Blind GSM Channel Estimation under Channel Coding Conditions

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Abstract

As a fundamental component of the Global System for Mobile communications (GSM), channel coding aims at improving speech and data transmission quality when the signal encounters disturbances. In this paper, we put the main emphasis on investigating the application of blind channel estimation approaches based on Second Order Statistics (SOS) and/or Higher Order Statistics (HOS) to the identification of a full rate data Traffic CHannel (TCH/F9.6). Giving the bit error rates (BERs) after equalization and channel decoding in terms of the $\bar{E}_b/N_0$ ratio, we show that there are blind channel estimation algorithms which are almost as efficient as non-blind methods even if the disturbance is non-Gaussian.

1 Introduction

Optimum receivers in digital communication systems require the knowledge of the transmission channel. In general, this knowledge is not available so that the channel's equivalent baseband impulse response has to be estimated. State-of-the-art mobile communication systems transmit *training sequences* to assist the receiver in estimating the channel impulse response. However, their repeated transmission leaves the communication system with a considerable overhead which could be used for the transmission of additional data sequences, if the channel estimation was solved *blindly*. In recent publications [2, 3, 8], we have shown that some blind HOS based approaches yield promising results in estimating GSM channels. However, these papers did not consider the influence of channel coding which enables a sufficient error protection in the GSM transmission scheme.

According to the needs for user data and speech transmission, the GSM *Specifications* distinguish two kinds of Traffic CHannels (TCHs) with different transmission modes [7]:

- TCH/FS, TCH/F9.6, TCH/F4.8, and TCH/F2.4 are the transmission modes for *full rate* speech at 13 kbit/s and data at 12, 6 and 3.6 kbit/s, respectively.
- TCH/H4.8 and TCH/H2.4 are the *half rate* transmission modes, respectively for 6 and 3.6 kbit/s.

All these modes are combined with different interleaving and coding schemes. Since high rate data transmission over mobile radio channels is playing a growing role, we have carried out our simulations for the full rate data channel TCH/F9.6.

The paper is organized as follows: Section 2 gives an overview of the GSM communication system. Based on the so-called *burst* structure, we outline in section 3 the approaches of non-blind and blind channel estimation. This section also presents the solution of a common drawback of all blind estimation algorithms: The *unknown complex factor*. In section 4, we describe the interleaving and coding schemes used for the TCH/F9.6. Finally, simulation results are presented in section 5 and the paper is concluded in section 6.

2 GSM Data Transmission

Fig. 1 depicts the equivalent baseband representation of the physical layer of a GSM communication system. In the

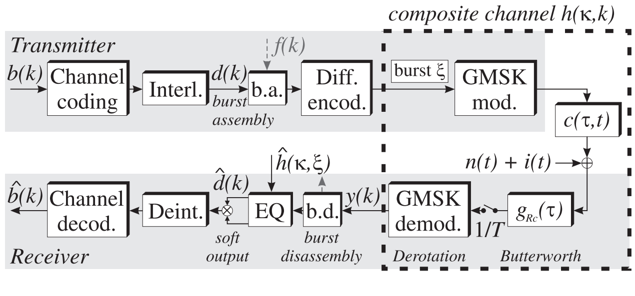

Figure 1: Physical layer of a GSM communication system

transmitter, *blocks* of uncoded data bits $b(k) \in \{0,1\}$ are encoded by a half-rate convolutional coder ($R_c = 1/2$). Thus, every bit $b(k)$ is encoded into a code word of two bits. In order to scramble the coded bit stream, the blocks are interleaved, and assembled together with the reference bits $f(k)$ into differentially encoded radio *bursts* of $N_b = 142$ bits valued $\{-1, 1\}$. Then, each burst is modulated by Gaussian Minimum Shift Keying (GMSK) and transmitted over

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the *time-variant* (TV) channel $c(\tau, t)$. Apart from linear TV distortions, Additive White Gaussian Noise (AWGN) $n(t)$ is present. Moreover, in cellular systems such as GSM, transmitters in remote cells may use the same carrier frequency so that another unwanted signal $i(t)$, called *Co-Channel Interference* (CCI), is superimposed. In the receiver, an anti-aliasing filter $g_{Rc}(\tau)$ is applied in order to suppress adjacent channel interference. Due to differential encoding and GMSK modulation, a simple derotation demodulator can be used upon symbol-rate sampling in order to obtain the sequence $y(k)$.

Maximum Likelihood Sequence Estimation (MLSE) represents a well known procedure to remove intersymbol interference (ISI) from the received digital communication signal. Since MLSE requires the knowledge of the composite channel $h(\kappa, k)$ in order to equalize a block of the demodulated sequence $y(k)$, the following section focuses on blind and non-blind channel estimation. Therefore, let $\hat{h}(\kappa, \xi)$ denote the channel estimate, which will be used to equalize the ξ-th demodulated burst.[1] By using soft-decision algorithms within the equalizer, the system performance can be improved by about 3 dB [9]. Therefore, a Max-Log-MAP soft output detector computes probabilities for each bit $\hat{d}(k) \in \{-1, 1\}$ which can be applied to the Viterbi channel decoder after deinterleaving.

3 Blind and Non-Blind Channel Estimation

According to Fig. 2a), each GSM "normal" burst contains a training sequence (midamble) $f(k)$ of 26 bits surrounded by two packets of 58 data bits emerging from the coded i.i.d. information sequence $d(k)$. Non-blind channel estimates can be derived by solving the relation between the received (corrupted) and the stored training sequences in the *least squares* (LS) sense [2]. However, their repeated transmission leaves a GSM system with an overhead capacity of $26/116 = 22.4\%$, which can be used for other purposes if the channel is estimated blindly.

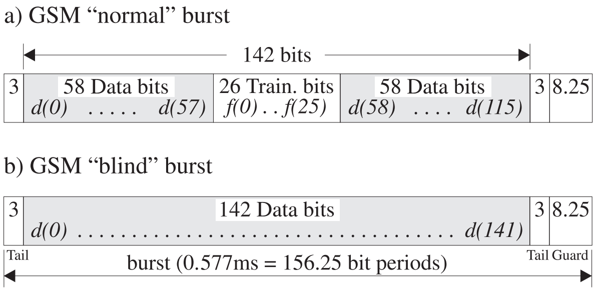

Figure 2: GSM burst structure

The fundamental idea of blind system identification is to derive the complete channel characteristics (including phase information) from the received signal only, i.e. *without* training sequences. Fig. 2b) shows the structure of a "blind" GSM burst which consists of 142 data bits.

Besides the non-blind LS scheme, we have selected the following blind HOS and SOS based channel estimation algorithms for an analysis in this paper:

- EIGENVECTOR APPROACH TO BLIND IDENTIFICATION (EVI) by Boss, Jelonnek, and Kammeyer [1],
- W-SLICE (WS) algorithm by Fonollosa and Vidal [5],
- TXK method by Tong, Xu and Kailath [10].

While EVI is a non-linear method maximizing a fourth order cross-cumulant on a second order boundary condition, WS computes the channel impulse response as a linear combination of 1D (auto)cumulant slices. Finally, TXK is a SOS based approach which can be applied to an excellent new idea of Ding & Li [4] exploiting channel diversity from a single symbol-rate sampled GMSK signal, such as $y(k)$. This method will therefore be referred to as the DL-TXK channel estimation. The TXK algorithm can also be applied to Second Order Cyclostationary Statistics (SOCS), if the received sequence is oversampled, i.e. if the sampling rate is a fraction of the symbol period T. Since in typical GSM receivers the received sequence is symbol-rate sampled, we do not investigate this procedure in this paper. A comparison of the DL and SOCS based TXK can be found in [8].

Generally, algorithms demanding channel diversity (DL and SOCS) are sufficient to retrieve the complete channel characteristics, but there are "singular" channel classes which *cannot* be identified this way. They include channels with common subsystems (i.e. common zeros) in all polyphase subchannels.

Remember that all blind channel estimates have to be obtained from only 142 samples of the received sequence $y(k)$ (see Figs. 1, 2, and 3).

3.1 Blind Channel Estimation Correction

In order to achieve MLSE, the Viterbi algorithm is dependent upon the complete channel characteristics which include correct magnitude *and* phase information. However, a common drawback of *all* existing blind approaches whether they are based on HOS or SOS is the interference of the estimated channel impulse response $\tilde{h}(\kappa, \xi)$ by an unknown complex factor α_ξ (see Fig. 3).

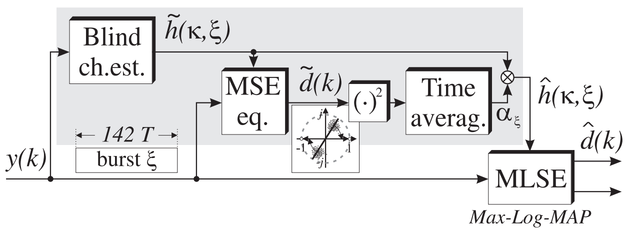

Figure 3: Blind Channel Estimation Correction (BECO)

[1] While in $h(\kappa, k)$, the index $k = t/T$ refers to the symbol period $T \approx 3.7\ \mu$s, the index ξ used in $\hat{h}(\kappa, \xi)$ changes from burst to burst only.

The motivation for blind channel estimation is the avoidance of midambles. Hence, α_ξ cannot be determined from reference symbols either. Fig. 3 illustrates the BLIND CHANNEL ESTIMATION CORRECTION (BECO) scheme, where the $N_b = 142$ bits $y(k)$ of each demodulated burst as well as the q blind estimated coefficients $[\tilde{h}(0,\xi), \cdots, \tilde{h}(q,\xi)]$ are fed into the linear Mean Square Error (MSE) equalizer

$$\mathbf{e}_\xi = (\tilde{\mathbf{H}}_\xi^* \tilde{\mathbf{H}}_\xi + \gamma_\xi \mathbf{I})^{-1} \tilde{\mathbf{H}}_\xi^* \mathbf{i}_\xi. \qquad (1)$$

In equation (1) the n-dimensional column vector $\mathbf{e}_\xi$ is defined as

$$\mathbf{e}_\xi \triangleq [e_\xi(0), \cdots, e_\xi(n)]^T \qquad (2)$$

and $\tilde{\mathbf{H}}_\xi^*$ represents the conjugate transpose of the $(q+n) \times n$ filtering matrix

$$\tilde{\mathbf{H}}_\xi \triangleq \begin{bmatrix} \tilde{h}(0,\xi) & 0 & \cdots & 0 \\ \vdots & \ddots & \ddots & \vdots \\ \tilde{h}(q,\xi) & \cdots & \tilde{h}(0,\xi) & 0 \\ 0 & \tilde{h}(q,\xi) & \cdots & \tilde{h}(0,\xi) \\ \vdots & \ddots & \ddots & \vdots \\ 0 & \cdots & 0 & \tilde{h}(q,\xi) \end{bmatrix}. \qquad (3)$$

For mixed-phase transfer functions the "1" within the $(q+n)$-dimensional column vector $\mathbf{i}_\xi = [0, \cdots, 0, 1, 0, \cdots, 0]^T$ should be located at the centre position $i_0 = \lceil (q+n)/2 \rceil$. For each burst ξ, we have to add the mean noise energy γ_ξ, including CCI, on the diagonal of the auto-correlation matrix $\tilde{\mathbf{H}}_\xi^* \tilde{\mathbf{H}}_\xi$. Filtering $y(k)$ with the equalizer (1) will lead to the linearly equalized sequence $\tilde{d}(k) = e_\xi(\kappa) * y(k)$ which is influenced by the blind channel estimate $\tilde{h}(\kappa, \xi)$. Finally, the complex factor can be calculated by time averaging over one burst,

$$\alpha_\xi = \sqrt{\frac{1}{N_b} \sum_l |\tilde{d}(l)|^2} \cdot e^{\frac{j}{2} \cdot \arg\left\{\sum_l \tilde{d}^2(l)\right\}}, \qquad (4)$$

which yields the corrected blind channel estimate $\hat{h}(\kappa,\xi) = \alpha_\xi \cdot \tilde{h}(\kappa,\xi)$. Note that the transients of $\tilde{d}(k)$ have to be cut off in order to avoid estimation errors.

Hence, we have shown that the complex factor can be estimated blindly except of an ambiguity of π, due to the transmitted *BASK (Binary Amplitude Shift Keying)* symbols $d(k) \in \{-1, 1\}$. In order to estimate the absolute phase of α_ξ, we need at least one reference symbol (e.g. the tail bits of each burst) or exert differential methods in the receiver.

4 Channel Coding and Interleaving

Channel coding plays a decisive role in protecting bits during transmission. Especially data transmission requires very low bit error rates in order to keep re-transmission within certain limits and avoid unnecessary high system loads. The principle of channel coding is that it introduces redundancy into the source data calculated from the source information. Exploiting this additional redundancy for error detection or correction, the channel decoder is able to reconstruct the source information from the output of the equalizer. This operation is much more efficient when the equalizer indicates an a-priori error likelihood for each bit, called *soft output* (refer to Fig. 1 and [9]).

Due to TV multipath propagation, time-selective fading often causes the signal to fall below the noise level. This may result in a large number of consecutive bit errors within one burst. These error clusters are not usually corrected by codes designed for statistically independent errors. In order to suppress the occurrence of such burst errors, *interleaving* in the transmitter scrambles the bits of each encoded block into different bursts. In the receiver, deinterleaving re-scrambles the bits of these bursts in order to rebuild the block structure.

The GSM system applies different interleaving and coding schemes to different transmission modes. In the following, we will outline the schemes assigned to full rate data transmission, where each input block of the TCH/F9.6 consists of 240 bits containing 16 header, 200 data, and 24 FCS (frame check sequence) bits. Hence, the input rate amounts to 240 bits/20 ms = 12 kbit/s. A half-rate convolutional code then applies two convolutions whose polynomials are respectively $D^4 + D^3 + 1$ and $D^4 + D^3 + D + 1$. Since convolutional codes are adapted to infinite sequences, 4 appended tail bits (set to "0") enable a termination of the finite input block. After puncturing, each output block consists of 456 coded bits interleaved unequally over 22 bursts [7].

Depending on the application of non-blind or blind channel estimation algorithms, we have seen from section 3 that the burst structure differs for both cases. This fact also influences the block structure which makes necessary some modifications to the coding and interleaving schemes in order to achieve a fair comparison between non-blind and blind channel estimation approaches. Therefore, Figs. 4 and 5 illustrate our modifications respectively to the channel encoder and interleaver for the non-blind case.

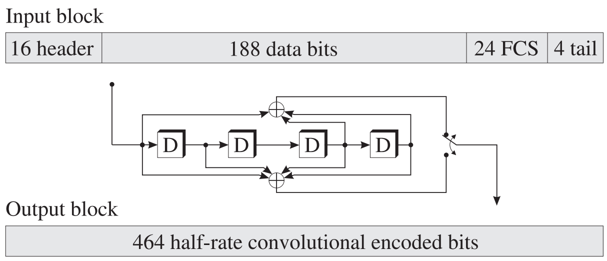

Figure 4: Modified "non-blind" channel coding scheme

Differing from the GSM *Specifications*, each input block contains 228 bits $b(k)$ reducing the input rate to 11.4 kbit/s. Thus, after half-rate encoding an output block consists of

464 non-punctured bits. Finally, a block interleaver scrambles the bits of 5 consecutive blocks over 20 bursts in the manner depicted in Fig. 5.

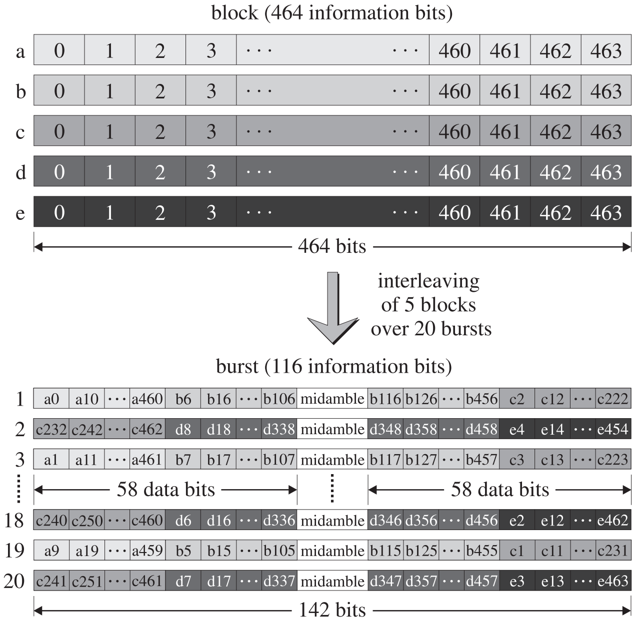

Figure 5: Modified "non-blind" interleaving scheme

If the impulse response of the transmission channel is estimated blindly, each burst consists of 142 data bits $d(k)$. In order to apply the same modified interleaver structure as in the non-blind case, an output block must therefore contain 568 encoded bits. Hence, we have to transmit 280 uncoded bits $b(k)$ per input block within the same period of time which results in an increased input rate of $14\,\mathrm{kbit/s}$. Regarding the energy E_b of an uncoded bit, this also leads to an E_b/N_0-gain of $10 \cdot \log(280/228) \approx 0.9$ dB of blind approaches relative to non-blind ones, where N_0 characterizes the power spectral density of the real valued baseband AWGN $n(t)$.

5 Simulation Results

Referring to Fig. 1 and the modifications to the full rate data transmission mode TCH/F9.6 described in the previous section, a block of 232 (284) bits $b(k)$ is channel encoded and block interleaved into bursts of 116 (142) data bits $d(k)$, where the numbers in front of and within the brackets refer to the non-blind and blind case, respectively. Together with the midamble[2] $f(k)$, a burst of 156 bits is differentially encoded, modulated, and then propagated through each slice $c(\tau, t_{8\xi})$ of the respective sample channel[3] derived from the stochastic GAUSSIAN STATIONARY UNCORRELATED SCATTERING (GSUS) model [6]. We use standard COST-207 *Typical Urban* (TU), *Bad Urban* (BU), and *Hilly Terrain* (HT) profiles, where we take the channel's time-variance *within each burst period* into account. According to a given $\bar{E}_b/N_0$ and signal-to-interference ratio (SIR), AWGN $n(t)$ and CCI $i(t)$ are added to the signal. Upon receive filtering, symbol-rate sampling, and demodulation 148 samples of $y(k)$ are obtained per burst, where we assume perfect synchronization. Then, the non-blind LS approach as well as the blind EVI, WS, and DL-TXK channel estimation schemes are applied to $y(k)$, where the former exploits the training sequences while the latter corrected by the BECO method[4] use the central 142 samples of each demodulated burst. The resulting channel estimates $\hat{h}(\kappa, \xi)$ are then passed to the Max-Log-MAP equalizer delivering the hard output $\hat{d}(k) \in \{-1, 1\}$ as well as the reliability information for soft-decision decoding. After deinterleaving, the channel decoder at last provides the output sequence $\hat{b}(k) \in \{0, 1\}$.

In the frame of MONTE-CARLO simulations, this procedure was executed for all channel slices $c(\tau, t_{8\xi}), 1 \leq \xi \leq 4000$, and $\bar{E}_b/N_0$ values ranging from 0 to 15 dB. In order to examine the influence of non-Gaussian CCI, we selected one co-channel interferer with an SIR set to a fixed value of 10 dB which seems to be realistic in a typical cellular GSM system [7]. Finally, bit error rates (BERs) are calculated before and after channel decoding respectively from bursts and blocks transmitted at a given $\bar{E}_b/N_0$ and SIR.

Table 1 depicts the line styles attributed to the estimation algorithms. For comparison, "ideal non-blind channel estimates" are obtained when the composite channel is perfectly known: $\hat{h}(\kappa, k_{8\xi}) = h(\kappa, k_{8\xi})$.

line style	associated channel est. method
∗—∗	blind DL-TXK method (SOS)
▽—▽	blind WS approach (HOS)
□—□	blind EVI algorithm (HOS)
×—×	non-blind LS scheme
o—o	"ideal non-blind estimation"

Table 1: Line styles used in Figs. 6 and 7

Figure 6: Here, we display the BERs for the TU, BU, and HT sample channels with a maximum Doppler shift set to $f_{\mathrm{D,max}} = 50\,\mathrm{Hz}$. Assuming a carrier frequency of 950 MHz, the velocity of the mobile unit is approximately $60\,\mathrm{km/h}$. The upper row of subplots (a-c) shows the performance of the non-blind and blind channel estimation algorithms based on a known complex factor which is determined by fitting the estimated impulse response to $h(\kappa, k_{8\xi})$ in the least squares sense. For reasons of equality, this is applied to the non-blind LS scheme, too. On the other hand, the bottom row of subplots (d-f) shows the algorithms' performances when the blind channel estimates are corrected by the BECO method. Furthermore, the dashed and solid lines are associated with BERs calculated from the received

[2] which is only required for non-blind channel estimation

[3] Since at most each 8th burst is sent from/to the same mobile station, just each 8th slice $c(\tau, t_{8\xi})$, with $t_{8\xi} = (8 \cdot 156.25)\xi T$, is used.

[4] under assumption of a known noise power γ_ξ

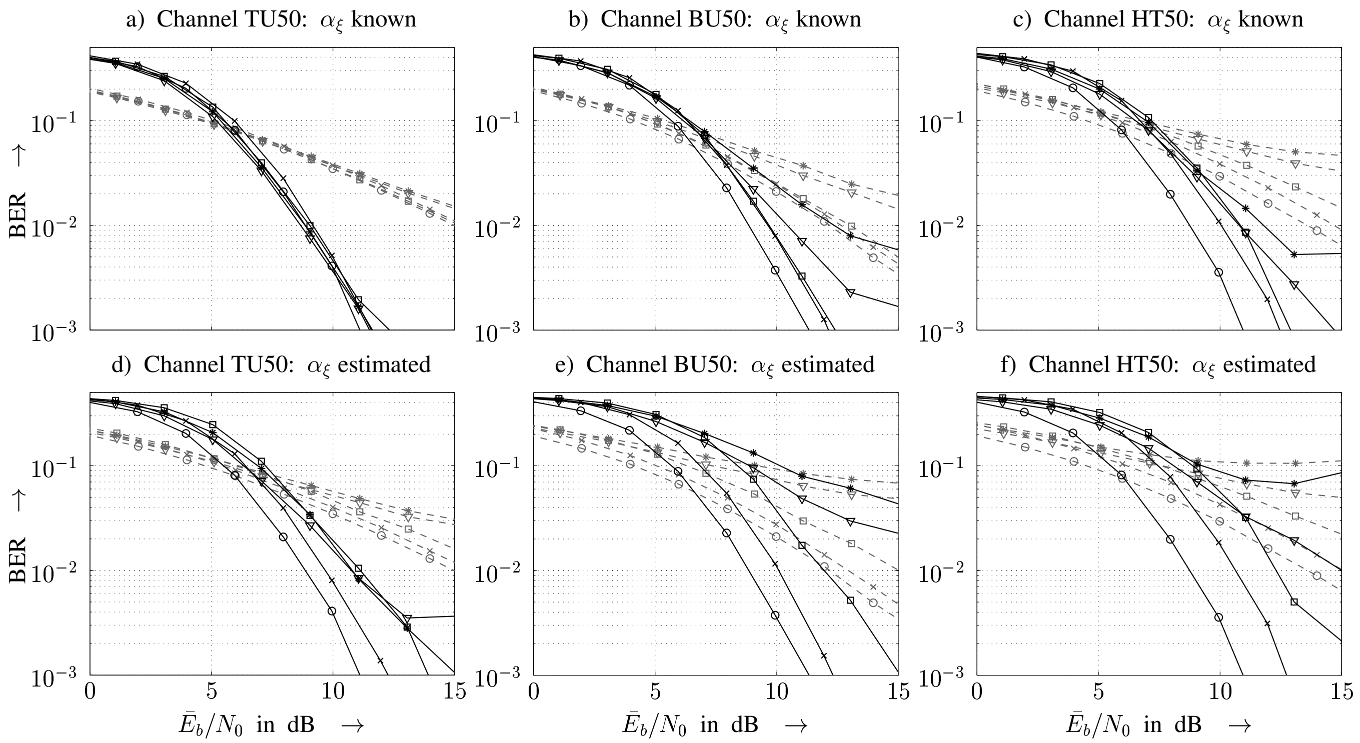

Figure 6: BER in terms of $\bar{E}_b/N_0$ for the estimates of three COST-207 channels with a maximum Doppler shift of $f_{\mathrm{D,max}} = 50$ Hz considering a known (a-c) and estimated complex factor α_ξ (d-f). The dashed and solid lines are associated with the sequences before and after channel decoding, respectively.

bursts and blocks, respectively, i.e. before and after deinterleaving and channel decoding.

Let us first evaluate the impact of channel coding in general by comparing the dashed and solid lines of all subplots. As long as the BERs pertaining to the non-decoded bursts are above a level of approx. $8 \cdot 10^{-2}$, i.e. for low $\bar{E}_b/N_0$ values, channel coding results in a reduced performance. Below this level, the combination of interleaving and channel coding leads to significantly improved BERs [9].

Assuming a known complex factor α_ξ, we realize from subplots a-c that the blind EVI approach and the non-blind LS method deliver nearly similar channel estimates. Averaged over the three channel profiles TU, BU, and HT, EVI requires just about 0.1 dB more in $\bar{E}_b/N_0$ than LS. Although in the noisy case ($\bar{E}_b/N_0 < 5$ dB), the BER values of WS and DL-TXK are lower than those of EVI, for higher $\bar{E}_b/N_0$ rates, WS's and DL-TXK's modest estimation qualities lead to a significant deterioration in BER. This behaviour may be caused by the sensitivity of WS and TXK in estimating long channel impulse responses[5] from a small quantity of symbols. Additionally, the estimation performance of DL-TXK could be impaired by the appearance of "singular" or "quasi-singular" channels.

From subplots d-f, it is obvious that all blind channel estimates suffer from an estimation of the complex factor α_ξ using the BECO scheme, since after decoding the *averaged* $\bar{E}_b/N_0$ *loss* between EVI and LS raises to almost 2 dB. But, remember that only this method ensures absolute blind channel estimates including both the channel's magnitude *and* phase information required for correct equalization.

Figure 7: CCI appears just as an additional source of additive noise to non-blind approaches, whereas blind algorithms cannot distinguish between the interfering and the wanted signal, because both reveal the same statistical properties. Therefore, we investigate the behaviour of blind (EVI) and non-blind (LS) channel estimation schemes after channel decoding for slow (subplots a-c) and fast fading channels (subplots d-f), if a non-Gaussian CCI is superimposed. Furthermore, we assume the complex factor α_ξ to be known within the receiver. The solid lines characterize the estimates in case of an additional CCI with one interferer at SIR $= 10$ dB while the dashed lines represent the estimation results without CCI. Because of their modest estimation quality at higher $\bar{E}_b/N_0$ rates, WS and DL-TXK are omitted in order to enhance clarity.

Irrespective of the channel estimation scheme and the relative velocity between base and mobile station, we can see from all subplots that the additional CCI leads to an increased error floor. By comparing the dashed and solid lines for slow (subplots a-c) and fast fading channels (subplots d-f), we realize that CCI results in a small additional $\bar{E}_b/N_0$ loss of approx. 0.5 dB in both cases. From this, we conclude that it is due to the attenuation that the effect of interference

[5] The number of channel coefficients to be estimated is determined prior to a simulation from the effective length of the power delay spectrum of the GSUS sample channel. We use 3, 4 and 6 coefficients, respectively for the TU, BU, and HT sample channel profiles.

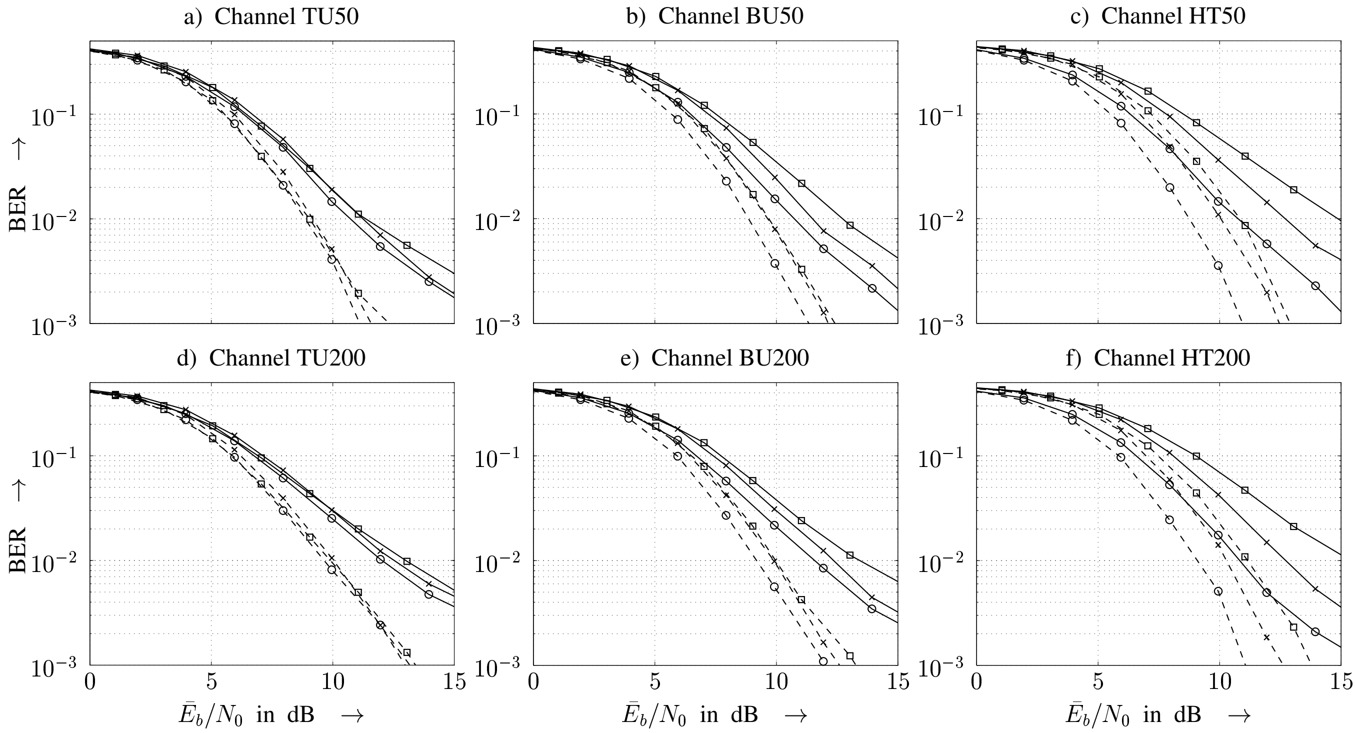

Figure 7: BER in terms of $\bar{E}_b/N_0$ after channel decoding for the estimates of three COST-207 channels with maximum Doppler shifts of $f_{\mathrm{D,max}} = 50$ Hz (a-c) and 200 Hz (d-f). The solid lines characterize the estimates in case of an additional CCI with one interferer at SIR $= 10$ dB while the dashed lines represent the estimation results without CCI.

between EVI and LS is nearly negligible.

6 Conclusion

In this paper, we have compared non-blind and blind channel estimation approaches for the GSM full rate transmission mode TCH/F9.6 under channel coding conditions. In order to facilitate a fair simulation environment, we have slightly modified the channel coding and interleaving schemes with regard to the utilized block sizes. Furthermore, we have presented the new BECO algorithm which overcomes the severe problem of interfered blind channel estimates caused by an unknown complex factor. We have shown that it is possible to use blind channel estimators (especially the HOS based EVI) for MLSE in a GSM receiver, even if the disturbance is non-Gaussian. Let us finally emphasize that we have selected the GSM standard, although we are well aware of the fact that, on account of the GSM *Specifications*, the training sequences are quite unlikely to be banned from the GSM "normal" burst. The reason for choosing GSM anyway is that, with a burst duration of 142 symbols, it is very demanding from the channel estimation point of view.

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