# Efficient Receivers for GSM MUROS Downlink Transmission


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***Abstract*—Currently, Multiple Users Reusing One Slot (MUROS) is discussed in 3GPP GERAN as an extension of the GSM standard. In MUROS, two overlaid GMSK signals are transmitted in the same time slot and at the same frequency resource. By this, capacity of existing GSM networks in principle can be doubled and up to four half rate voice users can share one time slot. In this paper, channel estimation and detection is investigated for the MUROS downlink. Two novel channel estimation algorithms are presented, taking into account the specifics of the MUROS downlink. For detection, a joint MLSE of both user signals can be applied in case of noise limited scenarios. For interference limited environments, it turns out that approaches based on the mono interference cancellation (MIC) algorithm for single antenna interference cancellation (SAIC) are more favorable. It is shown that the standard MIC algorithm performs sufficiently well and could be used for a fast introduction of MUROS in existing GSM networks. For enhanced performance, a novel algorithm based on MIC along with successive interference cancellation is proposed. The presented results demonstrate that an even better performance as the GSM reference performance before introduction of SAIC can be obtained for MUROS if well–designed receivers are used.**


## I. Introduction

The Global System for Mobile communications (GSM) is still by far the most popular cellular communication system in the world. Every year, there is a significant increase in the number of GSM users. According to operators, the growth can be considered as especially dramatic in emerging markets such as Asian and African countries. Because in particular voice services are demanded, there is the need for a major voice capacity enhancement within the GSM system in order to meet the demands of the customers.

In recent years, the capacity and transmission quality of GSM has been already significantly improved. For this, Enhanced Data Rates for GSM Evolution (EDGE) has been introduced which employs 8–ary phase–shift keying (8PSK) modulation in addition to Gaussian minimum–shift keying (GMSK) which has been adopted for the original GSM system. Furthermore, Enhanced GPRS phase 2 (EGPRS2) is currently standardized where 16–ary and 32–ary quadrature amplitude modulation (16QAM and 32QAM), respectively, are used together with turbo coding in order to further increase the data rates.

Another approach for improving the spectral efficiency of GSM is to employ a tighter frequency reuse. This, however, increases adjacent and cochannel interference from other users within the system. As a consequence, the performance of a conventional GSM equalizer based on maximum–likelihood sequence estimation (MLSE), the BCJR algorithm [1] or the soft–output Viterbi algorithm (SOVA) [2], respectively, is no longer sufficiently good. Interference suppression techniques are required in order to avoid a performance degradation for small frequency reuse factors. To this end, single antenna interference cancellation (SAIC) algorithms have been developed, e.g. [3]–[9], exploiting the special properties of the GMSK modulation which can be well approximated by filtered binary phase-shift keying (BPSK) modulation. SAIC algorithms are highly beneficial especially for downlink transmission because only a single receive antenna is required for interference suppression; at the mobile station, most often no antenna diversity is incorporated. Capacity improvements due to SAIC of up to 50 % have been reported in field trials [10]. In 3GPP TSG GERAN, corresponding performance requirements have been specified (Downlink Advanced Receiver Performance (DARP) phase I), and commercial mobile stations with SAIC capability are now widely available.

Mobile terminals of low cost are required for emerging markets. Therefore, SAIC seems to be a more promising avenue in this case than voice transmission with higher order modulation such as 8PSK or 16QAM which implies the necessity of highly linear amplifiers and complex equalizers. As a consequence, Multiple Users Reusing One Slot (MUROS) and Voice Services over Adaptive Multiuser Orthogonal Subchannels (VAMOS), respectively, are currently considered in 3GPP TSG GERAN, cf. [11], [12]. In MUROS/VAMOS, capacity is increased by deliberately overlaying two users in the same time slot and at the same frequency resource within a cell. By this, capacity can be almost doubled with an only small loss in performance compared to the single user case if the users can be separated. MUROS/VAMOS can be applied to data as well as full and half rate voice transmission. While in the uplink, cf. [13], a 2×2 virtual multiple–input multiple–output (MIMO) system with different subchannels results, if the base station has two sufficiently separated receive antennas, the situation is more difficult in the downlink. Here, only a single receive antenna can be assumed for both involved mobile terminals, and for each of the mobile terminals the two transmit signals emitted by the base station travel through the same propagation

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channel. In order to enable a sufficiently good user separation, the Orthogonal Subchannels (OSC) concept has been proposed for the downlink of MUROS as a signaling scheme among other approaches [11].

In this paper, we consider efficient receivers for a MUROS downlink OSC transmission. In Section II, the corresponding system model is introduced. In Section III, approaches for channel estimation are discussed. Here, it should be noted that the channel estimation problem of MUROS differs significantly from that of a conventional GSM transmission, requiring novel algorithms. Several different approaches for equalization and interference cancellation are presented in Section IV. Finally, simulation results are discussed in Section V, and some conclusions are drawn in Section VI.

## II. System Model

In the considered scenario of a MUROS downlink OSC transmission, the base station transmits two user signals in the same time slot and at the same frequency resource, where the second GMSK signal is rotated by 90° according to the OSC concept [11]. Hence, after GMSK derotation at the receiver, the discrete–time received signal at one of the two mobile terminals in the equivalent complex baseband can be written as

$$r[k]=\sum_{\kappa=0}^{q_h} h[\kappa]\, a_1[k-\kappa] + j\, b\sum_{\kappa=0}^{q_h} h[\kappa]\, a_2[k-\kappa] + n[k] + q[k]. \tag{1}$$

Here, the discrete–time channel impulse response $h[k]$ of order $q_h$ comprises the effects of GMSK modulation, the mobile channel from the base station to the considered user, receiver input filtering, and GMSK derotation at the receiver. For all derivations, $h[k]$ is assumed as constant within a transmission burst but varies randomly between bursts (block fading). $a_1[k]$ and $a_2[k]$ denote the BPSK transmit symbols of both users with variance $\sigma_a^2$ each, and $n[k]$ and $q[k]$ refer to discrete–time white Gaussian noise of variance $\sigma_n^2$ and adjacent and cochannel interference from other cells, respectively. The factor $b>0$ represents the difference in power of both transmit signals introduced by downlink power control at the base station to cope with the near–far problem. There is no explicit signaling of $b$ from the base station to the mobile stations. It is assumed that $a_1[k]$ is the desired signal of the considered mobile station.

## III. Channel Estimation

For channel estimation, it should be taken into account that both user signals in principle propagate through the same channel and the channel impulse response of the user 2 signal is that of the user 1 signal scaled by a factor $b$ (and $j$). If the received symbols corresponding to the time–aligned training sequences of both users are collected in a vector $\boldsymbol{r}$, this vector can be expressed as

$$\boldsymbol{r} = \boldsymbol{A}_1\, \boldsymbol{h} + b\, \boldsymbol{A}_2\, \boldsymbol{h} + \boldsymbol{n} + \boldsymbol{q}, \tag{2}$$

where $\boldsymbol{A}_1$ and $\boldsymbol{A}_2$ represent $(N-q_h)\times(q_h+1)$ Toeplitz convolution matrices corresponding to the training sequences of user 1 and user 2, respectively, with training sequence length $N$, and $\boldsymbol{h}=[h[0]\,h[1]\,\ldots\,h[q_h]]^T$ ($(\cdot)^T$: transposition). $\boldsymbol{n}$ and $\boldsymbol{q}$ are vectors containing the noise and interference contributions, respectively. For simplicity, factor $j$ in (1) has been absorbed in $\boldsymbol{A}_2$. Furthermore, for channel estimation it is assumed that the composite impairment $\boldsymbol{w}=\boldsymbol{n}+\boldsymbol{q}$ is a Gaussian vector with statistically independent entries of variance $\sigma_w^2$ each.

### A. Joint ML Estimation of $\boldsymbol{h}$ and $b$

The joint maximum–likelihood (ML) estimates for $\boldsymbol{h}$ and $b$ result from minimizing the $L_2$–norm of the error vector $\boldsymbol{e}=\boldsymbol{r}-\boldsymbol{A}_1\,\hat{\boldsymbol{h}}-\hat{b}\,\boldsymbol{A}_2\,\hat{\boldsymbol{h}}$, where $\hat{\boldsymbol{h}}$ and $\hat{b}$ denote the estimated quantities. Differentiating $\boldsymbol{e}^H\boldsymbol{e}$ ($(\cdot)^H$: Hermitian transposition) with respect to $\hat{\boldsymbol{h}}^*$ and $\hat{b}$ and setting the derivatives to zero results in the following two conditions for the ML estimates of $\boldsymbol{h}$ and $b$:

$$\hat{\boldsymbol{h}} = \Big(\underbrace{(\boldsymbol{A}_1^H+\hat{b}\,\boldsymbol{A}_2^H)}_{\boldsymbol{V}^H}\underbrace{(\boldsymbol{A}_1+\hat{b}\,\boldsymbol{A}_2)}_{\boldsymbol{V}}\Big)^{-1}\underbrace{(\boldsymbol{A}_1^H+\hat{b}\,\boldsymbol{A}_2^H)}_{\boldsymbol{V}^H}\,\boldsymbol{r} \tag{3}$$

or

$$\hat{\boldsymbol{h}} = \left(\boldsymbol{V}^H\boldsymbol{V}\right)^{-1}\boldsymbol{V}^H\boldsymbol{r}, \tag{4}$$

$$\begin{aligned}\hat{b} \;=\; & \frac{1}{2}\left(\hat{\boldsymbol{h}}^H\,\boldsymbol{A}_2^H\,\boldsymbol{A}_2\,\hat{\boldsymbol{h}}\right)^{-1}\Big(\left(\hat{\boldsymbol{h}}^H\,\boldsymbol{A}_2^H\right)\left(\boldsymbol{r}-\boldsymbol{A}_1\,\hat{\boldsymbol{h}}\right)\\ & +\left(\boldsymbol{r}^H-\hat{\boldsymbol{h}}^H\,\boldsymbol{A}_1^H\right)\left(\boldsymbol{A}_2\,\hat{\boldsymbol{h}}\right)\Big).\end{aligned} \tag{5}$$

Eqs. (4) and (5) may be also viewed as ML channel estimate for given $b$, cf. e.g. [14], and ML estimate of $b$ for given channel vector, respectively. However, it does not seem possible to obtain a closed–form solution for $\hat{\boldsymbol{h}}$ and $\hat{b}$ from the two coupled equations. Thus, a solution might be calculated iteratively by inserting an initial choice for $\hat{b}$ in (4), using the resulting channel vector for refining $\hat{b}$ via (5), etc. until convergence is reached.

### B. Blind Estimation of $b$

In an alternative approach, $b$ is first estimated from the received vector according to an ML criterion, assuming only knowledge of the channel statistics and both training sequences. ML channel estimation for given $\hat{b}$ according to (4) might be performed subsequently.

Assuming $\boldsymbol{h}$ as a complex Gaussian vector with autocorrelation matrix $\boldsymbol{\Phi}_{hh}=\mathcal{E}\{\boldsymbol{h}\,\boldsymbol{h}^H\}$ ($\mathcal{E}\{\cdot\}$: expectation), the pdf of the received vector conditioned on $b$ may be expressed as

$$\mathrm{pdf}(\boldsymbol{r}\,|\,b) = \frac{1}{\pi^M\,\det\left(\boldsymbol{\Phi}_{rr|b}\right)}\exp\left(-\boldsymbol{r}^H\,\boldsymbol{\Phi}_{rr|b}^{-1}\,\boldsymbol{r}\right), \tag{6}$$

where $M=N-q_h$ and $\boldsymbol{\Phi}_{rr|b}=\mathcal{E}\left\{\boldsymbol{r}\,\boldsymbol{r}^H\,|\,b\right\}$,

$$\boldsymbol{\Phi}_{rr|b} = (\boldsymbol{A}_1+b\,\boldsymbol{A}_2)\,\boldsymbol{\Phi}_{hh}\,(\boldsymbol{A}_1+b\,\boldsymbol{A}_2)^H+\sigma_w^2\,\boldsymbol{I}_M. \tag{7}$$

The ML estimate for $b$ can be obtained by maximizing $\ln(\mathrm{pdf}(\boldsymbol{r}\,|\,b))$:

$$\begin{aligned}\hat{b} \;&=\; \underset{\tilde{b}}{\operatorname{argmax}}\left\{-\boldsymbol{r}^H\,\boldsymbol{\Phi}_{rr\,|\,\tilde{b}}^{-1}\,\boldsymbol{r}-\ln\left[\det\left(\boldsymbol{\Phi}_{rr\,|\,\tilde{b}}\right)\right]\right\}\\ &=\; \underset{\tilde{b}}{\operatorname{argmin}}\Big\{\boldsymbol{r}^H\,\Big[\left(\boldsymbol{A}_1+\tilde{b}\,\boldsymbol{A}_2\right)\boldsymbol{\Phi}_{hh}\,\cdot\\ &\qquad\qquad \left(\boldsymbol{A}_1+\tilde{b}\,\boldsymbol{A}_2\right)^H+\sigma_w^2\,\boldsymbol{I}_M\Big]^{-1}\boldsymbol{r}\end{aligned}$$

$$+\ln\left[\det\left(\left(\boldsymbol{A}_1+\tilde{b}\,\boldsymbol{A}_2\right)\,\boldsymbol{\Phi}_{hh}\left(\boldsymbol{A}_1+\tilde{b}\,\boldsymbol{A}_2\right)^H+\sigma_w^2\,\boldsymbol{I}_M\right)\right]\Bigg\}. \quad (8)$$

Minimization of the one–dimensional function in (8) might be performed by a golden section search technique [15].

Simulations have shown that both proposed estimation approaches in principle perform equally well under practically relevant conditions. Thus, the approach which is better suited for a given platform of implementation might be selected by the system designer.

## IV. Approaches for Equalization and Interference Cancellation

### A. Joint MLSE

In noise limited scenarios, joint MLSE of sequences $a_1[\cdot]$ and $a_2[\cdot]$ (or a corresponding SOVA or BCJR algorithm producing soft output) is optimum. For this, a Viterbi algorithm (VA) in a trellis diagram with states

$$\tilde{\boldsymbol{S}}[k]=[\tilde{a}_1[k-1]\,\tilde{a}_2[k-1]\,\ldots\,\tilde{a}_1[k-q_h]\,\tilde{a}_2[k-q_h]] \quad (9)$$

($\tilde{a}_1[\cdot]$, $\tilde{a}_2[\cdot]$: trial symbols of sequence estimator) can be used, where the branch metric of state transitions is given by

$$\lambda[k]=\left|r[k]-\sum_{\kappa=0}^{q_h}\hat{h}[\kappa]\,\tilde{a}_1[k-\kappa]-j\,\hat{b}\sum_{\kappa=0}^{q_h}\hat{h}[\kappa]\,\tilde{a}_2[k-\kappa]\right|^2. \quad (10)$$

Equivalently, an MLSE for the modified 4QAM constellation $\{-1-j\,\hat{b},\,-1+j\,\hat{b},\,+1-j\,\hat{b},\,+1+j\,\hat{b}\}$ can be applied. In both cases, the VA requires $4^{q_h}$ states. The estimates for $h[\kappa]$ and $b$ needed in (10) can be calculated according to Section III.

### B. Mono Interference Cancellation (MIC)

For reconstruction of the sequence of interest, also a standard SAIC algorithm can be employed. Therefore, legacy DARP phase I mobile terminals can be used also for MUROS without any change if legacy training sequences are employed. By a simple pure software update, also the eight new MUROS training sequences [11] can be taken into account in a straightforward way in a mobile terminal with SAIC receiver. Thus, in the following, the MIC algorithm from [6]–[8] is briefly reviewed.

An arbitrary non–zero complex number $c$ is selected and a corresponding number $c^{\perp}=\mathrm{Im}\{c\}-j\,\mathrm{Re}\{c\}$ is generated, where $\mathrm{Im}\{\cdot\}$ and $\mathrm{Re}\{\cdot\}$ denote the imaginary and real part, respectively. $c$ and $c^{\perp}$ may be interpreted as mutually orthogonal two–dimensional vectors. The received signal is first filtered with a complex–valued filter with coefficients $p[\kappa]$ and then projected onto $c$, i.e., the real–valued signal

$$y[k]=\mathcal{P}_c\Big\{\sum_{\kappa=0}^{q_p}p[\kappa]\,r[k-\kappa]\Big\} \quad (11)$$

is formed, where $\mathcal{P}_c\{x\}$ denotes the coefficient of projection of a complex number $x$ onto $c$,

$$\mathcal{P}_c\{x\}=\frac{<x,\,c>}{|c|^2}=\frac{\mathrm{Re}\{x\,c^*\}}{|c|^2} \quad (12)$$

($<\cdot,\cdot>$: inner product of two vectors). It is shown in [8], that the filter impulse response $p[\kappa]$ can be chosen for perfect elimination of $a_2[k]$ (assuming $a_1[k]$ is the desired sequence) if the filter order $q_p$ is sufficiently high. After filtering and projection, $a_1[\cdot]$ can be reconstructed by trellis–based equalization. An adaptive implementation of the MIC algorithm is also described in [8] which requires only knowledge of the training sequence of the desired user but no explicit channel knowledge.

In typical urban (TU) environments, channel snapshots where a single tap dominates arise frequently. Therefore, we consider the case $h[0]\neq 0$, $h[\kappa]=0$, $\kappa\neq 0$ ($q_h=0$). The single effective channel tap $j\,b\,h[0]$ of the second user is rotated by 90° compared to that of the first user. Therefore, in this case orthogonal subchannels result also at the receiver side. According to [8], suppression of the second user is possible without any loss in signal–to–noise ratio (SNR), and $\mathrm{SNR}=2\,|h[0]|^2\,\frac{\sigma_a^2}{\sigma_n^2}$ is valid after MIC if interference from other cells is absent ($q[k]=0$). However, both subchannel contributions are not orthogonal anymore at the receiver side for $q_h>0$, and in general an SNR loss due to filtering and projection cannot be avoided. Hence, if interference from other cells is absent, joint MLSE performs better than MIC which may be viewed as a suboptimum equalizer for QPSK–type signals in this case.

It should be noted that MIC is beneficial also for scenarios with several interferers [8]. Here, the minimum mean–squared error (MMSE) filter found by adaptation is a kind of compromise solution adjusted to the interference mixture. Given this and the fact that interference created by the other user of the same base station is close to orthogonal to the desired user signal in many cases for TU scenarios, it can be expected that MIC performs better than joint MLSE in scenarios with additional interference from other cells (cf. also Fig. 1 in Section V).

### C. MIC Receiver with Successive Interference Cancellation

Because joint MLSE degrades significantly under external interference and DARP phase I receivers such as MIC typically exhibit a good performance only if the signal of the second MUROS user is not much stronger than that of the considered user, more sophisticated schemes might be designed for interference limited scenarios. For this purpose we can exploit the fact that in contrast to the standard SAIC problem, the training sequences corresponding to $a_1[k]$ *and* $a_2[k]$ are known at the mobile station, and both signals are time aligned. Therefore, in principle it is possible to reconstruct $a_1[\cdot]$ and $a_2[\cdot]$ in the same mobile station using two separate MIC algorithms.

In a MIC receiver with successive interference cancellation (SIC), channel estimation according to Section III is performed first. If $\hat{b}\geq b_0$ (e.g. $b_0=1.0$), $a_2[\cdot]$ is reconstructed first by a MIC algorithm and subsequent trellis–based equalization yielding estimates $\hat{a}_2[\cdot]$. In the next step, the contribution of $a_2[\cdot]$ is canceled from the received signal, resulting in the signal

$$r_c[k]=r[k]-j\,\hat{b}\sum_{\kappa=0}^{q_h}\hat{h}[\kappa]\,\hat{a}_2[k-\kappa], \quad (13)$$

which is fed into another MIC and equalization stage in order to reconstruct $a_1[\cdot]$. Because $r_c[k]$ contains no (or much reduced) contributions from $a_2[\cdot]$, interference from other cells can be much better combatted now by the second MIC.

If $\hat{b} < b_0$, only a standard MIC is employed for reconstruction of $a_1[\cdot]$ because successive interference cancellation most likely would suffer from error propagation[1].

In a typical implementation, the complexity of MIC with SIC is about 2.5 times higher than that of the standard MIC, which is supposed affordable in a typical modern mobile terminal.

## V. Simulation Results

For all numerical results, a typical urban channel profile is considered, and the terminal speed is 3 km/h (TU3). Ideal frequency hopping is used in the 900 MHz band. To users 1 (desired signal) and 2, legacy training sequence TSC0 [16] and new MUROS TSC0 from [11] are assigned, respectively. Speech transmission with adaptive multirate (AMR) speech coding with half rate (TCH/AHS 5.9 codec) is investigated.

For the interference from other cells, the MTS-1 and MTS-2 models from [17] have been used. In MTS-1, only a single cochannel interferer is present, while an interference mixture has been defined for MTS-2, with $I_{\mathrm{tot}}/I_1 = 0.6$ dB ($I_{\mathrm{tot}}$: average receive power of total external interference, $I_1$: average receive power of dominant external interferer).

A fixed–point receiver implementation with channel estimation and filter adaptation has been used in each case, and a time slot based frequency offset compensation has been active [18]. Receiver impairments such as phase noise and I/Q imbalance have been also taken into account, and typical values for an implementation have been selected, cf. [19].

In Fig. 1, the frame error rate (FER) of user 1 after channel decoding versus $C_1/I_1$ ($C_1$: average receive power of user 1) is shown for joint MLSE, MIC, and SIC. Results are given for different ratios of $C_1/C_2$ ($C_2$: average receive power of user 2)[2]. Also shown is the performance of a conventional GSM equalizer (CEQ) for a transmission without second MUROS user and the reference performance requirement according to [16] for non–DARP capable mobile terminals ($\diamond$). For the external interference from other cells, the MTS-1 scenario has been assumed. MIC corresponding to the first level of MUROS introduction [20] (only standard SAIC mobile terminals used) exhibits a performance degradation for increasing $C_2$, but still meets the reference performance requirement. Due to the requirement of suppressing a strong second MUROS user, MIC cannot take into account the external interference in an optimum way. In contrast, SIC corresponding to the second level of MUROS mobile terminals [20] degrades only slightly for increasing $C_2$ because interference of the second user is removed by successive interference cancellation before a second MIC is applied for suppression of external interference. The performance of joint MLSE is also only slightly dependent on $C_2$, similar to SIC. However, there is a loss of about 2 dB compared to SIC because joint MLSE does not have an appropriate remedy against external interference. Therefore, SIC should be preferred in interference limited scenarios. MIC can be also recommended if $C_1 \gtrapprox C_2$ is fulfilled but exhibits a noticeable performance degradation if the second MUROS user is significantly stronger than the considered user, e.g. $C_1/C_2 = -8$ dB, cf. Fig. 1. Nevertheless, all considered receivers fulfill the performance required in [16], and in many cases an even better performance results as for the conventional equalizer without a second MUROS user (reference interference performance of GSM before introduction of SAIC).

In Fig. 2, FER of user 1 versus $C_1/I_1$ is shown for the MTS-2 interference scenario and different receiver algorithms. Except for the external interference scenario, the same assump-

[1] Iterative approaches might be used in this case which, however, are beyond of the scope of this paper.

[2] In contributions to standardization, MUROS performance is often also shown versus $C/I_1$ with $C = C_1 + C_2$, which includes also the loss in transmit power due to the second MUROS user, cf. e.g. [19]. In our representation, a direct comparison of receiver performance is possible for varying $C_2$.

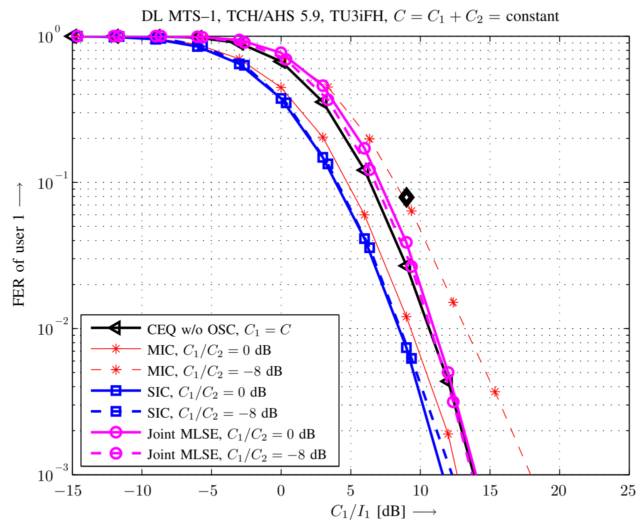


Fig. 1. FER of user 1 versus $C_1/I_1$ for different $C_1/C_2$ for joint MLSE, MIC, and SIC. Also shown is the performance of a conventional GSM equalizer (CEQ) without second MUROS user and the reference performance requirement ($\diamond$). MTS-1 scenario for external interference.

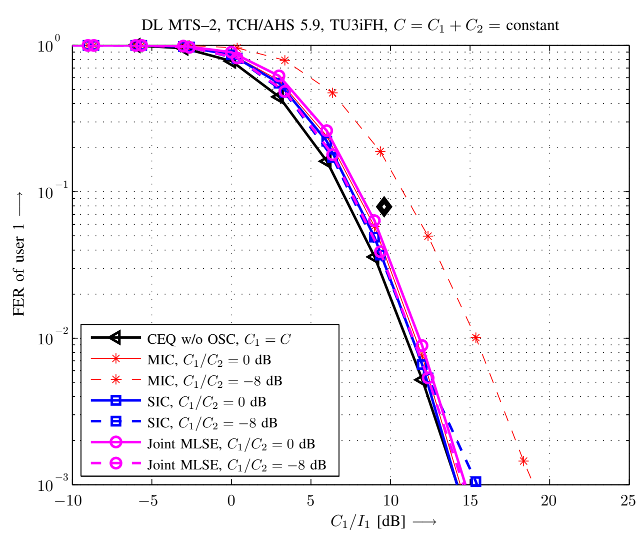


Fig. 2. FER of user 1 versus $C_1/I_1$ for different $C_1/C_2$ for joint MLSE, MIC, and SIC. Also shown is the performance of a conventional GSM equalizer (CEQ) without second MUROS user and the reference performance requirement ($\diamond$). MTS-2 scenario for external interference.

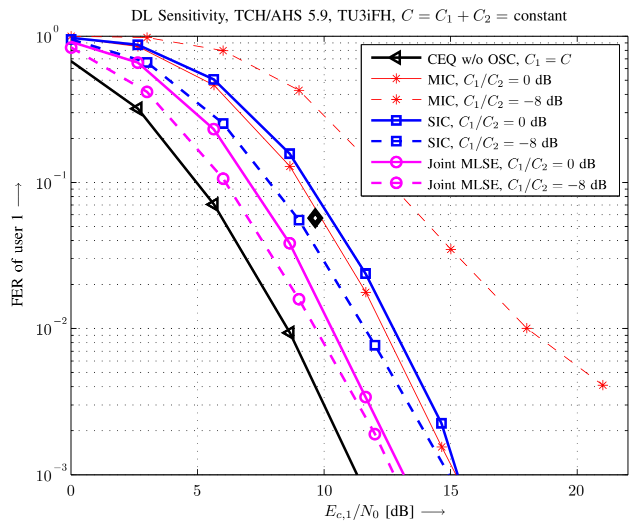


Fig. 3. FER of user 1 versus $E_{c,1}/N_0$ for different $C_1/C_2$ for joint MLSE, MIC, and SIC. Also shown is the performance of a conventional GSM equalizer (CEQ) without second MUROS user and the reference performance requirement (◇). Noise limited scenario.

tions have been made as for Fig. 1. The main difference to the results in Fig. 1 is that the gap between SIC and joint MLSE is reduced now. This is because MIC can combat a single external interferer as in MTS-1 more efficiently than an external interference mixture. Due to this reason, also the conventional equalizer without second MUROS user is slightly better than SIC.

Fig. 3 shows the FER of user 1 versus $E_{c,1}/N_0$ ($E_{c,1}$: average receive energy per encoded bit of user 1, $N_0$: power spectral density of underlying passband noise process) for a noise limited scenario. As expected, joint MLSE which is the optimum receiver for this scenario exhibits the best performance, followed by SIC and MIC. SIC is able to improve the performance of a pure MIC noticeably for $C_1/C_2 \ll 1$. In contrast to MIC, joint MLSE and SIC perform even better for increasing $C_2$ because then, the second user can be better detected which also aids in detecting the user of interest. The gap between the proposed SIC receiver and joint MLSE can be almost closed by using a SIC receiver with iterative processing (results not shown here). This, however, increases complexity significantly. Therefore, joint MLSE may be viewed as most favorable receiver algorithm for noise limited environments.

## VI. Conclusions

Algorithms for channel estimation and detection have been presented for the GSM MUROS downlink. In channel estimation, the single channel impulse response and power imbalance factor of both users have to be estimated. A joint maximum–likelihood approach and an approach for consecutive estimation of both quantities have been introduced. The calculated estimates can be used for a joint MLSE for equalization or detection approaches based on MIC. As an alternative to the standard MIC algorithm, an enhanced MIC with additional successive interference cancellation (SIC) can be employed for MUROS, which has the best performance of all considered schemes in interference limited scenarios. In the absence of external interference, joint MLSE is the most favorable detection scheme. Therefore, the detection algorithms presented in this paper have to be combined with a proper switching scheme which selects the most favorable algorithm according to the given environment. The results of this paper confirm that MUROS is highly beneficial for upgrading existing GSM networks in a straightforward way without introducing a loss compared to the reference performance.

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