On Equalization for EDGE Mobile Communications

Wolfgang H. Gerstacker and Robert Schober

Lehrstuhl für Nachrichtentechnik II

Universität Erlangen-Nürnberg

Cauerstraße 7/NT, D-91058 Erlangen, Germany

Tel.: +49 9131 85 27667

Fax: +49 9131 85 28919

E-mail: gersta@nt.e-technik.uni-erlangen.de

Abstract

In this letter, an equalization concept for the novel radio access scheme EDGE is proposed, by which high performance can be obtained at moderate computational complexity. Because high–level modulation is employed in EDGE, optimum equalization as performed in most GSM receivers is too complex, and suboptimum schemes have to be considered. It is shown, that delayed decision–feedback sequence estimation (DDFSE) is a promising candidate. For various channel profiles, approximations for the bit error rate of DDFSE are given and compared with simulation results. It turns out, that a discrete–time allpass prefilter creating a minimum–phase overall impulse response is indispensible for a favourable tradeoff between performance and complexity. Additionally, the influence of the receiver input filter is investigated.

Index Terms: Mobile communications, equalization.

1 Introduction

Currently, several third-generation mobile communication systems are under standardization. Besides completely new radio access schemes, such as Universal Mobile Telecommunications Service (UMTS), which is based on wideband code–division multiple access (WCDMA), these efforts also include the evolution of the existing time–division multiple access (TDMA) standards GSM and IS-136 towards significantly higher spectral efficiency. For this, a novel common physical layer, EDGE (Enhanced Data Rates for GSM Evolution) [1], will be introduced for both TDMA schemes. EDGE improves spectral efficiency by applying the modulation format 8–ary phase–shift keying (8PSK) instead of binary Gaussian minimum–shift keying (GMSK) which is used in GSM. In order to enable a smooth transition from GSM to EDGE, additional important system parameters such as symbol rate and burst duration remain unchanged.

For equalization, the modification of the modulation scheme is of high significance. This letter discusses suitable equalization strategies for EDGE, which guarantee high performance and can be implemented with moderate complexity.

2 Transmission Model

In the following, all signals are represented by their complex–valued baseband equivalents. At the transmitter, binary data is mapped onto 8PSK symbols $b[\cdot] \in \mathcal{A} = \{e^{j2\pi\nu/8} | \nu \in \{0, 1, \ldots, 7\}\}$. The discrete–time received signal, sampled at times kT (T is the symbol duration) at the output of the receiver input filter, can be written as

$$r[k] = \sum_{\nu=0}^{L-1} h[\nu] b[k-\nu] + n[k]. \tag{1}$$

Here, $h[\nu]$, $0 \leq \nu \leq L-1$, denote the coefficients of the combined discrete–time impulse response (CIR) of transmit filter, channel, and receiver input filter. L is the length of the CIR. For simplicity, it is assumed, that the channel is constant during each burst, but varies from burst to burst, which corresponds to low or moderate vehicle speeds, and that the receiver has perfect knowledge of the CIR. The transmit filter of EDGE is given by a linearized GMSK impulse. For the receiver input filter, we assume a square–root Nyquist frequency response; two special cases thereof are the whitened matched filter (WMF) [2], which is an optimum input filter, and the square–root raised cosine (SRC) filter. The (continuous–time) received signal is impaired by additive white Gaussian noise, which is characterized by the single–sided power spectral density N_0. This results in a zero mean discrete–time white complex Gaussian noise $n[\cdot]$. The mean received energy per bit is given by $\bar{E}_b = \bar{E}_h/3$ with mean channel energy $\bar{E}_h = \mathsf{E}\left\{\sum_{\nu=0}^{L-1} |h[\nu]|^2\right\}$, where $\mathsf{E}\{\cdot\}$ denotes expectation.

3 Equalizer Concepts

EDGE receivers should be able to cope with the same channel conditions as GSM receivers. Therefore, equalizers have to be tested for the channel profiles specified in [3], i.e., typical urban area (TU), rural area (RA), hilly terrain (HT), equalizer test (EQ), and static (ST, no channel dispersion). For the profiles HT and EQ, the discrete–time CIR has an effective length of $L = 7$, whereas for the remaining profiles, fewer taps are sufficient for an accurate characterization (e.g. $L = 4$ for TU).

In GSM, maximum–likelihood sequence estimation (MLSE) using the Viterbi algorithm (VA) [2], which is the optimum equalizer in terms of sequence error probability, is widely employed. The computational complexity of MLSE is directly related to the number of states of the underlying trellis diagram of the VA, which is given by $Z = M^{L-1}$, where M denotes the alphabet size. Hence, MLSE has a maximum complexity of Z=64 states for GSM. In contrast to that, complexity of MLSE is prohibitively high for EDGE. Even for TU, a full–state MLSE would require $Z = 512$ states, which is currently far too complex for a practical implementation. Thus, suboptimum equalizer concepts have to be considered for EDGE. In this letter, reduced–state trellis–based equalizers are discussed. Among the schemes in this class, delayed decision–feedback sequence estimation (DDFSE) [4] seems to be a very promising choice because of its high regularity and good performance. DDFSE takes only the first K, $1 \leq K \leq L$, taps of the CIR for definition of a trellis diagram with $Z = M^{K-1}$ states, whereas the influence of the remaining taps is taken into account by per–survivor processing. For $K = 1$ and $K = L$, the limiting cases decision–feedback equalization (DFE) and MLSE result, respectively. In order to obtain high performance, a minimum–phase CIR is essential for DDFSE [4]. Therefore, a discrete–time prefilter, which ideally has an allpass characteristic, should be introduced in front of equalization in order to transform the CIR into its minimum–phase equivalent. For an efficient calculation of the prefilter coefficients, several strategies are proposed in literature, cf. e.g. [5].

4 Analytical Performance Evaluation

For a fixed CIR, the bit error rate (BER) of DDFSE can be approximated by [4]

$$\text{BER} \approx C \cdot Q\left(\sqrt{d_{\min}^2 \frac{\bar{E}_b}{N_0}}\right). \tag{2}$$

Here, $Q(\cdot)$ denotes the complementary Gaussian error integral, $d_{\min}$ is the normalized minimum Euclidean distance being a function of the overall CIR (including prefilter if applied) and K, and C is a constant factor, which is close to unity in most cases for the considered scenario and therefore can be neglected. $d_{\min}$ can be calculated using the Dijkstra algorithm [6], which performs a path search in the state transition diagram of the equalizer and is highly efficient for minimum–phase CIRs.

In our case of random, time–invariant channels, the BER expression of Eq. (2) has to be averaged over a sufficient number N of random channel realizations, yielding

$$\text{BER} \approx \frac{1}{N} \sum_{i=0}^{N-1} Q\left(\sqrt{\frac{E_{h,i}}{\bar{E}_h}\, d_{\min,i}^2 \frac{\bar{E}_b}{N_0}}\right), \tag{3}$$

where $E_{h,i}$ and $d_{\min,i}$ denote the energy and the normalized minimum Euclidean distance of the CIR corresponding to the ith channel realization, respectively. For the following numerical results, $N = 10000$ is selected. It should be noted that performance of MLSE and DDFSE with K close to L cannot be assessed by means of simulations for the given scenario, because computational complexity of these schemes is even too high for this. Thus Eq. (3) is crucial in order to obtain theoretical limits for equalizer performance.

5 Numerical Results and Conclusions

Fig. 1 shows calculated and simulated BERs for HT. An SRC receiver input filter with roll–off factor $\alpha = 0.3$ has been chosen. In the case of DDFSE without prefiltering, only simulation results could be obtained, because the Dijkstra algorithm did not converge in reasonable time for many channel realizations having a nonminimum–phase characteristic. For additional prefiltering, calculated and simulated results are of the same order. The maximum difference is about 1.3 dB, mainly caused by error propagation, which plays a role especially for small K and is not taken into account for the BER approximation of Eq. (3). The loss of DDFSE ($K = 2$) compared to MLSE is limited to 1.6 dB at BER $= 10^{-5}$; for $K = 3$ and $K = 4$, only slight improvements can be obtained compared with $K = 2$. Hence, a DDFSE with $Z = 8$ states and allpass prefiltering seems to be a reasonable choice for HT. Without prefiltering, a significant performance degradation occurs.

In Fig. 2, results for MLSE and DDFSE ($K = 2$) with prefiltering are shown for all test profiles. In addition to the SRC filter, also the case of a WMF, individually designed for each of the random channel realizations, is considered. In general, simulated and calculated bit error rates are in good agreement. Obviously, the suboptimum SRC filter causes only minor performance degradation. DDFSE with $Z = 8$ states ($K = 2$) seems to be well suited for all profiles. With this choice, a performance rather close to the theoretical limit given by MLSE is possible with an only moderate computational complexity.

Because inherent path diversity is maximum for EQ and minimum for RA, which resembles the flat Ricean fading channel with small Ricean factor, these two profiles correspond to best (behind the limiting case ST) and worst equalizer performance, respectively.

References

[1] A. Furuskär, S. Mazur, F. Müller, and H. Olofsson. EDGE: Enhanced Data Rates for GSM and TDMA/136 Evolution. *IEEE Personal Communications Magazine*, pages 56–66, June 1999.

[2] G.D. Forney, Jr. Maximum–Likelihood Sequence Estimation of Digital Sequences in the Presence of Intersymbol Interference. *IEEE Transactions on Information Theory*, IT-18:363–378, May 1972.

[3] ETSI. *GSM Recommendations 05.05 Version 5.3.0*, December 1996.

[4] A. Duel-Hallen and C. Heegard. Delayed Decision–Feedback Sequence Estimation. *IEEE Trans. on Commun.*, COM-37:428–436, May 1989.

[5] W. Gerstacker and J. Huber. Improved Equalization for GSM Mobile Communications. In *Proceedings of International Conference on Telecommunications (ICT'96)*, pages 128–131, Istanbul, April 1996.

[6] T. Larsson. A state–space partitioning approach to trellis decoding. Technical report 222, Chalmers University of Technology, Göteborg, 1991.

Figures with Captions

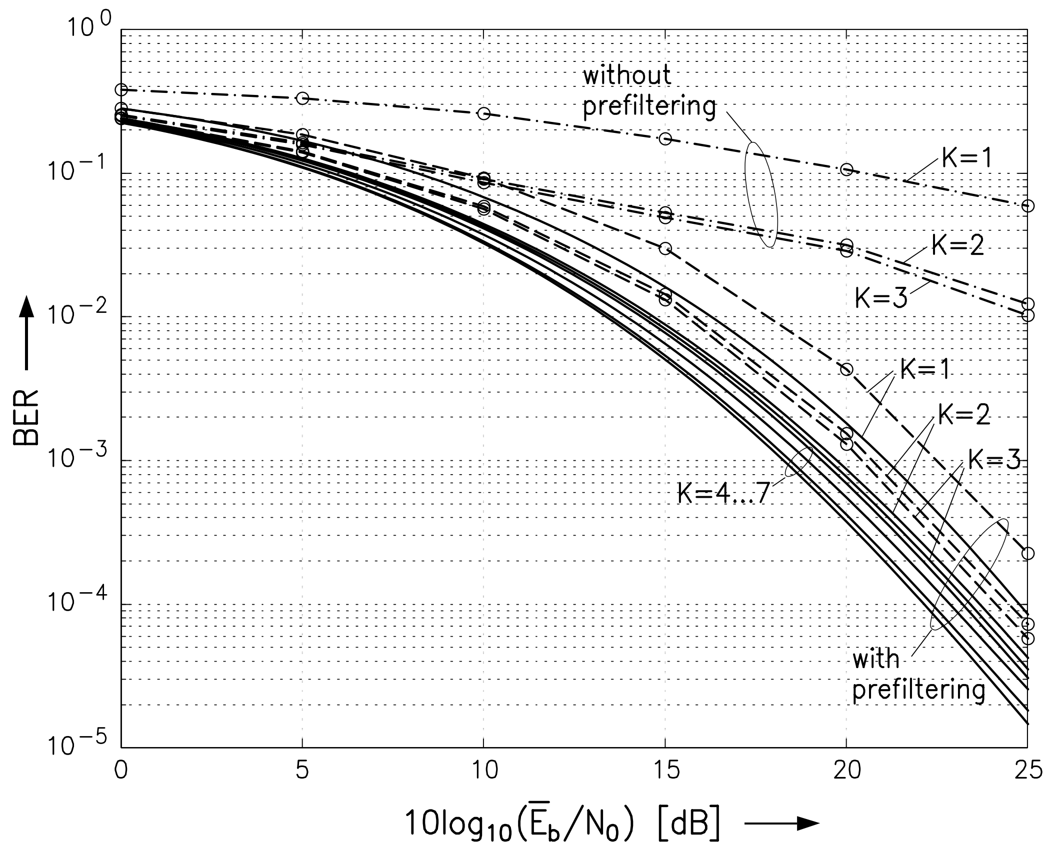

Figure 1: BER vs. $10\log_{10}(\bar{E}_b/N_0)$ for various DDFSE schemes including DFE and MLSE (HT profile). Solid lines: BER approximations according to Eq. (3) (with allpass prefiltering). Dashed lines: simulation results (with prefiltering). Dash–dotted lines: simulation results (without prefiltering).

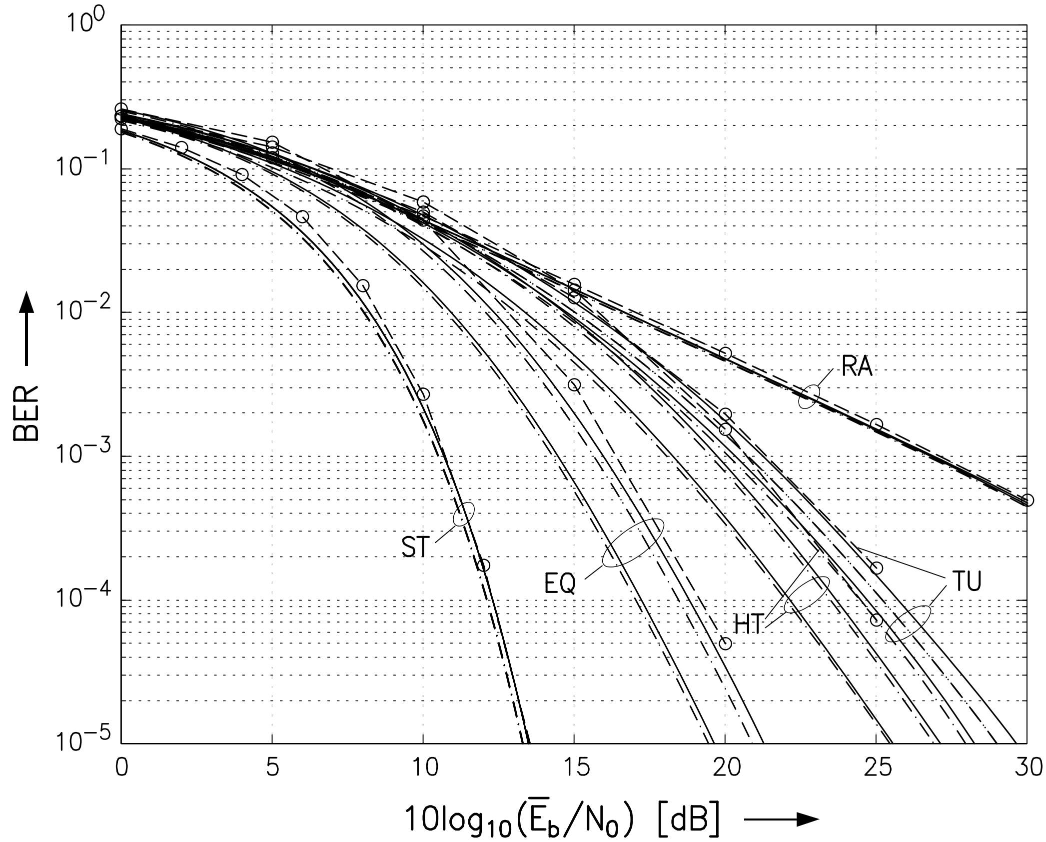

Figure 2: BER vs. $10\log_{10}(\bar{E}_b/N_0)$ for all profiles. Solid lines: BER approximations according to Eq. (3) for MLSE, DDFSE ($K = 2$) (SRC, $\alpha = 0.3$). Dashed lines: simulation results for DDFSE ($K = 2$) (SRC, $\alpha = 0.3$). Dash–dotted lines: BER approximations according to Eq. (3) for MLSE, DDFSE ($K = 2$) (WMF). Allpass prefiltering is assumed in each case.

Figure 1

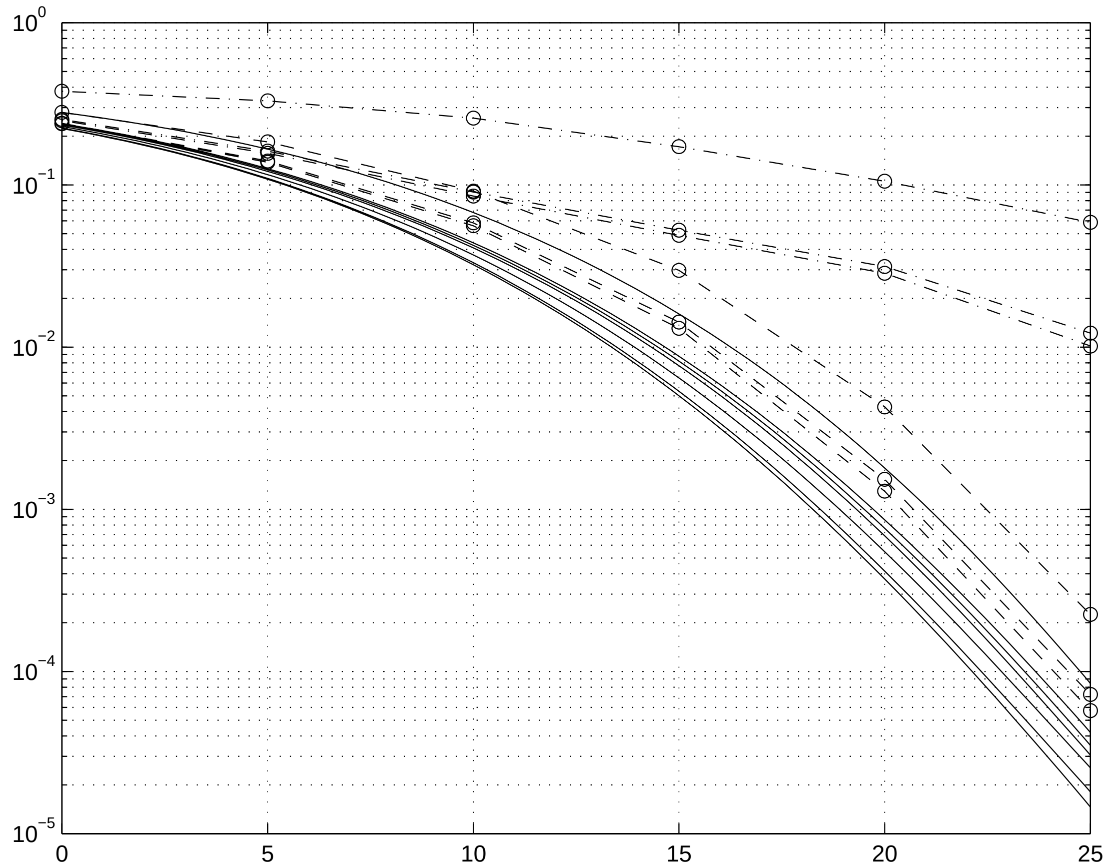

Figure 2