Abstract
Third-generation tetherless mobile/personal communication system development is faced with the conflicting requirements of providing both highest protection against noise/interference and keeping the transmission delay as low as possible.
In this paper, a novel algorithm for combined equalization and decoding of interleaved coded modulation is presented. The new \( M \)-CED algorithm is an extension of the "conventional" CED in two ways: i) per-survivor processing is used in the equalizer section in order to reduce the decoding latency, and ii) \( M > 1 \) equalizer states per code state are kept in the survivor memory so as to further improve on the bit error performance. Results on simulated bit error rates give insight into the effect of various channel and receiver parameters, in particular, interleaver length (thus latency), number of equalizer states, and the choice of trellis codes. The entire digital receiver with feedforward sync, appropriate prefiltering and \( M \)-CED equalization/decoding is shown to combine the advantages of being robust against fading and yielding good BER performance also at low delay.

1 Introduction
In the design of mobile/personal communication systems, one attempts to achieve both a high degree of protection against impairments (noise, adjacent- and co-channel interference) and as low a decoding latency as possible, particularly for speech transmission [5]. Obviously, these are conflicting requirements on fading channels since some kind of explicit or implicit time diversity, most often through bandwidth-efficient channel coding and interleaving [6,7,8,16,19,23] must be provided for in order to bridge periods of deep channel fades, especially if antenna diversity is not available.
Here, we are concerned with advanced TDMA-based multitzer transmission of linearly modulated data packets over dispersive fading channels. Each packet is assumed to be headed by a preamble \((P_k)\) for synchronization and decoder initialization purposes, followed by \( D \) trellis-encoded and interleaved random data symbols. If faster fading must be handled by the receiver, an additional training segment of length \( P_L \) may have to be appended at the end of the packet so as to maintain correct receiver sync [10]. Using interleaved trellis codes with large effective code length (ECL) combined with frequency hopping (FH) and feedforward receiver synchronization constitutes a very effective antiadual technique [12].
Recently, a computationally-efficient combined equalization and decoding (CED) algorithm has been proposed by Melhu [15] and developed further by the author [13,11]. In the CED scheme, the block interleaver matrix must have a special structure tailored to the packet length \( F \) and the survivor depth \( S \) of the Viterbi decoder (Fig. 2 below) [15]. Since \( S \) must be large w.r.t. the code constraint length, the interleaver size and thus the latency must also be large in order to achieve near-optimal decoder performance. For instance, a survivor depth of \( S = 20 \) leads to a late C3 in excess of 100 ms if the packets are transferred in a GSM-like time slot format [4].
In the new \( M \)-CED presented here, the latency \( F \times J \) is reduced by a factor of \( Z \) by means of per-survivor processing (PSP). The survivor depth \( S = Z \times J \) can be kept large by retarding the final decoder decisions until \( Z \) rows in the deinterleaver matrix have been processed. The operation of the conventional CED can be interpreted as a joint equalization and decoding algorithm that considers one equalizer state per code state. The equalizer portion therefore operates in the manner of a decision-feedback equalizer (DFE). In the case of spread modulation, the so-called \( M \)-algorithm is known to improve on the DFE performance by keeping \( M > 1 \) equalizer states in the survivor memory [12,18]. Interestingly, the performance of the optimal but much more complex Viterbi equalizer is approached very quickly already for very small \( M \), provided that suitable receiver prefiltering is performed in front of the equalizer [3,9].
Applying the \( M \)-algorithm concept to combined equalization and decoding leads to a natural extension of the CED, termed \( M \)-CED algorithm, where \( M \) equalizer states per code state are kept in the survivor memory. Introducing both per-survivor processing and the \( M \)-algorithm concept has the potential of reducing the latency and simultaneously improving the BER performance.

2 Signal Transmission and System Model
The block diagram of the entire transmission system is outlined in Fig. 1. The input bits \( b_k \) are trellis-encoded and mapped onto PSK or QAM data symbols \( a_k \) (index \( k \): non-interleaved time scale) which are written row-wise into the block interleaver matrix. After adding training segments for synchronization, the channel symbols \( a_k \) (index \( k \): interleaved channel symbol time scale) are read out column-wise and converted to a rate-2/\( T \) symbol stream which is filtered by the low-pass-equivalent physical channel \( h(k) \) whose memory spans \( L \) symbol intervals. The received \( T/2 \)-spaced signal \( r_k \) is prefiltered in the detection path by the time-variant channel matched filter (MF) \( m(k) \) and, following decimation to rate 1/\( T \), the whitening filter (WF) \( w(k) \). Assuming ideal filters, the cascade of rate converter, channel \( h(k) \), MF \( m(k) \), rate decimator and WF \( w(k) \) yields a causal, minimum phase equivalent \( T \)-spaced
channel impulse response \( f(k) \) [17, 8] so that the resulting rate-1/\( T \) signal \( v_k \) (see Fig. 1) can be expressed as

\[
v_k = \sum_{i=0}^{L} f_i(k) \cdot a_k + \eta_k
\]

where \( \eta_k \) is the whitened noise behind the WF.

The channel access scheme and packet duration assumed here are oriented at the GSM system [4]: throughput 270 kbit/s/carrier, 8 TDMA packets per carrier, 4.615 ms spacing between packets pertaining to a user, packet length \( 511 \mu s \). However, here the data bits are trellis-encoded and linearly modulated. Since each symbol contains two information symbols, the symbol rate is halved to \( 1/1.5 \) symbols. The preamble can be used for initialization of the decision-directed channel estimation present in decision-directed channel estimators, enables frequency-hopping, and is more robust against fading. Moreover, the preamble can be used for initialization of the decision-directed channel estimation procedure described below.

The training segments are designed for near-optimal feedforward receiver synchronization. Feedforward sync is most advantageous since it eliminates sync error propagation present in decision-directed channel estimators, enables frequency-hopping, and is more robust against fading. Moreover, the preamble can be used for initialization of the decision-directed channel estimation procedure described below.

Here, a packet of length \( F = 74 \) is assumed to contain two training segments (pre- and postamble, Fig. 2) comprising \( P_1 = 9 \) and \( P_2 = 7 \) symbols, respectively. From these training segments, channel estimates \( h(k) \) (here 10-dim. vectors spanning \( L = 4 \) symbol intervals) are computed for some positions \( k \) within the data block. The \( f_i(k) \) so obtained are then mapped onto MF-, WF-, and equalizer coefficient sets \( f(k), f(k) \) [10]. As test simulations have shown, only four channel/MF coefficient sets \( h(k) \), \( f(k) \) and only two WF/equalizer coefficient sets \( f(k) \) (9-dim.), \( f(k) \) (4-dim.) need to be generated per packet to achieve quasi-optimal sync performance.

For this study, two trellis codes have been selected that are very effective in fading environments due to their large effective code lengths (ECL). The first code (“8-PSK”) is a rate-2/3, 16-state, 8-PSK Ungerboeck code [20] with asymptotical gain (AWGN channel) 4.1 dB and ECL 3. The second code (“16-QAM”) by Mohor and Lodge [16] is a 2-rate-1/2, 2×16-state, 2×4-PAM (16-QAM) code with asymptotical gain 3.4 dB and ECL 5.

3 The Conventional CED Algorithm

The frequency-hopped TDMA packets form the columns of the interleaver matrix so that the number of rows (including training) is \( F = P + D \) where \( P \) denotes the number of training symbols per packet (here \( P = P_1 + P_2 = 16 \), \( D = 58 \)). The interleaver depth is therefore the number \( D \) of encoded data symbols in a packet.

As soon as the deinterleaver matrix has been filled with received and preprocessed packets (samples \( v_k \), see Fig. 2), the row-wise equalization and decoding process begins. Every iteration \( i = 1 \rightarrow 58 \) starts with a DFE-like equalization step where the ISI is canceled by subtracting the ISI postcursor from \( v_k \):

\[
\hat{a}_k = \left( a_k + \sum_{i=1}^{L} f_i(k) \cdot \hat{a}_{k-i} + \eta_k \right) - \left( \sum_{i=1}^{L} f_i(k) \cdot \hat{a}_{k-i} \right)
\]

\( v_k \)

\[
\text{ISI estimate}
\]

\[
\hat{a}_k = a_k + \eta_k \quad \text{if} \quad \hat{a}_{k-l} = a_{k-l}
\]

The ISI estimate is formed via the final (thus more reliable) symbol decisions \( \hat{a}_{k-l} \) \( (l = 1, \ldots, L) \) “above” the current symbol \( a_k \). Assuming ideal prefiltering and correct final
decisions, the ISI-canceled sample $\hat{a}_k$ is just a noisy replica of the transmitted symbol $a_k$, so that $\hat{a}_k$ can be passed on to the Viterbi decoder. Switching to decoder indexing $k \rightarrow i$ and using the ISI-canceled sample $\hat{a}_k = a_k$, metric increments $\lambda_{mn,i}$ associated with state transitions $m \rightarrow n$ are computed and added to the previous path metrics $\gamma_{mn,i}$ so as to form "extended" path metrics $\gamma_{mn,i}$. From the paths merging in a successor state $n$, the path with the best metric is selected and its survivor bit sequence $b_{n,i} = \{b_{n,i}, b_{n,i+1}, \ldots\}$ as well as its survivor channel symbol sequence $a_{n,i} = \{a_{n,i}, a_{n,i+1}, \ldots\}$ is stored in the survivor memory.

In order to be able to perform ISI cancellation $v_k \rightarrow \hat{a}_k$ as described above, the channel symbols $a_{1,0} = \{a_{1,0}, a_{1,1}, \ldots\}$ contain all current samples $v_k$. In particular $a_{k-1} = a_{k-1,i}$, the CED generates a final symbol decision $a_{k-1,i}$ on survivor symbols $a_{n,i}$.Before metric computation can be performed. Therefore, the survivor memory of the Viterbi decoder must be truncated to a survivor depth $S \leq J$ not larger than the interleaver length $J$.

Near-optimal decoder performance is obtained for survivor depths $S$ that are several times larger than the code constraint length. If the GSM-like transmission format and trellis codes described above are used, survivor depth $S$ (number of packets per interleaver block) in the order of 20-30 lead to latencies in excess of 100 ms which is hardly tolerable for speech transmission. Latencies below 50 ms call for truncating the survivor depth to no more than $S = 10$ symbols.

In Fig. 3, the simulated BER performance of the entire receiver with conventional CED and frequency-hopped packet transmission over the GSM hilly-terrain channel as specified by CEPT [4] is shown for the 8-PSK code. Curves are displayed for survivor depths of $S = 25$ (near-optimal, 125 ms latency), $10$ (50 ms) and $5$ (25 ms), and for Doppler frequencies of $\leq 200$ Hz (900/1800 MHz: $\leq 250/125$ km/h) and 400 Hz.

**Figure 3: BER Performance of CED. 8-PSK Code**

From the figure and similar results for the 16-QAM code (not shown here), one observes the following:

- The receiver is robust against fading up to Doppler frequencies of at least 200 Hz (relative Doppler $1.5 \times 10^{-3}$). At higher Doppler, channel estimation and thus receiver sync experiences a degradation.
- Under ideal conditions ($S=25$, Doppler $\leq 200$ Hz), the 16-QAM code with ECL 5 (8/12 dB SNR) is needed to obtain a BER of $10^{-4}$/$10^{-5}$ performs better than the 8-PSK code with ECL 3 (9/13 dB SNR for BER $10^{-7}$/$10^{-4}$).
- The BER degradation resulting from truncating the survivor depth is substantial; due to its larger constraint length, the 16-QAM code is more sensitive than the 8-PSK code.

These results reflect a tradeoff between latency and performance which needs to be improved upon if quasioptimal performance at lower delays is desired.

### 4 The New M-CED Algorithm with Per-Survivor Processing

In order to circumvent the delay limitations of the conventional CED, the algorithm is modified in two ways. As discussed above, the symbols $a_{k-1,i}$ are the current symbol $v_k$ must be known (see Fig. 2) in order to perform ISI cancellation. Now instead of forming final decisions $\hat{a}_{k-1} = a_{k-1,i}$ by survivor memory truncation, the concept of per-survivor-processing (PSP) is invoked; i.e. the symbols $a_{n,k-1} = a_{n,i-1,i}$ contained in the survivor memory $a_{n,i}$ are used for per-survivor-dependent ISI cancellation:

$$\hat{a}_{n,k} = v_k - \sum_{i=1}^{L} f_i(k) \cdot a_{n,k+i}$$  \hspace{1cm} (3)

By such, the survivor memory can be extended to more than one row in the deinterleaver matrix. The process of equalization and decoding of the CED algorithm with PSP is illustrated in Fig. 4. In the figure, the survivor memory is extended to $Z$ rows so that the survivor depth becomes $S = Z \cdot J$. Therefore, $J$ and thus the latency can be reduced without having to truncate the survivor memory. However, $J$ must not be smaller than the code ECL. In order to provide for sufficient interleaving: remember that the symbols $a_{n,k-1} = a_{n,i-1,i}$ in one column tend to fade simultaneously since they belong to the same packet.

The second modification to the CED algorithm is motivated by the fact that a large percentage of symbols $a_{n,k-1} = a_{n,i-1,i}$ in the channel memory "above" $v_k$ may in fact be wrong since they belong to survivors ending in an incorrect code state $n$. This is best explained via an example of survivors in the code trellis diagram as displayed in Fig. 5. Since $J$ is now smaller than the survivor depth $S$, many of the survivor paths pertaining to code states $n$ have not yet merged at decoding instants $i = J, i = 2J, \ldots$ so that their associated survivor symbols $a_{n,i}$ in particular $a_{n,J-1} = a_{n,i-1,i}; a_{n,J-2} = a_{n,i-2,i}, \ldots$ in the channel memory "above" $v_k$, may differ from those of the correct path. Using these symbols $a_{n,k+i} = a_{n,i+i,i}$ for survivor-dependent ISI cancellation therefore yields incorrectly "compensated" samples $\hat{a}_{n,i}$ so that the metrics computed from $a_{n,i}$ are likewise incorrect.

An elegant remedy is found by recognizing that a collection of symbols $a_{n,k-1} (i = 1, \ldots, L)$ in the channel memory can be interpreted as an equalizer state, just like in the case of uncoded modulation. In this context, the ISI cancellation mechanism of the CED is identical to the operation of a decision-feedback equalizer (DFE) since only one equalizer state is kept per code state. At the other end of the complexity scale, the full-fledged Viterbi equalizer (VE) [14] keeps
(M-)CED Algorithm with Per-Survivor-Processing

\[ W_F \]

output signal

WF output signal

\[ a_n, k = v_k - \sum_{l=1}^{L} f_l(k) a_{n,k-l} \]

Figure 4: The (M-)CED Algorithm with PSP

\[ Q^L \]

equalizer states per code state, where \( Q \) is the cardinality of the symbol alphabet (here 8 or 16) and \( L \) the channel memory (here 4). Fortunately, it is not necessary to implement the VE since it has been shown that keeping a small number \( M \) of equalizer states already yields quasi-optimal performance [1, 2, 18].

Therefore, it seems reasonable to extend this \( M \)-algorithm concept also to joint equalization and decoding. This leads to the \( M \)-CED algorithm where \( M \geq 1 \) survivors (equalizer states) are kept per code state. At each iteration \( i = 1 \rightarrow i \), the \( M \) survivors \( a_{n,m,i} \) \((m = 1, \ldots, M)\) of each code state \( n \) are used to perform \( M \) associated ISI cancellations yielding samples \( \hat{a}_{n,m,i} \). From these, \( M \) extended path metrics for each possible state transition are computed in the Viterbi decoder. Depending on the code trellis, a number (8-PSK code: \( 4M \), 16-QAM code: \( 2M \)) of paths merge in each successor state. Of these merging paths, the \( M \) paths with best metrics are selected and their survivors stored for the next iteration.

5 Receiver Performance

The bit error performance results reported here are of preliminary nature; as yet, only a few transmission scenarios have been investigated in simulation experiments. Simulations have been conducted for the GSM hilly terrain channel [4] (mild ISI spanning \( L = 4 \) symbol intervals) only. Also, the simulation of reference curves for ideal channel estimation will have to be completed in the near future. Nevertheless, the benefits of the \( M \)-CED algorithm with PSP and the impact of some important parameters do become apparent from the results already obtained.

Here, BER results are shown for the entire digital receiver with feedforward sync and \( M \)-CED joint equalization and decoding. Interleaver lengths of \( J = 5, 10 \) and 25 have been tested; the survivor depth \( S \) is fixed at 25. The BER curves for the 8-PSK code (ECL 3) are given in Figs. 6 and 7 for Doppler frequencies of (up to) 200 and for 400 Hz, respectively.

\[ \text{M-CED Performance} \]

GSM hilly terrain channel, 8-PSK code, feedforward sync, \( \leq 200 \) Hz Doppler

Again, the receiver is seen to be robust against fading up to Doppler frequencies of at least 200 Hz (relative Doppler 1.5 \( \times 10^{-3} \)). The BER degradation at faster fading (Doppler 400 Hz) is entirely due to the compromised quality of receiver synchronization.

Concerning the 8-PSK code (ECL 3), one observes the following:

- The use of per-survivor processing (PSP) already yields substantial gains w.r.t. the conventional CED, especially for small interleaver depths \( J \).
- Applying the \( M \)-algorithm concept yields additional gains; and the BER curves "converge" rapidly: beyond \( M = 4 \), further performance gains are very marginal so that \( M = 4 \) is the best compromise between complexity.
The results (not shown here) for the 16-QAM code (ECL 5) reveal the following:

- The gains obtained by virtue of PSP are noticeable (especially for J=10) but are smaller than for the 8-PSK code.
- The additional M-CED gains (M ≥ 1) are also smaller than for 8-PSK, but the BER curve "converge" even faster; here, M=2 is the best compromise between complexity and performance.
- The latency reduction factors are smaller (usually ≤ 2) than for the 8-PSK code. Only at 400 Hz Doppler and with J=10, the latency can be reduced by a factor of 2 without compromising BER performance (the M-CED [J=10] performs almost as well as the CED [J=25]).

In summary, the M-CED algorithm significantly improves on the tradeoff between latency and performance whereby quasi-optimal performance can be achieved at lower delays. Interestingly, the results reveal another tradeoff: between latency and effective code length (ECL). Without delay limitations, codes with large ECL are known to yield the best performance on fading channels due to their high degree of inherent time diversity. However, large ECL’s can become counterproductive when it comes to minimizing the latency. The results obtained so far clearly indicate that, under severe delay constraints, the 8-PSK code with smaller ECL (→ smaller constraint length) should be preferred to the 16-QAM code.

6 Conclusions

A digital receiver including feedforward sync, receiver prefiltering and M-CED processing has been studied. BER simulation results give insight into the effect of the interleaver length J (thus latency), number M of equalizer states, and the choice of trellis codes with different ECL’s. In summary, the receiver with feedforward synchronization and the new M-CED algorithm has been shown to be a good candidate for 3rd generation digital mobile/personal radio since, apart from being robust against fading, it combines the advantages of good BER performance with low delay and moderate cost of implementation.

References