Blind Channel Estimation in GSM Receivers: A Comparison of HOS and SOCS Based Approaches

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Abstract

Due to the solution of an eigenvalue problem, a common drawback of all blind estimation algorithms is the interference of the estimated channel impulse response caused by an unknown complex factor. Since Maximum Likelihood Sequence Estimation (MLSE) requires the knowledge of the exact channel characteristics, we present in this paper a method based on a linear Mean Square Error (MSE) solution which enables the estimation of this factor using a-priori information of the source signal. On the assumption of COST-207[1] *mobile radio channels, we compare the blind channel estimation approaches with the non-blind cross correlation method (CC) used in state-of-the-art GSM receivers (Global System for Mobile communications) and give the bit error rates (BER) after Viterbi detection in terms of the signal-to-noise ratio (SNR). Finally, we also study the influence of non-Gaussian* co-channel interference (CCI) *on the blind and non-blind estimation schemes.*

1 Introduction

In wireless mobile communication systems, the application of blind channel estimation approaches whether they are based on Higher Order Statistics (HOS) or Second Order Statistics (SOS) has been examined very rarely. In 1997, we have shown that some blind HOS approaches yield a good performance in estimating GSM channels [3]. In this paper, we expand our studies on the applicability of SOS based methods. Therefore, we examine the utilization of a new method published by Ding and Li [5], which generates channel diversity in a GSM receiver from a single symbol-rate sampled sequence in contrast to the Second Order Cyclostationary Statistics (SOCS) procedure [11].

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Fig. 1 depicts the equivalent baseband representation of the physical layer of a GSM communication system [10], where channel coding is omitted in order to enhance clarity. In the transmitter, coded information bits $d(k)$ and reference bits $f(k)$ are assembled into differentially encoded *bursts* of $N_b = 142$ bits valued from $\{-1, 1\}$. Then, each burst is modulated by Gaussian Minimum Shift Keying (GMSK) and transmitted over the channel.

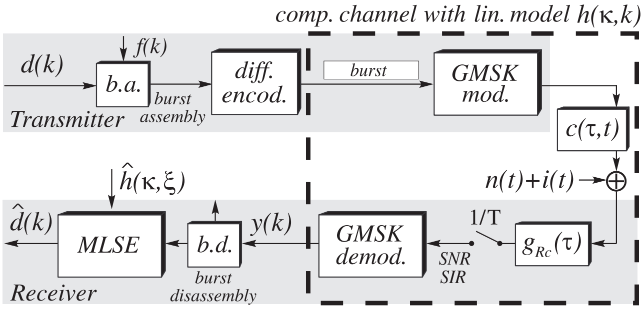

Figure 1. GSM communication system

In a mobile scenario, the physical multipath radio channel is *time-variant (TV)* with a baseband channel impulse response $c(\tau, t)$ derived from the stochastic Gaussian Stationary Uncorrelated Scattering (GSUS) model [7]. We use standard COST-207 *Typical Urban (TU)*, *Bad Urban (BU)*, and *Hilly Terrain (HT)* profiles. Apart from linear TV distortions, additive white Gaussian noise $n(t)$ is present. Moreover, in cellular systems such as GSM, transmitters in remote cells may use the same carrier frequency, which results in an additional interfering signal $i(t)$, called CCI. In the receiver, a Butterworth 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 piecewise stationary sequence $y(k)$.

Maximum Likelihood Sequence Estimation (MLSE) represents a well known procedure to remove intersymbol interference (ISI) from a received digital communication signal such as $y(k)$. However, it assumes the "composite channel" (see Figs. 1 and 4)

$$h(\kappa, k) = j^{-\kappa} \cdot g_0(\kappa, k) \tag{1}$$

to be linear with a finite impulse response, which is complied under consideration of the linear model approximation

$$g_0(\kappa, k) \triangleq [c_0(\tau) * c(\tau, t) * g_{Rc}(\tau)]_{\tau=\kappa T,\, t=kT}. \tag{2}$$

In eq. (1), the factor $j^{-\kappa}$ takes into account the rotation in the transmitter and the derotation in the demodulator. The $c_0(\tau)$ impulse describes the Laurent approximation [9] of the (non-linear) GMSK modulator and the asterisk "$*$" denotes convolution. For any value of k, $h(\kappa, k)$ is limited to the range $\mathcal{K}$ of "relevant" indices κ in order to obtain a (TV) FIR model. Since MLSE requires the knowledge of $h(\kappa, k)$ in order to equalize a block of the demodulated sequence $y(k)$, the following section is focused on the 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[2].

2 Blind and non-blind channel estimation

According to Fig. 2, each GSM normal burst contains a training sequence $f(k)$ 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 from the cross-correlation (CC) between the received (corrupted) signal after burst disassembling (see Fig. 1) and the known training sequence $f(k)$. 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*.

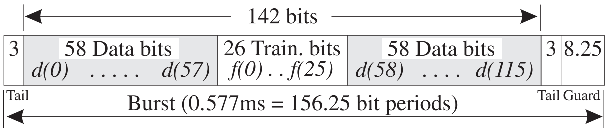

Figure 2. GSM "normal" burst

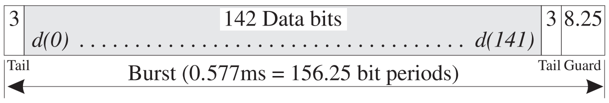

Figure 3. GSM "blind" burst

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. 3 shows the structure of a "blind" GSM burst which consists of 142 data symbols.

We have selected the following blind HOS and SOS based channel estimation algorithms for an analyzes in this paper:

- EIGENVECTOR APPROACH TO BLIND IDENTIFICATION (EVI) by Boss, Jelonnek, and Kammeyer [1, 8],
- W-SLICE (WS) algorithm by Fonollosa and Vidal [6],
- TXK method by Tong, Xu and Kailath [11].

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. TXK, finally, is one of the first subspace based methods exploiting SOCS for fractionally sampled channel identification. On the other hand, this SOS based approach can be applied to a new method of Ding & Li explained in the following section. Remember that all blind channel estimates have to be obtained from only 142 samples of the received sequence $y(k)$ (see Figs. 1, 3, and 5).

2.1 Exploiting channel diversity in a GSM system

The application of SOS methods[3] to blind channel estimation requires the exploitation of channel diversity, which can be achieved in several ways. If the sampling frequency is a fraction of T (characterized by the M fold oversampling in Fig. 5), SOCS based approaches can be applied in principle presumed that the transmitting filter delivers sufficient excess bandwidth. Alternatively, it is possible to exploit channel diversity by using additional antennas (in the uplink, e.g.), which requires extra radio frequency (RF) receivers and significantly increases hardware costs and computational efforts. Consequently, utilizing HOS to the symbol-rate sampled (quasi) stationary demodulated sequence $y(k)$ seems to be the only way to estimate the channel blindly without any additional hardware effort.

In [5] Ding & Li have maintained that the application of SOCS will not generate the necessary channel diversity due to the lack of excess bandwidth in the $c_0(\tau)$ impulse of the GMSK Laurent approximation. Therefore, they have shown that the generation of channel diversity can also be achieved if the GSM derotation scheme is used in order to create two channel outputs from a single symbol-rate sampled GMSK signal. This procedure will shortly be named as DL. Referring to eq. (2) and Fig. 4, where noise and symbol-

[2] 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.

[3] and some HOS algorithms [4]

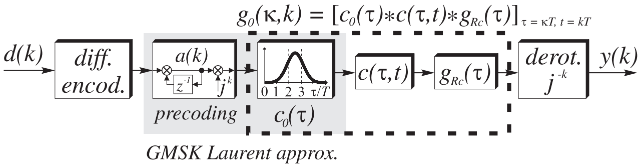

Figure 4. GMSK Laurent approximation

rate sampling are omitted in order to enhance clarity, we obtain the demodulated (derotated) sequence

$$y(k) = j^{-k} \cdot \sum_{\kappa=-\infty}^{\infty} g_0(\kappa,k) \cdot a(k-\kappa) \cdot j^{k-\kappa} \quad (3)$$

$$= \sum_{\kappa=-\infty}^{\infty} \underbrace{[g_0(\kappa,k) \cdot j^{-\kappa}]}_{h(\kappa,k)} \cdot a(k-\kappa) \quad (4)$$

with the differentially encoded and precoded burst sequence $a(k) \in \{-1, 1\}$. Since $a(k)$ is binary real, two independent sub-channel outputs

$$\begin{bmatrix} Re\{\mathbf{y}\} \\ Im\{\mathbf{y}\} \end{bmatrix} = \begin{bmatrix} Re\{\mathbf{H}\} \\ Im\{\mathbf{H}\} \end{bmatrix} \cdot \mathbf{a} \quad (5)$$

can be generated, where $\mathbf{a}$ is defined as

$$\mathbf{a} \triangleq [a(k), \cdots, a(k-q-N_b+1)]^T \quad (6)$$

and $\mathbf{H}$ is the $N_b \times (q+N_b)$ channel Toeplitz matrix containing the TV FIR channel $\mathbf{h} = [h(0,k), \cdots, h(q,k)]$ of order q. Note that $\mathbf{H}$ will have no full column rank if the two sub-channels share common zeros. These so-called "singular" channels *cannot* be identified, which is a problem similar to the application of SOCS.

As we do not agree with the statement of Ding & Li that the $c_0(\tau)$ impulse has zero-excess bandwidth, we also examine the usability of the SOCS based TXK method besides the DL based one.

2.2 Blind channel estimation correction

In order to achieve an ideal MLSE, the Viterbi algorithm is dependent upon the complete channel characteristics which include correct magnitude and phase information. However, due to the solution of an eigenvalue problem, a common drawback of all existing blind approaches is the interference of the estimated channel impulse response $\tilde{h}(\kappa,\xi)$ by an unknown complex factor α_ξ (see Fig. 5).

The fundamental idea of blind channel estimation is the avoidance of any midamble. Hence, we neither can determine α_ξ using reference symbols. Fig. 5 illustrates the BLIND CHANNEL ESTIMATION CORRECTION (BECO) scheme, where $y(i)$ will be a fractionally sampled sequence ($iT = M \cdot kT, M > 1$), if the blind channel estimation is based on SOCS. Note that in this case $y(i)$ has been demodulated using a fractionally tap-spaced derotation scheme $j^{-k/M}$. After downsampling (which is necessary for SOCS

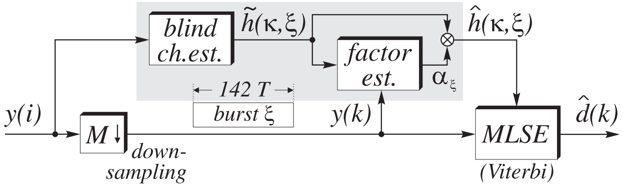

Figure 5. Blind channel estimation correction

only), the demodulated sequence $y(k)$ as well as $\tilde{h}(\kappa,\xi)$ are fed to the factor estimation block. Based upon the MSE equalizer

$$\mathbf{e}_\xi = (\tilde{\mathbf{H}}_\xi^* \tilde{\mathbf{H}}_\xi + \gamma_\xi \mathbf{I})^{-1} \tilde{\mathbf{H}}_\xi^* \mathbf{i}_\xi, \quad (7)$$

where

$$\mathbf{e}_\xi \triangleq [e_\xi(0), \cdots, e_\xi(n)]^T \quad (8)$$

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

$$\tilde{\mathbf{H}}_\xi \triangleq \begin{bmatrix} \tilde{h}(0,\xi) & 0 & \cdots & 0 \\ \vdots & \ddots & \ddots & \vdots \\ \tilde{h}(q,\xi) & \cdots & \tilde{h}(0,\xi) & 0 \\ 0 & \tilde{h}(q,\xi) & \cdots & \tilde{h}(0,\xi) \\ \vdots & \ddots & \ddots & \vdots \\ 0 & \cdots & 0 & \tilde{h}(q,\xi) \end{bmatrix}, \quad (9)$$

we are able to estimate α_ξ. In case of 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 $\tilde{\mathbf{H}}_\xi^* \tilde{\mathbf{H}}_\xi$ in order to avoid a zero-forcing solution, which will be affected adversely by noise. The utilization of (7) to $y(k)$ will lead us to the linear equalized sequence $\tilde{d}_\xi(k) = e_\xi(k) * y(k)$ 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

$$\alpha_\xi = \sqrt{\frac{1}{N_b} \sum_k |\tilde{d}_\xi(k)|^2} \cdot e^{\frac{j}{2} \cdot \arg\left\{\frac{1}{N_b} \sum_k \tilde{d}_\xi^2(k)\right\}}, \quad (10)$$

which yields the corrected blind channel estimation $\hat{h}(\kappa,\xi) = \alpha_\xi \cdot \tilde{h}(\kappa,\xi)$. Note that the transients of $\tilde{d}_\xi(k)$ 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 Ampl. Shift Keying)* symbols

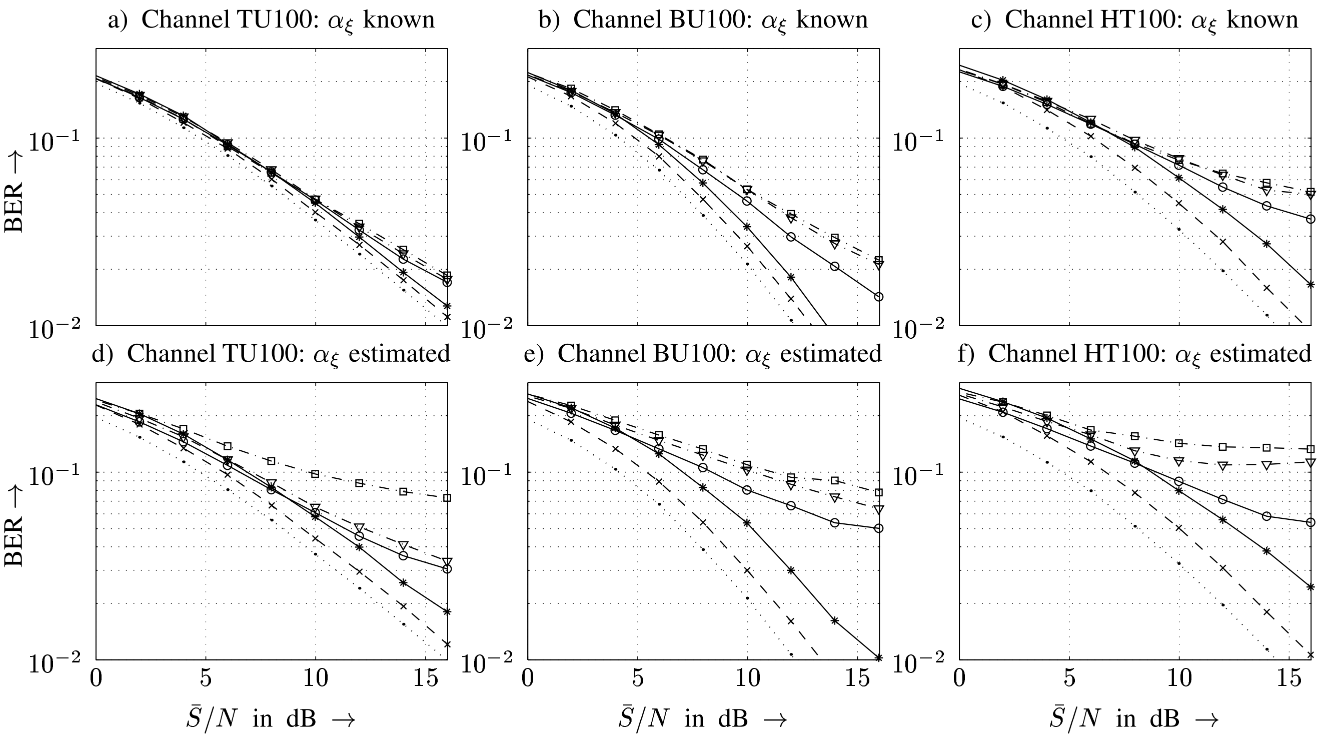

Figure 6. BER in terms of SNR for the estimates of three COST-207 channels with a maximum Doppler shift of $f_{\mathrm{D,max}} = 100\ \mathrm{Hz}$ considering a known (a-c) and estimated complex factor α_ξ (d-f).

$d(k) \in \{-1, 1\}$. 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.

3 Simulation results

Referring to Fig. 1, a burst of 156 BASK symbols $d(k)$ is encoded, modulated, and then propagated through each slice $c(\tau, t_{8\xi})$ of the respective sample channel[4]. Additive Gaussian noise $n(t)$ and CCI $i(t)$, both colored by the receiving filter $g_{Rc}(\tau)$, are added according to a given SNR and SIR, respectively. Upon symbol-rate sampling and demodulation, 148 samples of $y(k)$ are obtained. Then, the non-blind and blind GSM channel estimation schemes are applied to $y(k)$, where the non-blind CC method utilizes the training midamble of 26 bits while the blind algorithms use 142 samples of $y(k)$. Note that in case of SOCS based blind approaches, the received sequence $y(i)$ is fractionally tap-spaced (see Fig. 5) and consists of $M \cdot 142$ samples. After applying the BECO method to the blind channel estimates, $\hat{h}(\kappa, k_{8\xi})$ is passed to the Viterbi detector.

In the frame of Monte-Carlo runs, this procedure was executed for all channel slices $c(\tau, t_{8\xi})$, $1 \le \xi \le 2000$, and SNR values ranging from 0 to 16 dB. In order to examine the influence of non-Gaussian CCI, we selected three co-channel interferers with an SIR set to a fixed value of 10 dB which seems to be realistic in a typical cellular GSM system. Finally, the bit error rate (BER) was calculated from all bursts transmitted at a given SNR and SIR.

Table 1 depicts the line styles attributed to the respective algorithm, where the TXK method is based on SOCS as well as on DL. For comparison, the dotted line is obtained when using ideal channel estimates $\hat{h}(\kappa, k_{8\xi}) = h(\kappa, k_{8\xi})$.

line style	associated channel est. method
□ - · - □	blind TXK method based on SOCS
▽ - · - ▽	blind TXK method based on DL
o —— o	blind WS approach (HOS)
* —— *	blind EVI algorithm (HOS)
× - - - ×	non-blind CC scheme
• ········ •	"ideal non-blind estimation"

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

Figure 6 displays BER after MLSE in terms of SNR for the *TU* (a, d), *BU* (b, e), and *HT* sample channel (c, f). The maximum Doppler shift was set to $f_{D,max} = 100$ Hz. Assuming a carrier frequency of 950 MHz, the velocity of the mobile unit is approx. 115 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 com-

[4]Since at most each 8th burst is sent from/to the same mobile station, just each 8th slice $c(\tau, t_{8\xi})$ is used for simulation.

plex factor, which is determined by fitting the estimated impulse response $\hat{h}(\kappa, k_{8\xi})$ to $h(\kappa, k_{8\xi})$ in the least squares (LS) sense. For reasons of equality, this is also applied to the non-blind CC method. On the other hand, the bottom row (d-f) displays the algorithms' performance when the blind channel estimates are corrected by the BECO scheme.

We realize that CC delivers the best channel estimates and EVI outperforms all blind approaches. Although in the noisy case (SNR < 5 dB) , the BER values of WS and TXK are lower than those of EVI, at higher values of SNR, WS's and TXK's modest estimation qualities lead to a significant degradation in BER. This behaviour could be due to the sensitivity of WS and TXK to the small quantity of symbols used for blind channel estimation. Furthermore, the estimation performance of TXK could be impaired by the appearance of "singular" channels. For a known complex factor α_ξ (a-c), EVI requires about 0.3 dB *(TU)*, 0.7 dB *(BU)*, and 1.7 dB *(HT)* more in SNR than the non-blind CC scheme. This corresponds to an *everaged loss* of about 0.9 dB, while it raises to almost 2 dB, if α_ξ has been estimated (d-f). Hence, the estimation of α_ξ leads to an additional SNR loss of about 1.1 dB.

Although the performance of the SOCS based TXK method is a little worse than the DL based one (regarding Fig. 6a-c), we can state that in contrast to Ding & Li's opinion the $c_0(\tau)$ impulse delivers enough excess bandwidth. In our simulations the oversampling was set to $M = 4$.

Figure 7: Using the same *BU* channel as in Fig. 6e, we examine the impact of three co-channel interferers on the BER performance at SIR = 10 dB. As expected, there is a

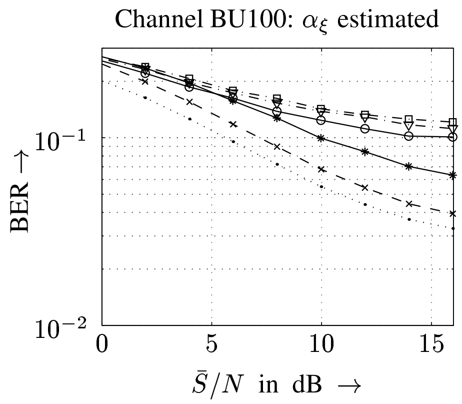

Figure 7. BER in terms of SNR for the estimates of a BU channel ($f_{\mathrm{D,max}} = 100\,\mathrm{Hz}$) with three interferers at SIR $= 10\,\mathrm{dB}$

degradation for all algorithms due to the higher noise level. In [2] we have shown that blind HOS estimation schemes do not suffer from a less Gaussian interference. Here, we see that this also seems to be valid for the SOS based approaches, even if α_ξ has been estimated.

4 Conclusions

In this paper, we have presented a new algorithm based on an MSE solution, which overcomes the fundamental problem of interfered blind channel estimates by an unknown complex factor. Furthermore, 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. Although the DL method is an excellent idea creating channel diversity in a GSM receiver, we finally have proven that the GMSK modulator delivers sufficient excess bandwidth for the application of SOCS.

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