ITERATIVE SOURCE CODED MODULATION: TURBO ERROR CONCEALMENT BY ITERATIVE DEMODULATION

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ABSTRACT

In this paper we propose to combine the exploitation of residual source redundancy and iterative demodulation of multilevel modulations in a Turbo-like process to improve error concealment for the source codec parameters. The developed system, denoted as *iterative source coded modulation* (ISCM), uses the band-width efficiently since no channel code increases the transmitted bit rate. Thus, ISCM is suitable for communication systems without channel coding, but it can be enhanced by a channel code to resemble a Multiple Turbo Code. Furthermore, without a complex channel decoder ISCM requires relatively low computational complexity. Nonetheless, simulations demonstrate considerable performance gains by ISCM with few iterations and small interleaver sizes. Additionally, measures are derived to ensure a correct relation between the unequal error protection in multilevel modulation and the different significance of bits of source codec parameters.

1. INTRODUCTION

In digital communications speech, audio and video signals source encoding is applied by extracting parameters such as scale factors or predictor coefficients. These source codec parameters may be independent of each other, but each parameter exhibits usually considerable residual redundancy in terms of, e.g., non-uniform probability distribution or auto-correlation. With the *a priori* knowledge on this residual redundancy the error robustness can be enhanced at the receiver, e.g., by error concealment using *softbit source decoding* (SBSD) [1]. The basic idea of SBSD is to estimate the codec parameters within the source decoding process by taking *a priori* knowledge (residual redundancy) and reliability of the received bits into account.

The capabilities of SBSD have been extended by *iterative source-channel decoding* (ISCD) [2], [3], a Turbo-like algorithm. However, ISCD usually requires additional bit rate (and band-width) on the channel due to channel coding, which furthermore requires computational complex decoding algorithms. In [4] an ISCD system without demand for additional bit rate is proposed.

For efficient use of the limited band-width multilevel modulation is used in modern communication systems, e.g., up to 64QAM in IEEE 802.11a WLAN or 8PSK in EDGE. Usually, bit-wise interleaving is applied to cope the effects of flat fading on the channel. In presence of a channel code such a scheme can be considered as *bit-interleaved coded modulation* (BICM) [5], [6]. Also the capabilities of BICM can be extended by performing demodulation and channel decoding in a Turbo-like algorithm. This is called BICM with *iterative decoding* (BICM-ID) [7], [8].

In this paper we propose the enhancement of the error concealment capabilities of SBSD by using the residual redundancy of the source with SBSD in a Turbo-like process together with *iterative demodulation* of multilevel modulation. We will denote the emerging system as *iterative source coded modulation* (ISCM). Since ISCM performs error concealment with a Turbo-like algorithm this scheme can also be considered as *Turbo error concealment by iterative demodulation*. Without a channel code ISCM uses a simple soft-demodulator instead of a complex channel decoder and does not require additional bit-rate on the channel. Simulations demonstrate considerable performance gains by ISCM with respect to conventional SBSD combined with multilevel modulation. ISCM can be applied to communications systems without channel coding, e.g., the European standard DECT (Digital Enhanced Cordless Telecommunications). Furthermore, by adding channel coding to ISCM the emerging system resembles a Multiple Turbo code.

2. THE ISCM SYSTEM

In Fig. 1 the baseband model of the proposed ISCM system is depicted. At time instant τ, a source encoder determines a frame $\underline{u}_\tau$ of K_S source codec parameters $u_{\kappa,\tau}$ with $\kappa=1,\ldots K_S$ denoting the position in the frame. The single elements $u_{\kappa,\tau}$ of $\underline{u}_\tau$ are assumed to be statistically independent from each other. The rate of source codec parameters $u_{\kappa,\tau}$ is denoted by R_S. Each value $u_{\kappa,\tau}$ is individually mapped to a quantizer reproduction level $\bar{u}_\kappa^{(\xi)}$, $\xi=1,\ldots 2^{M_\kappa}$. To each quantizer reproduction level $\bar{u}_\kappa^{(\xi)}$ selected at time instant τ a unique bit pattern $\mathbf{x}_{\kappa,\tau}$ of M_κ bits is assigned. For simplicity we assume $M_\kappa=M$ for all κ. The single bits of a bit pattern $\mathbf{x}_{\kappa,\tau}$ are indicated by $x_{\kappa,\tau}^{(m)}$, $m=1,\ldots M$. The frame of bit patterns at time instant τ is denoted by $\underline{\mathbf{x}}_\tau$.

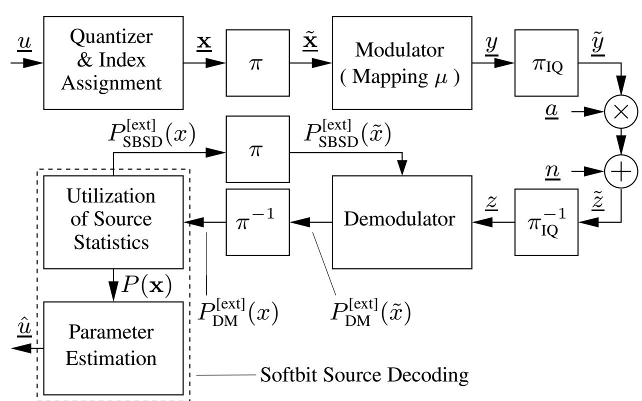

Fig. 1. Baseband model of the ISCM system

Since the used multilevel modulation will introduce an unequal error protection (UEP) performance the different significance of

Table 1. Average Noise Energy $\overline{N}_E^{(m)}$ of a single bit error for a Lloyd-Max quantizer with $M = 3$ bit/parameter

Index Assignment	$\overline{N}_E^{(1)}$	$\overline{N}_E^{(2)}$	$\overline{N}_E^{(3)}$
Natural Binary (NB)	5.03	1.51	0.38
SNR optimized (SO)	4.66	5.04	4.66

the single bits $x_{k,\tau}^{(m)}$ is required to make an appropriate assignment of these bits to the modulated symbols. As an indicator for the significance of a bit serves the average (w.r.t. the probability of occurence $P(\bar{u}^{(\xi)})$) noise energy $\overline{N}_E^{(m)}$ which a single bit error of bit $x^{(m)}$ will generate for a parameter,

$$\overline{N}_E^{(m)} = \sum_{\xi=1}^{2^M} \left(\bar{u}^{(\xi)} - \overline{\bar{u}^{(\xi)}}\right)^2 \cdot P\left(\bar{u}^{(\xi)}\right) \quad . \tag{1}$$

$\overline{\bar{u}^{(\xi)}}$ is the quantizer level with same assigned bit pattern $\mathbf{x}$, except for an inverted bit $x^{(m)}$. A high $\overline{N}_E^{(m)}$ implies a high significance of bit m. Table 1 shows values of $\overline{N}_E^{(m)}$ for Lloyd-Max quantization of a Gaussian source ($\sigma_u^2 = 1$) with $M = 3$ bit/parameter and two index assignments, natural binary (NB) and SNR optimized (SO). The latter one was develop in [9] to improve ISCD. As visible for NB index assignment the bit $m = 1$ is by far the most significant bit (MSB) while for SO index assignment the bit $m = 2$ has the highest significance. The relation between quantizer levels and bit patterns is depicted in Fig. 2.

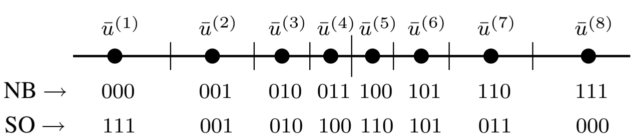

Fig. 2. Index Assignments to Quantizer Levels $\bar{u}^{(\xi)}$ for $M = 3$ bit/parameter, Natural Binary (NB) and SNR optimized (SO) [9]

After the source encoder the bit patterns $\underline{\mathbf{x}}_\tau$ are permuted by a pseudo-random block interleaver π to $\underline{\tilde{\mathbf{x}}}_\tau$. Similar to the BICM(-ID) systems in [5], [7] the interleaver π consists of I separate different sub-interleavers, where I is the number of bits that will be mapped to one channel symbol later on, e.g., $I = 3$ in case of 8PSK (phase shift keying). Thus, the set $\underline{\tilde{\mathbf{x}}}_\tau$ consists of bit patterns $\tilde{\mathbf{x}}_{k,\tau}$, $k = 1, \dots K_C$, with I single bits $x_{k,\tau}^{(i)}$, $i = 1, \dots I$. For a system in which the rate of channel symbols R_C is identical to the rate of source codec parameters R_S, i.e., $R_C = R_S$, we set $I = M$ and $K_C = K_S$. Furthermore, the interleaver π reassigns the bit positions m of $\underline{\mathbf{x}}_\tau$ to the bit positions i of $\underline{\tilde{\mathbf{x}}}_\tau$ according to their significance. As described in Section 3.1 the *signal constellation sets* of the modulator are designed such that the bit at position $i = 1$ is the most protected bit (MPB) and the bit at position $i = I$ the least protected bit (LPB). Note that with $I = M$ the relation between i and m is identical for all parameters. Thus, a common reassignment function ψ can be used. Since the index i is already sorted according to its inherent bit protection, the complete reassignment process can be described by

$$(i_{\text{MPB}}, \dots i_{\text{LPB}}) = \psi(m_{\text{MSB}}, \dots m_{\text{LSB}}) \quad , \tag{2}$$

e.g., for the $M = 3$ bit/parameter SO index assignment and $I = 3$ bit/channel symbol we obtain $\psi^{\text{SO}} = (2, 1, 3)$ or $\psi^{\text{SO}} = (2, 3, 1)$. By $\psi^{\text{SO}} = (2, 1, 3)$ we indicate that the MSB of SO index assignment $m_{\text{MSB}} = 2$ is reassigned to the MPB $i_{\text{MPB}} = 1$. The next significant bit $m = 1$ is reassigned to the next best protected bit $i = 2$ and the LSB $m_{\text{LSB}} = 3$ to the LPB $i_{\text{MPB}} = 3$. In case of $I \neq M$ a more complex reassignment function may be required.

The modulator maps an interleaved bit pattern $\tilde{\mathbf{x}}_{k,\tau}$ according to a mapping rule μ to a complex modulated symbol $y_{k,\tau}$ out of the *signal constellation set* $\mathcal{Y}$

$$y_{k,\tau} = \mu(\tilde{\mathbf{x}}_{k,\tau}) \quad . \tag{3}$$

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

$$\tilde{\mathbf{x}}_{k,\tau} = \mu^{-1}(y_{k,\tau}) = [\mu^{-1}(y_{k,\tau})^{(1)}, \dots \mu^{-1}(y_{k,\tau})^{(I)}] \quad . \tag{4}$$

The modulated symbols are normalized to an average energy of $E\{\|y_{k,\tau}\|^2\} = 1$.

In case of Rayleigh fading an IQ interleaver π_{IQ} [8] is applied to the modulated symbols. With an IQ interleaver the in-phase (I) and quadrature (Q) component of a modulated symbol shall be made to fade independently. For a memoryless fading channel this can be achieved by a simple wraparound shift by one symbol of the Q component of the modulated symbol,

$$\tilde{y}_{k,\tau} = \begin{cases} \text{Re}\{y_{k,\tau}\} + j \cdot \text{Im}\{y_{k-1,\tau}\} & 2 \leq k \leq K_C \\ \text{Re}\{y_{k,\tau}\} + j \cdot \text{Im}\{y_{K_C,\tau}\} & k = 1 \end{cases} \quad . \tag{5}$$

On the channel the transmitted symbols $y_{k,\tau}$ are faded by the real-valued Rayleigh distributed coefficients $a_{k,\tau}$ with $E\{|a_{k,\tau}|^2\} = 1$. In this paper we assume that these coefficients are known at the receiver, i.e., perfect knowledge of the channel state information (CSI). Thus, $a_{k,\tau}$ can be real-valued since a possible phase of a complex fading coefficient could be perfectly corrected at the receiver. Next, complex additive white Gaussian noise (AWGN) $n_{k,\tau} = n'_{k,\tau} + jn''_{k,\tau}$ with a known power spectral density of $\sigma_n^2 = N_0$ ($\sigma_{n'}^2 = \sigma_{n''}^2 = N_0/2$) is applied. Thus, the received channel symbols $\tilde{z}_{k,\tau}$ can be written as

$$\tilde{z}_{k,\tau} = a_{k,\tau} \cdot \tilde{y}_{k,\tau} + n_{k,\tau} \quad . \tag{6}$$

3. THE PROPOSED ISCM RECEIVER

In this section the proposed ISCM receiver design is derived. After the IQ deinterleaver π_{IQ}^{-1} is applied the received symbols $z_{k,\tau}$ are evaluated in a Turbo process, which exchanges *extrinsic* probabilities between the demodulator and the softbit source decoder.

3.1. Extrinsic Probabilities Computation in Demodulator

The demodulator (DM) computes *extrinsic* probabilities $P_{\text{DM}}^{[\text{ext}]}(\tilde{x})$ for each bit $\tilde{x}_{k,\tau}^{(i)}$ being $b = 0, 1$ according to [7], [8]

$$P_{\text{DM}}^{[\text{ext}]}(\tilde{x}_{k,\tau}^{(i)} = b) \sim \sum_{\hat{y} \in \mathcal{Y}_b^i} P(z_{k,\tau}|\hat{y}) \prod_{j=1, j \neq i}^{I} P_{\text{SBSD}}^{[\text{ext}]}\left(\tilde{x}_{k,\tau}^{(j)} = \mu^{-1}(\hat{y})^{(j)}\right). \tag{7}$$

Each $P_{\text{DM}}^{[\text{ext}]}(\tilde{x})$ consists of the sum over all possible channel symbols $\hat{y}$ for which the i^{th} bit of the corresponding bit pattern $\tilde{\mathbf{x}} = \mu^{-1}(\hat{y})$ is b. 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 case of IQ interleaving the I and Q component of $\underline{z}_{k,\tau}$ are faded independently. Thus, the conditional probability density $P(z_{k,\tau}|\hat{y})$ is given by

$$P(z_{k,\tau}|\hat{y}) = \frac{1}{\pi\sigma_n^2} \exp\left(-\frac{d^2}{\sigma_n^2}\right) \quad \text{, with} \tag{8}$$

$$d^2 = |\text{Re}\{z_{k,\tau}\} - a_{k,\tau}\text{Re}\{\hat{y}\}|^2 + |\text{Im}\{z_{k,\tau}\} - a_{k-1,\tau}\text{Im}\{\hat{y}\}|^2 \tag{9}$$

Without IQ interleaving (9) simplifies to $d^2 = \|z_{k,\tau} - a_{k,\tau}\hat{y}\|^2$. In the first iteration the $P_{\text{SBSD}}^{[\text{ext}]}(\tilde{x})$ are set as equiprobable.

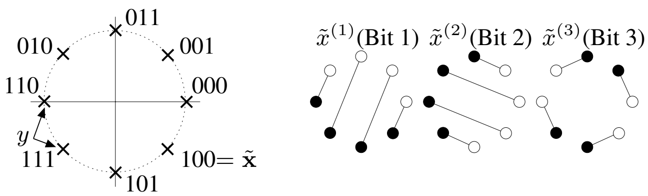

Fig. 3. 8PSK-Gray mapping with signal-pair distances $\|y-\check{y}\|$ for $\check{d}_h^{2(i)}(\mu)$. $\circ \leftrightarrow \tilde{x}^{(i)}=0$ and $\bullet \leftrightarrow \tilde{x}^{(i)}=1$.

Based on the performance bound of the bit-error rate (BER) P_b of BICM-ID on Rayleigh fading channels [7] we derive a measure for the protection level of a bit position i. Assuming error free feedback (EFF), i.e., $P_{\text{SBSD}}^{\text{[ext]}}(\tilde{x}) \in \{0.0, 1.0\}$, the asymptotic behavior of P_b in BICM-ID can be described by

$$\log_{10} P_b \approx \frac{-d_H(\mathcal{C})}{10}\left[\left(R\cdot\check{d}_h^2(\mu)\right)_{dB}+\left(\frac{E_b}{N_0}\right)_{dB}\right]+const. \quad (10)$$

In BICM-ID $d_H(\mathcal{C})$, the minimum Hamming distance of the channel code used in BICM-ID, defines the slope of the bound. $\check{d}_h^2(\mu)$ is the *harmonic mean* of the *squared Euclidean distances* of the two symbols in $\mathcal{Y}$ which do not eliminate when applying (7) for $b=0,1$ and EFF. Assuming EFF the decision is made between these two symbols. For a mapping μ with I bits the *harmonic mean* $\check{d}_h^2(\mu)$ is given by

$$\check{d}_h^2(\mu)=\left(\frac{1}{I2^I}\sum_{i=1}^{I}\sum_{b=0}^{1}\sum_{y\in\mathcal{Y}_b^i}\frac{1}{\|y-\check{y}\|^2}\right)^{-1}, \quad (11)$$

where $\check{y}$ possesses the identical bit pattern $\mathbf{x}$ as y except for an inverted bit at position i. The product of the information rate R and the *harmonic mean* $\check{d}_h^2(\mu)$ determines an E_b/N_0 offset of the bound, i.e., a horizontal offset in a BER vs. E_b/N_0 plot.

Figs. 3 and 4 depict the *signal constellation sets* and bit pattern assignment for two mappings with $I=3$ bits, 8PSK-Gray mapping and 8PSK-SSP (semi Set-Partitioning) mapping [7]. On the right side of Figs. 3 and 4 the *Euclidean distances* $\|y-\check{y}\|$ used in (11) are depicted. Black dots denote $\tilde{x}^{(i)}=\mu^{-1}(y)^{(i)}=1$ and white dots $\tilde{x}^{(i)}=\mu^{-1}(y)^{(i)}=0$. 8PSK-Gray mapping performs best in the non-iterative BICM case [5] due to its Gray bit pattern assignment. In the iterative case of BICM-ID 8PSK-SSP mapping has the best asymptotic performance [7] since it possesses the highest $\check{d}_h^2(\mu)$ of all regular 8PSK mappings. Mappings with non-regular signal constellation sets providing superior asymptotic performance for BICM-ID are presented in [10], but in this paper we limit the discussion to regular mappings.

However, in ISCM there exists no channel code with a $d_H(\mathcal{C})$. Thus, (10) cannot be used in the context of ISCM. But, we still have a decoder supplying *extrinsic* feedback probabilities, which can asymptotically resemble EFF. Thus, a high $\check{d}_h^2(\mu)$ is still an indicator for a good asymptotic performance of a mapping μ in ISCM. Moreover, since no smearing effect of a convolutional code is present in ISCM each of the I bits of a mapping can be considered separately and $\check{d}_h^{2(i)}$, the *harmonic mean* of the *squared Euclidean distances* $\|y-\check{y}\|$ for bit i, is a measure for the error protection of bit i. The highest $\check{d}_h^{2(i)}$ identifies the most protected bit (MPB), resp. the lowest $\check{d}_h^{2(i)}$ the least protected bit (LPB). Similar to (11) $\check{d}_h^{2(i)}$ can be determined by

$$\check{d}_h^{2(i)}(\mu)=\left(\frac{1}{2^I}\sum_{b=0}^{1}\sum_{y\in\mathcal{Y}_b^i}\frac{1}{\|y-\check{y}\|^2}\right)^{-1}. \quad (12)$$

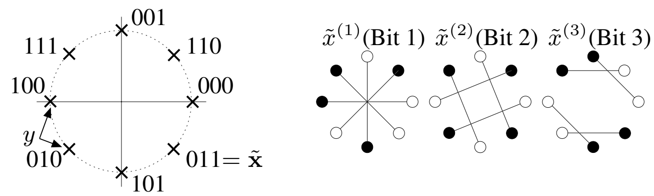

Fig. 4. 8PSK-SSP mapping with signal-pair distances $\|y-\check{y}\|$ for $\check{d}_h^{2(i)}(\mu)$. $\circ \leftrightarrow \tilde{x}^{(i)}=0$ and $\bullet \leftrightarrow \tilde{x}^{(i)}=1$.

Table 2 lists the $\check{d}_h^2$ and $\check{d}_h^{2(i)}$ for 8PSK-Gray and 8PSK-SSP mapping. As visible in Figs. 3 and 4 and stated before, the bit pattern assignment of all mappings is arranged such that $i=1$ is the MPB and $i=I$ the LPB.

Table 2. $\check{d}_h^2$ and $\check{d}_h^{2(i)}$ for 8PSK-Gray and 8PSK-SSP mapping

Mapping μ	$\check{d}_h^2$	$\check{d}_h^{2(1)}$	$\check{d}_h^{2(2)}$	$\check{d}_h^{2(3)}$
8PSK-Gray	0.81	1	1	0.59
8PSK-SSP	2.88	4	3.41	2

3.2. Softbit Source Decoding

After deinterleaving the $P_{\text{DM}}^{\text{[ext]}}(x)$ are fed into the SBSD. The algorithm how to compute the $P_{\text{SBSD}}^{\text{[ext]}}(x)$ of SBSD is based on a fully connected Trellis structure for the parameters and has been derived in [2], [3]. It shall briefly reviewed next:

- Firstly, the branch transition probabilities $\gamma(\mathbf{x}_{\kappa,\tau})$ are computed for all 2^M permutations of a bit pattern $\mathbf{x}_{\kappa,\tau}$ by
$$\gamma(\mathbf{x}_{\kappa,\tau})=\prod_{m=1}^{M}P_{\text{DM}}^{\text{[ext]}}(x_{\kappa,\tau}^{(m)}) \quad . \quad (13)$$
Furthermore, the *extrinsic* branch transition probabilities are required. By extracting the bit under consideration from the present bit pattern $\mathbf{x}_{\kappa,\tau}=\{x_{\kappa,\tau}^{(m)},\mathbf{x}_{\kappa,\tau}^{\text{[ext},m]}\}$ we obtain
$$\gamma(\mathbf{x}_{\kappa,\tau}^{\text{[ext},m]})=\prod_{j=1,j\neq m}^{M}P_{\text{DM}}^{\text{[ext]}}(x_{\kappa,\tau}^{(j)}) \quad . \quad (14)$$
- Secondly, if the codec parameters exhibit a 1st order Markov property the *a priori* knowledge $P(\mathbf{x}_{\kappa,\tau}|\mathbf{x}_{\kappa,\tau-1})$ is exploited by a forward recursion with $\alpha(\mathbf{x}_{\kappa,0})=P(\mathbf{x}_{\kappa,0})$ and
$$\alpha(\mathbf{x}_{\kappa,\tau})=\sum_{\mathbf{x}_{\kappa,\tau-1}}P(\mathbf{x}_{\kappa,\tau}|\mathbf{x}_{\kappa,\tau-1})\cdot\gamma(\mathbf{x}_{\kappa,\tau})\cdot\alpha(\mathbf{x}_{\kappa,\tau-1}). \quad (15)$$
The summation runs over all $\mathbf{x}_{\kappa,\tau-1}$.
- Finally, the forward metrics $\alpha(\mathbf{x}_{\kappa,\tau-1})$ of the previous time instance $\tau-1$ are combined with the *a priori* knowledge $P(\mathbf{x}_{\kappa,\tau}|\mathbf{x}_{\kappa,\tau-1})$ and the present $\gamma(\mathbf{x}_{\kappa,\tau}^{\text{[ext},m]})$ to
$$P_{\text{SBSD}}^{\text{[ext]}}(x_{\kappa,\tau}^{(m)}=b)= \quad (16)$$
$$\sum_{\mathbf{x}_{\kappa,\tau}^{\text{[ext},m]}}\gamma(\mathbf{x}_{\kappa,\tau}^{\text{[ext},m]})\sum_{\mathbf{x}_{\kappa,\tau-1}}P(\mathbf{x}_{\kappa,\tau}^{\text{[ext},m]}|\mathbf{x}_{\kappa,\tau-1},x_{\kappa,\tau}^{(m)}=b)\cdot\alpha(\mathbf{x}_{\kappa,\tau-1}).$$
These $P_{\text{SBSD}}^{\text{[ext]}}(x)$ are fed back to the demodulator.

In the final iteration parameter-oriented *a posteriori* knowledge $P(\mathbf{x}_{\kappa,\tau}|\underline{z}_{[1,\tau]})\triangleq P(\mathbf{x}_{\kappa,\tau}|\underline{z}_1,\ldots\underline{z}_\tau)$ is determined by

$$P(\mathbf{x}_{\kappa,\tau}|\underline{z}_{[1,\tau]})=C\cdot\gamma(\mathbf{x}_{\kappa,\tau})\sum_{\mathbf{x}_{\kappa,\tau-1}}P(\mathbf{x}_{\kappa,\tau}|\mathbf{x}_{\kappa,\tau-1})\cdot\alpha(\mathbf{x}_{\kappa,\tau-1}). \quad (17)$$

The constant factor C ensures that the *total probability theorem* is fulfilled. Thus, if the *minimum mean squared error* (MMSE) serves as fidelity criterion, the individual estimates are given by [1]

$$\hat{u}_{\kappa,\tau}=\sum_{\xi}\bar{u}_\kappa^{(\xi)}\cdot P(\mathbf{x}_{\kappa,\tau}\hat{=}\xi|\underline{z}_{[1,\tau]}) \quad . \quad (18)$$

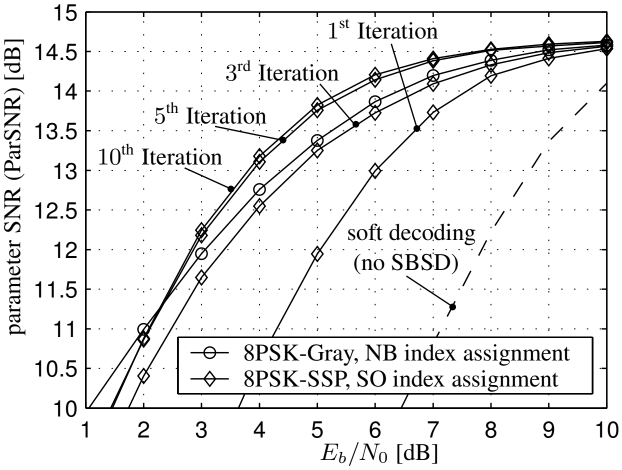

Fig. 5. Parameter SNR (ParSNR) for an AWGN channel

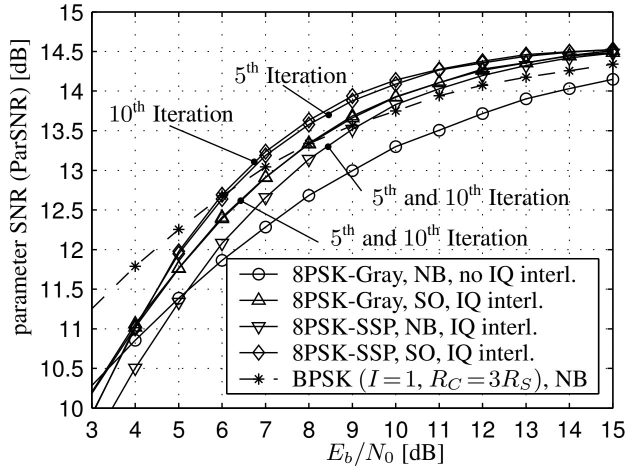

Fig. 6. Parameter SNR (ParSNR) for a Rayleigh channel

4. SIMULATION RESULTS

The capabilities of the proposed ISCM scheme shall be demonstrated by simulation. Instead of using any specific speech, audio, or video encoder, we model $K_S = 6$ statistically independent source codec parameters u by 1^{st} order Gauss-Markov process with auto-correlation $\rho = 0.95$, a typical value, e.g., for the scale factors of audio transform codecs. Each parameter $u_{\kappa,\tau}$ is scalarly quantized by a Lloyd-Max quantizer using $M = 3$ bits/parameter. The index assignment is either NB or SO with $\psi^{\text{NB}} = (1, 2, 3)$ and $\psi^{\text{SO}} = (2, 1, 3)$. $I = 3$ bits are modulated to one channel symbol. The *parameter signal-to-noise ratio* (ParSNR) between the originally generated parameters $u_{\kappa,\tau}$ and the reconstructed estimates $\hat{u}_{\kappa,\tau}$ is used for quality evaluation.

Fig. 5 shows results for an AWGN channel without Rayleigh fading ($a_{k,\tau} = 1$), i.e., no IQ interleaving is required. The solid curve marked "○" depicts the result of the non-iterative baseline system with NB index assignment, 8PSK-Gray mapping and a single iteration using SBSD. The performance of this baseline system cannot be noticeably improved by iterations. The solid lines marked "◊" represent results for the proposed ISCM scheme with SO index assignment and 8PSK-SSP mapping. A gain of up to $\Delta_{E_b/N_0} \approx 1$ dB can be achieved in the interesting ParSNR regions. The maximum gain is almost reached after 5 iterations. When non-iterative soft decoding by simple parameter estimation based on the $P_{\text{DM}}^{[\text{ext}]}(x)$ is used, i.e., not utilizing SBSD to exploit the *a priori* knowledge on the residual redundancy of the parameters, only the performance of the dashed curve can be obtained.

In Fig. 6 results for a memoryless Rayleigh channel are depicted. The curve marked "○" represents a similar baseline system as in Fig. 5. The performance of an ISCM system with all proposed improvements, i.e., 8PSK-SSP mapping, SO index assignment and IQ interleaving, is given by the solid curves marked "◊". Above a ParSNR of 13.3 dB the obtainable gain exceeds $\Delta_{E_b/N_0} = 3$ dB. Even a baseline system with BPSK modulation (dashed curve) is outperformed by more than $\Delta_{E_b/N_0} = 1$ dB. Note that this BPSK system requires a three times higher rate of channel symbols, $R_C^{\text{BPSK}} = 3 \cdot R_C^{\text{8PSK}}$! Comparing the curves "△" and "▽" with "◊" reveals that the two proposed modifications SO index assignment and 8PSK-SSP mapping are both required to obtain the best performance, but each alone already provides significant gains.

Other, not depicted, simulations showed that larger interleavers, i.e., an increased K_S, do not result in further performance improvements. Of course, in case the fading is not uncorrelated, the IQ interleaver size has to be modified such that the I and Q components are faded independently.

5. CONCLUSION

In this paper we improved error concealment by *softbit source decoding* by integrating it in a Turbo-like system with a low-complexity soft-demodulator as the second "decoder". Despite the lack of channel coding the proposed system, denoted *iterative source coded modulation*, achieves significant performance gains, e.g., up to and beyond 3 dB for a Rayleigh channel, already with few iterations and small interleaver sizes. An additional benefit of the presented ISCM system, a Turbo-like system without a channel code, is that the rate of transmitted symbols, and consequently the required band-width, is not increased as it is the case with channel coding. Of course, if the available band-width and the computational budget allow channel coding ISCM could be combined with, e.g., ISCD to resemble a Multiple Turbo code.

6. REFERENCES

[1] T. Fingscheidt and P. Vary, "Softbit Speech Decoding: A New Approach to Error Concealment," *IEEE Trans. Speech Audio Processing*, pp. 240–251, Mar. 2001.

[2] M. Adrat, P. Vary, and J. Spittka, "Iterative Source-Channel Decoder Using Extrinsic Information from Softbit-Source Decoding," in *Proc. of IEEE ICASSP*, Salt Lake City, Utah, May 2001.

[3] N. Görtz, "On the Iterative Approximation of Optimal Joint Source-Channel Decoding," *IEEE J. Select. Areas Commun.*, pp. 1662–1670, Sept. 2001.

[4] M. Adrat and P. Vary, "Turbo Error Concealment of Mutually Independent Source Codec Parameters," in *Proc. of ITG SCC 2004*, Erlangen, Germany, Jan. 2004.

[5] E. Zehavi, "8-PSK Trellis Codes for a Rayleigh Channel," *IEEE Trans. Comm.*, pp. 873–884, May 1992.

[6] G. Caire, G. Taricco, and E. Biglieri, "Bit-Interleaved Coded Modulation," *IEEE J. Select. Areas Commun.*, pp. 927–946, May 1998.

[7] X. Li, A. Chindapol, and J. A. Ritcey, "Bit-Interleaved Coded Modulation With Iterative Decoding and 8PSK Signaling," *IEEE Trans. Comm.*, pp. 1250–1257, Aug. 2002.

[8] A. Chindapol and J. A. Ritcey, "Design, Analysis, and Performance Evaluation for BICM-ID with Square QAM Constellations in Rayleigh Fading Channels," *IEEE J. Select. Areas Commun.*, pp. 944–957, May 2001.

[9] J. Hagenauer and N. Görtz, "The Turbo Principle in Joint Source-Channel Coding," in *Proc. of IEEE ITW 2003*, Paris, France, Mar. 2003, pp. 275–278.

[10] T. Clevorn and P. Vary, "Iterative Decoding of BICM with Non-Regular Signal Constellation Sets," in *Proc. of ITG SCC 2004*, Erlangen, Germany, Jan. 2004.