Abstract—For evolution of the GSM/EDGE radio access network (RAN), the use of higher order modulation like 16- and 32-ary quadrature amplitude modulation (QAM) is considered in standardization for increased peak data rates and reduced transmission delays. In this paper, an optimized receiver design for different packet data transmission schemes is proposed. Turbo coding and turbo equalization is discussed for improved power efficiency and interference robustness. An efficient complexity reduction of the equalizer enables the usage of turbo equalization at a complexity comparable to that of separate equalization and decoding for turbo-coded transmission.

I. INTRODUCTION

In order to increase the peak data rates and to reduce transmission delays in the GSM/EDGE system, higher order modulation like 16- and 32-ary quadrature amplitude modulation (QAM) and dual symbol rate (DSR) [1] are currently discussed for standardization [2]. In this paper, we focus on higher order QAM modulation for GSM/EDGE, but the results can be extended also to DSR in a straightforward way. Furthermore, an extension of channel coding in GSM/EDGE1 is discussed currently based on turbo coding of UMTS Terrestrial Radio Access Network (UTRAN) [3].

A performance analysis of turbo-coded transmission with 16QAM can be found in [4], where in addition the influence of transmission impairments on the performance as well as the gain in terms of network throughput is shown. In this paper here, we investigate turbo equalization [5] applied to the conventional convolutionally coded transmission schemes as competitor for the turbo-coded transmission schemes, and compare performance of different reduced-complexity (inner) equalizers2 for the 16QAM and 32QAM packet data transmission schemes. An advantage of the turbo equalization approach is, that the conventional coding format can be preserved, so that even conventional receivers (separate equalization and decoding) are applicable in the system. Optionally, the receiver may apply a varying number of equalization and decoding iterations, depending on the signal quality (e.g. signal-to-interference plus-noise ratio (SINR)).

Several turbo equalization schemes for time division multiple access (TDMA) systems based on PSK modulation can be found in the literature, e.g. [7]–[14]. For turbo equalization, the (inner) equalizer receives extrinsic a priori information input from the outer channel decoder and provides extrinsic a posteriori information output to the decoder. The decoder processes the input from the equalizer and produces extrinsic a posteriori information on the coded bits, so that equalization performance improves from iteration to iteration. As (inner) equalizer, reduced-complexity variants of the BCJR algorithm [15]–[17] are selected, that are based on joint reduced-state sequence estimation (JRSE) with Ungerboeck set partitioning [18].

We show, that a large complexity reduction (compared to the full-state equalizer) is feasible, similar to 8PSK modulation [19]. Furthermore, a simplified minimum mean-squared error (MMSE) soft-output detector [7]–[10] with soft-cancellation of pre- and postcursor intersymbol interference (ISI) extended to higher order modulations can be applied for subsequent turbo equalization iterations. For some of the considered transmission schemes, extrinsic information transfer (EXIT) chart [20] convergence analyses are performed.

The paper is structured as follows. The system model is introduced in Section II, and the receiver equalization algorithm is presented in Section III. In Section IV, performance is analyzed and simulation results are given for all currently considered GSM/EDGE modulation and coding schemes (MCs).

II. SYSTEM MODEL

The system model in equivalent discrete-time complex baseband representation is shown in Fig. 1. After encoding of the source information bits \( d[k] \) and successive interleaving, the encoded bits \( d[k] \) are mapped to linear modulation symbols \( a[k] \) and 4 bursts of \( N_f = 120 \) symbols each are transmitted over the channel with impulse response \( h[k] \) of order \( q_a \), which includes transmit pulse shaping and receive filtering (square-root raised cosine filter, roll-off factor \( \alpha = 0.3 \) [19]). As ideal frequency hopping is assumed, uncorrelated channel realizations are present for different bursts (block fading channel). The receive signal \( r[k] \) is disturbed by additive white Gaussian noise (AWGN) \( n[k] \) with variance \( \sigma_n^2 \). The MMSE decision-feedback equalization (DFE) prefilter of [21] with impulse response \( f[k] \) is employed in front of reduced-complexity BCJR equalization, whereas for soft-output detection with soft ISI cancellation the matched filter is used. Turbo equalization is based on sequence \( u[k] \) and delivers estimates

1 Convolutional codes of constraint length \( K = 7 \) are employed for GSM/EDGE modulation and coding schemes (MCs) with 8-ary phase-shift keying (PSK) modulation.

2 As the overall equalization process is comparable to the decoding of serially concatenated convolutional codes (SCCCs) [6] (regarding the ISI channel as inner code), we refer to the equalizer as the inner component of the turbo receiver [5].
of the transmit symbols \( \hat{a}[k] \) and the transmitted information bits \( d_k[k] \).

A. Coding and Interleaving

1) Turbo Coding: In [2], the usage of the UMTS UTRAN turbo code [3] is proposed, which is also adopted here with according internal interleaver and rate matching.

2) Convolutional Coding: For all transmission schemes with convolutional coding, the rate 1/3 convolutional basis code (constraint length \( K = 7 \) of GSM/EDGE [22] with appropriate puncturing is used in order to optimize performance [5].

3) Interleaving: A block interleaver has been adopted for all transmission schemes without precoding (cf. Section II-B). Only for precoded transmission, the S–random interleaver [23] is used in order to optimize performance [5].

B. Precoding and Mapping

The adopted mapping for 32QAM is given in [2]. As no explicit mapping is specified for 16QAM in the standard so far, we adopt the Gray mapping shown in Fig. 2 with 4–tuples \( d^{(i)} = [d_{0}^i d_{1}^i d_{2}^i d_{3}^i]^{T} \) ((i) = 1...7): transposition). The modulation alphabet is denoted as \( A = \{ A_i | i = 0,...,N_A - 1 \} \) consisting of the \( N_A = 2^{4m} \) modulation symbols \( A_i \) with arbitrary numbering, where \( q_a \) is the number of bits carried by the transmit symbols. The bit sequence \( d[k] \) is mapped to the vector–valued sequence \( \hat{d}[k] \) with \( \hat{d}[k] = d[kq_a + j] \) for \( k \in K = \{ 0,...,N_S - 1 \} \), and \( j \in J = \{ 0,...,q_a - 1 \} \), where \( [·]_m \) denotes the mth element of a column vector, and finally the transmit symbol \( A_i \) for which \( d[k] = d^{(i)} \) is selected for \( a[k] \), if no precoding is applied.

A comprehensive study of turbo equalization is given e.g. in [5] (and references therein), where it is demonstrated, that EXIT charts [20] may be used in order to optimize (turbo equalization) detection performance for static frequency–selective channels with Gaussian disturbance by fitting the according equalizer and decoder characteristics.

As turbo equalization schemes are comparable to the decoding of serially concatenated convolutional codes (SCCCs), the inner code should have a recursive structure in order to optimize the error probability of the turbo receiver in the water–fall region [6]. This can be accomplished by symbol–based precoding before transmitting the mapped symbols [5]. The structure introduced by the precoder needs to be taken into account by the equalizer in the receiver. In order to avoid an increase of the equalizer complexity and to allow application along with reduced–state equalization algorithms like delayed decision–feedback sequence estimation (DDFSE) and reduced–state sequence estimation (RSSE), only certain precoding structures should be selected. Here, we consider only precoding with memory one [5]. The general precoder is defined by matrices \( A, B, C \) and \( D \), and has input vector \( d[k] \) with binary components. The memory of the precoder is represented by \( m[k] \). State transition and output expression are given by

\[
\begin{align*}
\hat{m}[k + 1] &= Am[k] + Bd[k], \\
\hat{d}[k] &= Cm[k] + Dd[k],
\end{align*}
\]

respectively, where all operations are carried out in Galois field GF(2). In [5], suitable low state precoders have been found with \( A = 1, B = [1 1 1 . . . 1]^T, C \in \{1 0 0 . . . 0\}^T, D = I_{q_a} \) (identity matrix of size \( q_a \times q_a \) for 8PSK modulation. For these parameters, the precoder has only two states and is represented by a scalar \( m[k] \) (initialized with zero). The precoder memory is updated by the parity of the input vector \( d[k] \), and the output vector \( \hat{d}[k] \) is determined from the input vector by adding \( C \) for \( m[k] = 1 \) and leaving \( \hat{d}[k] \) unchanged for \( m[k] = 0 \).

We adopt this precoder structure for 16QAM and use \( C = [1 0 0 0 0]^T \), as this choice modifies the equalizer EXIT chart characteristic only slightly and yields the closest fit of equalizer and decoder EXIT chart curves.

III. RECEIVER STRUCTURE – TURBO EQUALIZATION

A. Reduced–Complexity BCJR Equalization

Reduced–complexity variants of the BCJR algorithm [15]–[17] based on reduced–state sequence estimation (RSSE) with Ungerboeck set partitioning [18] are employed as (inner) component for turbo equalization in order to obtain a low–complexity equalizer with reduced number of (hyper)states. Only a low number of states compared to joint maximum–likelihood sequence estimation (JMLSE) may be allowed. A survivor map is created within the forward recursion, so that the selected states of all considered hyperstates of each detection step are identical for the forward and backward recursion. Prefiltering [21] is applied in order to optimize detection quality, and equalization is based on the feedback filter impulse response \( b[k] \) of order \( q_o \).

B. Soft–Output Detection with Soft ISI Cancellation

For reduced–complexity turbo equalization, a soft–output symbol detector combined with soft–cancellation of pre– and postcursor ISI via matched filtering [7]–[10] is a reasonable candidate. In order to reduce complexity, soft symbols and error variances are determined block–wise. After each iteration, the derived a posteriori (not extrinsic) information values on the code bits\(^{6}\) are considered for soft QAM symbol calculation according to \( \hat{a}[k] = E\{a[k]\} \) (\( E\{·\} \): expectation operator). Initialization is important for soft–output detection, so that the RSSE–based BCJR algorithm\(^{8}\) (discussed in the

4Different choices have been evaluated, but as the considered convolutional code has a high constraint length \( K = 7 \), only \( C = [1 0 0 0 0]^T \) has shown reasonable performance. Please note, that for the used mapping shown in Fig. 2 identical results are obtained for \( C = [0 1 0 1 0]^T \).

5Using the a posteriori information instead of the extrinsic information for soft symbol calculation has shown advantages in convergence speed of turbo iterations. Furthermore, the influence of residual interference, modeled as Gaussian distributed, becomes smaller in early iterations.

6An RSSE equalizer with 2–4 states was observed to be adequate for the initialization step of the soft–output detector with soft ISI cancellation.
preceeding subsection) is employed in the 0th iteration. For
prefiltering, the matched filter is applied, \( f[k] = h^*[−k] \) (\( \cdot^* \): conjugation), so that the overall channel and prefilter impulse response \( h_{tot}[k] = h[k] + h^*[−k] \) (\( \cdot^* \): convolution) is obtained (proportional to the noise autocorrelation function). In order to cancel post–and precursors before detection, we apply the cancellation filter

\[
h_a[k] = \begin{cases} 
0 & \text{for } k = 0 \\
h_{tot}[k] & \text{otherwise}
\end{cases}.
\]

A proper estimate of the variance of noise and residual ISI is important for the considered equalizer. Self–interference is estimated by determining the minimum offset of the soft symbols to the closest signal point of the underlying constellation given by \( \hat{\alpha}[k] = \arg \min_{\alpha \in A} |\hat{\alpha}[k] − \alpha|^2 \). Disturbance with respect to the estimated transmit symbols is estimated (lower bounded) by \( \sigma_t^2[k] = |\hat{\alpha}[k] − \alpha[k]|^2 \). Finally, with \( g_{tot}[k] = |h_a[k]|^2 \), the variance of overall residual disturbance (due to residual ISI and noise) on the receive signal after the matched filtering and soft ISI cancellation, \( v[k] = f[k] + r[k] − h_a[k] * \hat{\alpha}[k] \), is given by \( \sigma_v^2[k] = g_{tot}[k] * \sigma_t^2[k] + |h_{tot}[0]|^2 \cdot \sigma_n^2 \). The extrinsic and a posteriori log–likelihood ratios (LLRs) of the transmit bits are determined by [5]

\[
\text{LLR}[k, j] = \ln \frac{\text{Prob} \{ d[k], j = 0 \mid v[k] \}}{\text{Prob} \{ d[k], j = 1 \mid v[k] \}}, \quad k \in K, \ j \in J.
\]

The distribution of \( v[k] \) is assumed as complex Gaussian, \( v[k] \sim N \left( \mu_v[k], \sigma_v^2[k] \right) \), with expectation value \( \mu_v[k] = h_{tot}[0] \hat{\alpha}[k] \) and variance \( \sigma_v^2[k] = \sigma_t^2[k] \), which has shown satisfying accuracy along with the used approximations.

IV. SIMULATION RESULTS

For Monte–Carlo simulations of the proposed transmission schemes, the GSM/EDGE typical urban (TU) channel profile is used \((q_0 = 5)\). For the shown simulation results, perfect channel knowledge is assumed. Using the conventional training sequence based channel estimation of GSM/EDGE, performance typically degrades by 1.2–1.3 dB compared to ideal channel knowledge. In [14] it has been shown, that the inaccuracy of the estimated channel impulse response may be reduced by including channel re–estimation in the iterative equalization scheme.

Applying RSSE for equalization, the number of trellis states is given by \( \prod_{k=1}^{h} L_k \), where \( L_k \) represents the number of subsets for the \( k \)th channel tap \((16QAM: L_k \in \{1, 2, 4, 8, 16\}, 32QAM: L_k \in \{1, 2, 4, 8, 16, 32\})\). Partitionings are distinguished by the number of used subsets: \((L_1 \times L_2 \times \ldots)\). For delayed decision–feedback sequence estimation (DDFSE), as a special case of RSSE, e.g. \( L_k \in \{1, 16\} \) is used for 16QAM.

In Figs. 3–5, block error rate (BER) results are shown versus \( E_b/N_0 \) \((E_b: \text{average received bit energy}; N_0: \text{single–sided power spectral density of the underlying passband noise process})\) for 16QAM modulation using a rate \( R = 1/2 \) convolutional code with constraint length \( K = 7 \). In Fig. 3, performance of RSSE with 2, 4, and 16 states is compared for separate (0th iteration) and multiple iterations of equalization and decoding. It has to be pointed out, that the first iteration gives the highest improvements of about 2.7 dB at \( \text{BLER} = 10\% \) for all three cases. Therefore, high gains are

![Fig. 3. BLER vs. 10 log10(Eb/N0) for 16QAM, R = 1/2, TU channel, ideal channel knowledge, no precoding applied.](image)

![Fig. 4. BLER vs. 10 log10(Eb/N0) for 16QAM, R = 1/2, TU channel, ideal channel knowledge, RSSE (16) with and without precoding and soft–output detector with soft ISI canceler without precoding (0th iteration: RSSE (16)).](image)

![Fig. 5. BLER vs. 10 log10(Eb/N0) for 16QAM, RSSE (16), TU channel, ideal channel knowledge. Turbo equalization (TE) without precoding and separate equalization and decoding with convolutional code \((K = 7)\) (SED–CC) and turbo coding (SED–TC).](image)
feasible at approximately double complexity. Performance degradations due to state reduction are observed to get smaller for a larger number of iterations.

Performance of different turbo equalization schemes is given in Fig. 4. Here, the RSSE–based equalizer without precoder is compared to that with precoding and to the soft detector combined with soft ISI canceler. For precoding, only results for $C = [1 \ 0 \ 0 \ 0 \ 0]^T$ are shown, as other choices of C have revealed worse performance for the considered scenario. On the other hand, for additional precoding, the 0th iteration always gives worse performance compared to schemes without precoding, but on the other hand, higher gains per iteration are obtained, if coding and precoding (including mapping) are chosen suitably. For the considered convolutional code, a worse overall performance of transmission over the fading channel is observed for the scheme with precoding, which will be explained by the corresponding EXIT charts in the following. Additionally, performance results for the soft detector with soft ISI canceler are depicted in Fig. 4, where initialization is done using the RSSE–based equalizer. The soft detector delivers less gain in the first iteration of turbo equalization, but gives even slightly better performance than the RSSE–based equalizer after 6 iterations.

For all currently considered modulation and coding schemes (MCS) of GSM/EDGE [2] are given in Table I and II in terms of the required $10 \log_{10}(E_b/N_0)$ [dB] to achieve a BLER of 10% and a bit error rate (BER) of $10^{-3}$, respectively. Results are shown for separate equalization and decoding using convolutional coding (SED-CC), which is equivalent to the zeroth iteration of turbo equalization, for turbo equalization (TE) after the first and fourth iteration (convolutional coding, constraint length $K = 7$), and for turbo–coded (SED-TC) transmission. Interleaving is performed over 4 bursts (2 bursts for MCS-8/9 with 8PSK). It can be observed, that in most cases turbo equalization has a better BLER performance than turbo–coded transmission already after a single iteration and after 4 iterations, a gain of 0.7–2.1 dB is achieved by turbo equalization for all proposed 16QAM and 32QAM MCS’s. As an example, the BLER curves of 16QAM MCS-7 and MCS-10 are shown in Fig. 5. The poor performance of the UTRAN turbo code can be explained by the relatively short block lengths of GSM/EDGE.

EXIT chart analyses have been performed for the equalizer and decoder components of turbo equalization. In Fig. 6(a), results are shown for the used convolutional code ($R = 1/2$, $K = 7$) and the RSSE–based equalizers for precoded transmission ($C = [1 \ 0 \ 0 \ 0 \ 0]^T$) and transmission without precoding at $10 \log_{10}(E_b/N_0) = 9$ dB and 11 dB, respectively. Extrinsic a priori information values $I_{EQ^{in}}$ are used as input to the equalizer and (extrinsic) output information values $I_{EQ^{out}}$ are obtained, which depend on the used (fading) channel realizations of the adopted TU channel. 1000 channel realizations have been drawn, and the corresponding input–output transfer characteristic has been stored for each. An adequate measure associated to the BLER performance is now necessary for the EXIT chart of the fading channel. For this, the average extrinsic information corresponding to 4 randomly picked fading channel realizations (according to the interleaving of a block over 4 transmission bursts) is calculated for each considered user. Furthermore, the 90% quantile of this random quantity has been determined (for each value of a priori information $I_{EQ^{in}}$ separately) and is shown in Fig. 6(b). Therefore, the depicted equalizer EXIT chart represents the 90% contour, and performance for only 10% of all transmission blocks is worse on average. For additional precoding, the effects on EXIT chart characteristic depend on the region of a priori information values, so that the curves for precoding end up in the upper right corner ($I_{EQ^{in}} = 1$, $I_{EQ^{out}} = 1$) of the diagram, with the drawback, that the $I_{EQ^{out}}$ values are decreased for small $I_{EQ^{in}}$ values. In the diagram, the curve for the considered convolutional code is additionally shown. Here, the a priori input information $I_{CC^{in}}$ is depicted versus the extrinsic output information $I_{CC^{out}}$ (swapped axes), so that convergence of turbo equalization can be predicted from the diagram to a certain extent. For the equalization scheme with precoding, it can be concluded from the EXIT chart, that for $10 \log_{10}(E_b/N_0) > 9.2$ dB convergence to the error–free state should be possible for more than 90% of all transmission blocks, if block lengths are chosen large enough and if an appropriate interleaver is applied. For the case without precoding, convergence to the upper right corner of the EXIT chart cannot be achieved, but due to the high constraint length of the code, large a posteriori information values of the source bits are obtained at about $10 \log_{10}(E_b/N_0) = 11.2$ dB. For this value, the intersection point of the equalizer and decoder EXIT chart curves (not shown in the diagram) is determined as $I_{CC^{in}} = 0.715$. Via the corresponding output information of the code with respect to the source bits, an estimated related BLER (assuming independent bit errors) of 10% can be derived [24]. In the simulated curves after 6 equalization and decoding iterations shown in Fig. 6(b), a BLER of 10% is valid at about $10 \log_{10}(E_b/N_0) = 10.3$ dB. Therefore, the BLER performance of the considered equalization schemes can be predicted by EXIT charts within less than 1 dB. For transmission with precoding, the number of used iterations as well as the considered block lengths are the main limiting factors, that result in slightly worse performance than the prediction. Additionally, an appropriate interleaver design is required. For the case without precoding, the assumption of
independent bit errors (used to derive the BLER) does not hold in most cases, so that the according prediction is slightly too pessimistic. Furthermore, consistency of the equalizer soft output information (which is usually not fulfilled for reduced-state equalizers) is required, strictly speaking, but for practical scenarios it is found, that consistency is of minor importance [5].

V. CONCLUSION

Different packet data transmission schemes for GSM/EDGE have been analyzed in this paper. It could be observed, that turbo equalization is an interesting candidate as a receiver algorithm for the convolutionally coded schemes of the evolved GSM/EDGE standard at moderate complexity increase. Compared to turbo-coded transmission without turbo equalization, similar performance is already observable after the first turbo equalization iteration for most considered transmission schemes. For more turbo equalization iterations, even better performance may be obtained.

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