtering uncorrelated Gaussian noise with a recursive filter $G(z) = z/(z-a)$ where a is the correlation coefficient. The EXIT characteristics are obtained for source vectors of $K = 20000$ symbols, a correlation coefficient $a = 0.9$, and a 4-bit scalar quantization with a Lloyd-Max quantizer.

As shown in Fig. 2(a), for standard bit mappings as the Gray mapping or the natural binary mapping, the outer source decoder provides only a low amount of information

[2]Note that the maximum amount of information which can be obtained on a bit B from a noisy observation Y is given as $I(B;Y) = H(B) - H(B|Y) \leq H(B)$. Throughout this paper we assume that by appropriate source coding $H(B) = 1$ can be achieved which is fulfilled in good approximation, e.g., for all VLCs with $d_f \geq 2$ in Fig. 2(a).

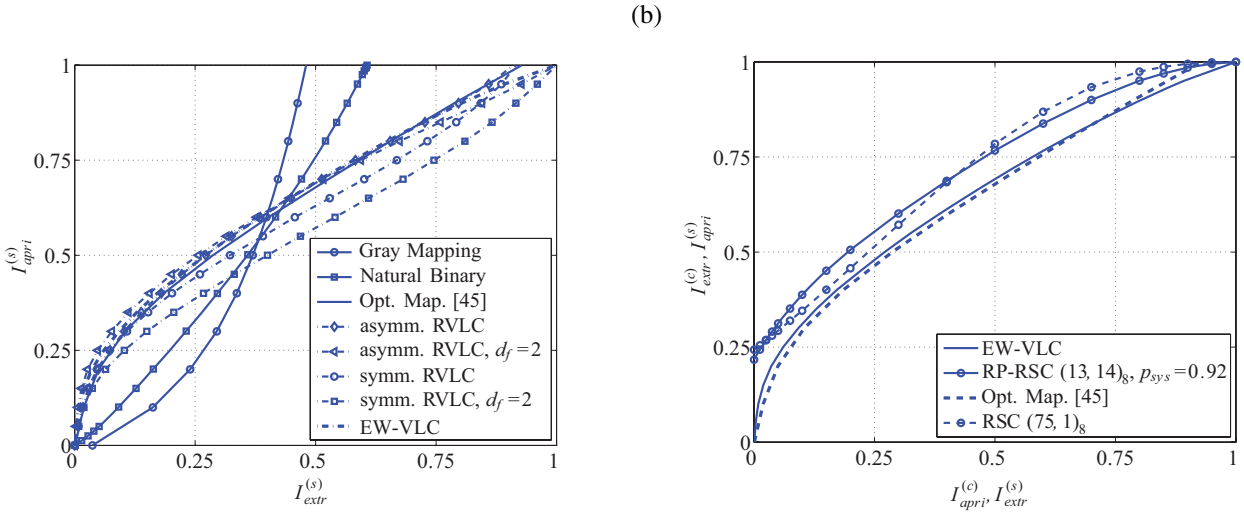

Fig. 2. Inverse EXIT characteristics $\mathcal{X}^{(s)\,-1}$ of the (outer) source encoder / decoder pair. (a): comparison between standard and optimized bit mappings, reversible VLCs, and the EW-VLC for a first-order autoregressive process with correlation coefficient $a = 0.9$, quantized by a 4-bit Lloyd-Max quantizer; (b): EXIT characteristics for outer source codes and the corresponding inner channel codes.

at its extrinsic output, given perfect *a priori* information ($\mathrm{I}^{(s)}_{apri} = 1\,\text{bits / channel use}$). Early intersections with the EXIT function of the inner decoder are the consequence, leading to a poor performance of the transmission system.

To overcome this problem, optimized mappings have been proposed in [45], [57], which are based on a maximization of the distortion for the case of one-bit errors and on a maximization of the extrinsic information for optimal *a priori* information, respectively. The inverse EXIT characteristic for the mapping from [45] is included in Fig. 2(a) which shows the superiority of the optimized mappings to the conventional ones. However, we are still not able to obtain iterative decoder convergence at $(\mathrm{I}_{apri}, \mathrm{I}_{extr}) = (1, 1)\,\text{bits / channel use}$.

A similar behavior of the iterative decoder can be expected for the asymmetrical RVLC proposed in [4]. By considering an additional symmetry constraint in the RVLC design [4], the robustness can be further improved, however, at the cost of a decreased code rate R_{Map}. In this case, the mean word length $\bar{l}$ is larger than the quantization word length W, i.e., the symmetrical RVLC introduces some redundancy.

Neither the analyzed mappings nor the RVLCs from [4] fulfill the condition $d^{(o)}_{min} \geq 2$. As a consequence, transmission systems which employ one of these codes can generally not benefit from an interleaver gain. This drawback can be avoided with the EW-VLC introduced in Section III. As Fig. 2(a) shows, $(\mathrm{I}_{apri}, \mathrm{I}_{extr}) = (1, 1)\,\text{bits / channel use}$ is achieved in the EXIT characteristics, and in addition a slight compression can be obtained compared to the bit mappings with constant word length. Assuming an appropriate (recursive) inner encoder, convergence at $(\mathrm{I}_{apri}, \mathrm{I}_{extr}) = (1, 1)\,\text{bits / channel use}$ becomes possible, and the interleaver gain can be exploited. On the other hand, RVLCs which satisfy the distance constraint $d^{(o)}_{min} \geq 2$ are proposed in [5], and their inverse EXIT characteristics are included in Fig. 2(a) as well. However, as we can see in Table II and conclude from the area under the inverse EXIT characteristic, the code rate is decreased compared to the EW-VLC; i.e., additional redundancy is introduced due to the additional reversibility constraint.

C. Code Matching for the Inner and Outer Encoder

In the following, we focus on the design of the inner RSC code. The goal is to find a good-matching inner code whose EXIT characteristic lies above the inverse EXIT characteristic of the outer source decoder. Therefore, we restrict ourselves to the EW-VLCs from the previous section, since they offer the best trade-off between amount of compression and distance properties. As a benchmarking scheme for our EW-VLC-based approach we consider fixed-length quantizer indices with an optimized bit mapping [45] in the source encoding stage.

As Fig. 2(a) shows, the outer source decoder provides only a small amount of extrinsic information $\mathrm{I}^{(s)}_{extr}$ for $\mathrm{I}^{(s)}_{apri} \leq 0.25\,\text{bits / channel use}$ for both EW-VLC and optimized bit mapping. Therefore, in order to avoid an early intersection of the EXIT characteristics, the inner channel encoder has to be chosen such that in absence of *a priori* knowledge ($\mathrm{I}^{(c)}_{apri} = 0\,\text{bits / channel use}$) a minimum amount of extrinsic information $\mathrm{I}^{(c)}_{extr}(0) \geq 0.1\,\text{bits / channel use}$ is guaranteed.

For rate-1 RSC codes, which are obtained by puncturing the systematic bits, it was observed in [50] that only codes whose feed-forward polynomials consist of a single coefficient are able to generate extrinsic information $\mathrm{I}^{(c)}_{extr}(0) > 0\,\text{bits / channel use}$ without additional *a priori* knowledge. However, especially for a short code memory, these codes have the drawback that only few candidates are available such that the code design becomes quite constrained.

As a remedy, we apply randomly punctured (RP) RSC codes [52] as suggested in Section II-A. Compared to the optimization of conventional puncturing patterns, the design of appropriate RP-RSC codes is drastically simplified. The puncturing process is parametrized by a single parameter, the puncturing rate p_{sys}. The result of the optimization for the EW-VLC from Section III is shown in Fig. 2(b) for the BI-AWGN channel with $E_b/N_0 = 2$ dB and $E_b = E_s/R$. Here, the overall code rate $R = 0.66$ is considered, obtained for the EW-VLC and the source parameters from above. A good-matching code is found by randomly puncturing a fraction $p_{sys} = 92\%$ of the systematic bits and a fraction $p_{par} = 8\%$

of the parity bits from the memory-3 RSC code with generator polynomials $(13, 14)_8$.

For the optimized bit mapping from [45] we apply the memory-5 RSC code [58], where the corresponding EXIT characteristic is shown in Fig. 2(b) for the same parameters of the BI-AWGN channel. It is obtained by puncturing all systematic bits from the rate-1/2 mother code with the generator polynomials $(75, 1)_8$. Since in this case the EXIT characteristic for the source decoder does not achieve the point $(\mathrm{I}_{apri}, \mathrm{I}_{extr}) = (1, 1)$ bits / channel use, an RSC code with larger memory must be applied in order to lower the error floor.

V. Simulation Results

In the following we verify the good performance of the EW-VLC from Section III in the context of iterative source-channel decoding. Monte Carlo simulations are performed for the BPSK-modulated BI-AWGN channel by using the code constructions from Section IV-C. For the sake of completeness, a second reference system based on the asymmetric RVLC (aRVLC) from [5] and the rate-1 randomly punctured RSC code $(13, 14)_8$ with $p_{sys} = 92\%$ is considered as well.

The results are presented in Fig. 3 for the symbol error rate (SER_L) based on the Levenshtein metric[3] and for the distortion in terms of reconstruction signal-to-noise ratio ($R-SNR$). In order to allow a fair comparison between different approaches the channels are parametrized by the channel SNR E_b/N_0 related to the transmit energy per information bit $E_b = E_s/R$. The simulations are carried out for source vectors of $K = \{200, 2000, 20000\}$ symbols, a correlation coefficient $a = 0.9$, and a 4-bit scalar quantization with a Lloyd-Max quantizer, similar to the parameters chosen in Section IV-B. All coding schemes achieve approximately the same overall code rate of $R \approx 0.66$. All interleavers are realized by s-random interleavers [59].

Fig. 3 shows the good performance of the EW-VLC combined with the randomly punctured RSC code $(13, 14)_8$. For a block length of $K = 20000$ symbols and $E_b/N_0 \geq 1.6$ dB no error could be observed within our simulation accuracy after 40 iterations[4], and we achieve a threshold within 0.6 dB of the capacity limit for $R = 0.66$. We outperform the coding scheme based on the aRVLC [5] by 0.2 dB. Compared to the fixed-length approach with an optimized mapping from [45] and a memory-5 RSC code a gain of 0.5 dB in channel SNR with respect to clear channel quality can be observed in Fig. 3(b). For the optimized mapping convergence is obtained after 10 iterations, even though the transmission suffers from a residual SER of approximately $2 \cdot 10^{-4}$.

For an intermediate block length of $K = 2000$ symbols the performances of both systems deteriorate. Despite the VLC approach now also shows an error-floor it still outperforms the optimized mapping: for $E_b/N_0 \geq 1.75$ dB and $K = 2000$ symbols the achieved distortion is comparable to the one obtained by the fixed-length-based scheme for a block length of $K = 20000$ symbols. If the block length is further reduced to $K = 200$ symbols the optimized mapping becomes superior for the channel SNR regime shown in Fig. 3.

The SER comparison in Fig. 3(a) for different block / interleaver lengths shows the behavior we expect according to the discussion from Section IV-B: due to a minimum distance of $d_{min}^{(o)} = 1$ the system based on the optimized mapping suffers from a residual error floor for $E_b/N_0 \geq 2.25$ dB. In contrast, the EW-VLC with $d_{min}^{(o)} = 2$ takes advantage of the available interleaver gain: while the SER only slowly decreases for $E_b/N_0 \geq 2.5$ dB and a block length of $K = 2000$ symbols, no error event could be observed within our simulation accuracy for $E_b/N_0 \geq 1.6$ dB and $K = 20000$ symbols.

Finally, the special case of an uncorrelated source ($a = 0$) is considered in Fig. 4 for a block length of $K = 20000$ symbols. As we can see, the joint source-channel coding scheme employing the EW-VLC clearly outperforms the scheme based on the optimized mapping [45]. The large gains can be explained from the properties of the EXIT functions for the optimized mapping: for optimal *a priori* information $\mathrm{I}_{apri}^{(s)} = 1$ bits/channel use the source decoder employing the optimized mapping provides only the extrinsic information $\mathrm{I}_{extr}^{(s)} = 0.187$ bits/channel use, leading to early intersections of the EXIT functions. In contrast, the EXIT functions in the EW-VLC case achieve the point $(1, 1)$ bits/channel use in the EXIT chart. It is thus straightforward to find a matching channel code which leads to a good performance as shown in Fig. 4. Note that these results hold similarly for small correlation coefficients $a > 0$ with the difference that the extrinsic information $\mathrm{I}_{extr}^{(s)}$ is slightly higher than for $a = 0$.

VI. Conclusions

We have presented a simple variable-length code construction which is well suited for transmission systems employing iterative source-channel decoding. While providing good compression properties compared to, e.g., RVLCs the proposed EW-VLC guarantees a minimum Hamming distance of two between equal-length code sequences. Both features are the key to transmission systems which are extremely efficient in terms of required overhead for error protection and robust at the same time. As the simulation results for correlated Gauss-Markov sources show, the proposed EW-VLC facilitates reliable communication close to the AWGN channel capacity solely by exploiting residual source redundancy. Finally, a comparison shows that the proposed EW-VLC-based scheme outperforms the best fixed-length bit mapping optimized for iterative source-channel decoding. These results suggest to employ well-designed variable-length codes instead of fixed-length mappings in the context of iterative source-channel decoding.

[3]The Levenshtein metric of two sequences gives the number of insertions, deletions, and / or substitutions which are needed to transform one sequence into the other. An efficient approximation is proposed in [6].

[4]Note that in this case a very low error floor exists which may not be measured by Monte Carlo simulations in a reasonable time.

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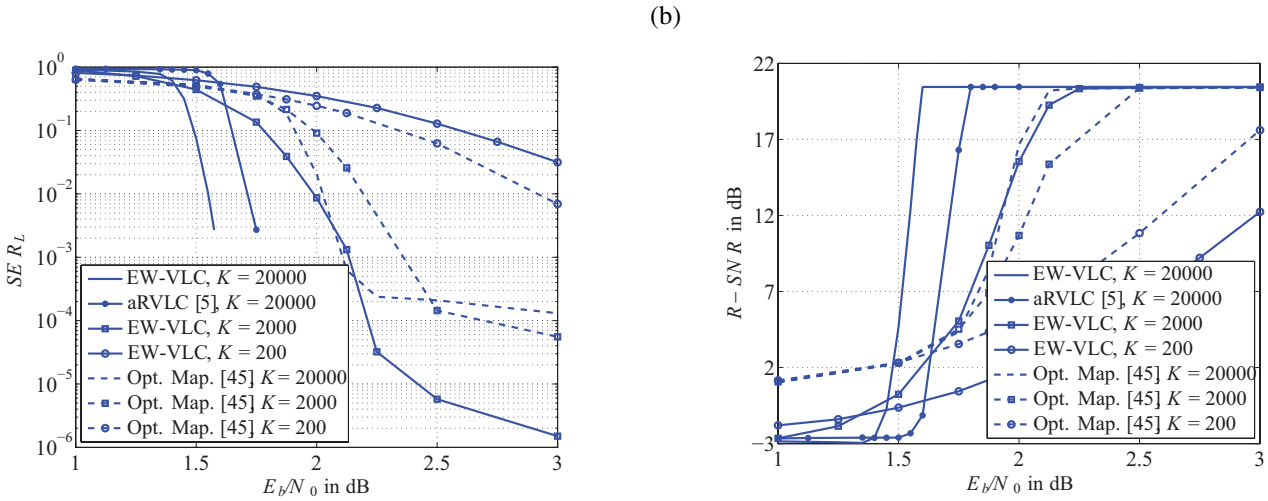

Fig. 3. Comparison between optimized mappings [45] and EW-VLCs for the AWGN channel with iterative source-channel decoding for varying block lengths $K \in \{200, 2000, 20000\}$. (a): symbol error rate (SER) versus E_b/N_0; (b): signal-to-noise ratio (SNR) versus E_b/N_0 after source reconstruction.

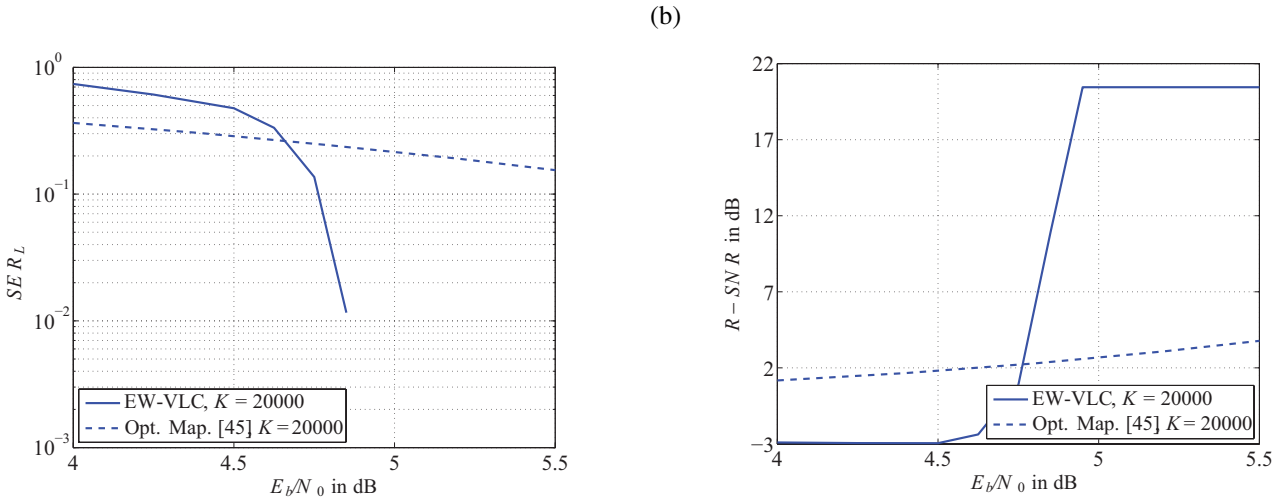

Fig. 4. Comparison between optimized mappings [45] and EW-VLCs for the AWGN channel with iterative source-channel decoding for an uncorrelated source and a block length $K = 20000$. (a): symbol error rate (SER) versus E_b/N_0; (b): signal-to-noise ratio (SNR) versus E_b/N_0 after source reconstruction.

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