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GSM/EDGE: A mobile communications system determined to stay

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This paper is dedicated to Prof. Johannes Huber on the occasion of his 60th birthday.

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ABSTRACT

The Global System for Mobile Communications (GSM) is still the most successful mobile communications system worldwide. During the last decade, major steps have been taken towards an improvement of GSM with respect to performance and spectral efficiency. To serve the increasing demands of more and more users for higher data rates, significant efforts are made towards a further evolution of GSM.

The standard extension Voice Services over Adaptive Multi-User Channels (VAMOS) has been specified in 3GPP Release 9. In VAMOS mode, two overlaid Gaussian minimum-shift keying (GMSK) signals are transmitted in the same time slot and frequency channel, potentially doubling the capacity of existing GSM networks in many cases. In this paper, we address channel estimation and detection techniques for the VAMOS downlink.

In a new 3GPP work item called TIGHTER, it is also foreseen to tighten all downlink receiver performance requirements for all data and voice services by at least 2 dB through the application of more sophisticated processing techniques at the receiver. Furthermore, a 3GPP study on Signal Precoding Enhancements for EGPRS2 (SPEED) to further modify the modulation and coding for data services by introducing OFDM has been started recently. We present ideas for improved receivers to enable tighter performance requirements and for SPEED transmission concepts, which might merit further investigation.

1. 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. For example, more than 500,000,000 additional GSM users have been reported in the year 2007. According to operators, the growth is considered especially dramatic in emerging markets in Asian and African countries. Because in particular voice services are requested, there is a need for a major voice capacity enhancement within the GSM system in order to meet the demands of the customers.

One approach for improving the spectral efficiency of GSM is to employ a tighter frequency reuse. This, however, increases adjacent channel and cochannel interference from other users in the system. As a consequence, the performance of a conventional GSM equalizer based on maximum-likelihood sequence estimation (MLSE) or the BCJR algorithm [1] 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. [2–8], exploiting the special properties of the Gaussian minimum-shift keying (GMSK) modulation of GSM, 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. It should be noted that in most cases antenna diversity is not available at the mobile station. Capacity improvements due to SAIC of more than 50% have been reported in field trials [9]. In 3GPP Release 6, corresponding Downlink Advanced Receiver Performance (DARP Phase I) requirements have been specified [10], and SAIC capable phones are widely used now even in the low-cost market segment.

Mobile terminals of low cost are required for emerging markets. Therefore, GMSK combined with SAIC seems to be a more promising avenue in this case than transmission with higher-order modulation such as 8-ary phase-shift keying (8PSK) or quadrature amplitude modulation (QAM). The latter is a possible alternative for improving the spectral efficiency but requires highly linear ampli-

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fiers and more complex equalizers. As a consequence, Multiple Users Reusing One Slot (MUROS) has been intensively studied in 3GPP [11] with focus on simpler modulations, cf. [12,13], for the specification of Voice Services over Adaptive Multiuser Orthogonal Subchannels (VAMOS) in 3GPP Release 9. In VAMOS, the capacity is increased by deliberately overlaying two users in the same time slot and at the same frequency resource within a cell. By this, the capacity can be doubled if the users can be separated with an only small loss in performance compared to the single-user case. VAMOS can be applied for full rate as well as half rate voice transmission. While in the uplink, cf. [14], a 2×2 virtual multiple-input multiple-output (MIMO) system with different subchannels results if the base station is equipped with two sufficiently separated receive antennas, the situation is more delicate in the downlink. Here, only a single receive antenna can be assumed for each one of the two considered mobile terminals, and for each of the mobile terminals the two transmit signals of the base station travel through the same propagation channel. In order to enable a sufficiently good user separation, the Orthogonal Subchannels (OSC) concept has been proposed for the downlink of VAMOS as a signaling scheme among other approaches [12]. It should be noted that for VAMOS, there is no need for a modification of the cellular architecture of the system, in contrast to the option of realizing capacity gains by decreasing the reuse factor.

For well-established markets with less stringent cost requirements, also transmission with higher-order modulation is an option for a capacity improvement of GSM. Making use of this, the capacity and transmission quality of GSM has been already improved significantly during the last decade. In particular, Enhanced Data Rates for GSM Evolution (EDGE) has been introduced which employs 8PSK modulation in addition to the GMSK modulation of the original GSM system. Furthermore, Enhanced GPRS Phase 2 (EGPRS2) has been standardized where 16-ary and 32-ary QAM (16QAM and 32QAM), respectively, are used along with turbo coding in order to further increase the data rates. For any modulation format introduced so far, the receiver performance is continuously improving because of increasingly more sophisticated signal processing techniques realized in hardware and software. The continuing growth of GSM/EDGE networks as the most ubiquitous mobile communications system and the short renewal cycles of mobile phones render the deployment of enhanced receivers a realistic and beneficial option. The specification of different performance levels is relevant to leverage the penetration of networks with high performance mobile terminals, as has been first done for SAIC and is now also considered in 3GPP TSG GERAN for higher-order modulation for data services [15,16]. To this end, the TIGHTER work item [17] has been introduced in 3GPP TSG GERAN with the objective to define a new performance level for the mobile terminal which improves the single antenna receiver performance significantly for all relevant voice and data services with various modulation types and for various test cases (sensitivity, adjacent channel and cochannel interference) without any changes to the air interface, i.e., only by employing improved signal processing algorithms.

Furthermore, also an alternative option for improving EGPRS2 performance is currently investigated in 3GPP TSG GERAN. Here, the downlink transmission of EGPRS2 is modified by employing a precoder in the transmitter. It is expected that via precoding, significant performance gains can be achieved both in interference and sensitivity limited scenarios, allowing for an increase in data rates and spectral efficiency of EGPRS2. The corresponding item in standardization is referred to as SPEED. Because key parameters of GSM/EDGE such as symbol interval, burst length, carrier frequency etc. will remain unchanged for SPEED, it is anticipated that SPEED can be smoothly integrated in the GSM/EDGE system. In particular, the spectral masks of GSM/EDGE will be still fulfilled by SPEED. Currently, the discrete Fourier transform (DFT) is considered as a precoder module in SPEED [18]. Thus, an orthogonal frequency division multiplexing (OFDM) transmission would result for SPEED, such as in the downlink of the Long Term Evolution (LTE) system, which simplifies equalization for higher-order modulation signals significantly.

In this paper, we shed some light on digital signal processing techniques for VAMOS, TIGHTER, and SPEED. In Section 2, efficient receivers for a VAMOS downlink OSC transmission are introduced, cf. also [19]. In particular, techniques for channel estimation are discussed. Here, it should be noted that the channel estimation problem for VAMOS differs significantly from that for a conventional GSM transmission, requiring novel algorithms. Furthermore, different approaches for equalization and interference cancellation are presented and performance results are given. In Section 3, a technique for cochannel interference reduction for arbitrary linear modulation formats is discussed, which may be viewed as a candidate for TIGHTER receivers. Also, it is shown to what extent the GSM reference performance can be improved, if sophisticated receiver concepts are employed. In Section 4, some transmission concepts for SPEED are proposed, and conclusions are drawn in Section 5.

2. Receiver algorithms for VAMOS downlink transmission

2.1. System model

In the considered scenario of a VAMOS 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. While in principle this strategy is equivalent to a transmission with filtered quaternary phase-shift keying (QPSK) modulation, it has the advantage that legacy GMSK transmitters can be directly used and that different transmit powers can be assigned to both signals. After GMSK derotation at the receiver, in equivalent complex baseband representation, the discrete-time received signal of one of the two mobile terminals can be written as

$$r[k] = \sum_{\kappa=0}^{q_h} h[\kappa] a_1[k-\kappa] + jb \sum_{\kappa=0}^{q_h} h[\kappa] a_2[k-\kappa] + n[k] + q[k]. \quad (1)$$

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. $h[k]$ is assumed to be 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 user 1 and user 2 of the VAMOS user pair with variance σ_a^2, respectively. In the following, the signal detection in the mobile terminal of user 1 is considered. $n[k]$ and $q[k]$ refer to discrete-time white Gaussian noise with variance σ_n^2 and adjacent channel 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.

2.2. 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 user 2 is that of user 1 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 (N: length of training sequences), 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 with variance σ_w^2. The joint maximum-likelihood (ML) estimates $\hat{\boldsymbol{h}}$ and $\hat{b}$ for $\boldsymbol{h}$ and b, respectively, result from minimizing the L_2-norm of the error vector $\boldsymbol{e} = \boldsymbol{r} - \boldsymbol{A}_1\hat{\boldsymbol{h}} - \hat{b}\boldsymbol{A}_2\hat{\boldsymbol{h}}$. Differentiating $\boldsymbol{e}^H\boldsymbol{e}$ $((\cdot)^H$: Hermitian transposition) with respect to $\hat{\boldsymbol{h}}^*$ $((\cdot)^*$: complex conjugation) and $\hat{b}$ and setting the derivatives to zero results in the following two conditions for the ML estimates of $\boldsymbol{h}$ and b:

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

$$\hat{b} = (\hat{\boldsymbol{h}}^H\boldsymbol{A}_2^H\boldsymbol{A}_2\hat{\boldsymbol{h}})^{-1}\,\mathrm{Re}\{(\hat{\boldsymbol{h}}^H\boldsymbol{A}_2^H)(\boldsymbol{r} - \boldsymbol{A}_1\hat{\boldsymbol{h}})\}. \tag{4}$$

Eqs. (3) and (4) may be also viewed as ML channel estimate for given b, cf. e.g. [20], and ML estimate of b for given channel vector, respectively. However, it does not seem to be possible to obtain a closed-form solution for $\hat{\boldsymbol{h}}$ and $\hat{b}$ from both coupled equations. Thus, a solution might be calculated iteratively by inserting an initial choice for $\hat{b}$ in (3), using the resulting channel vector for refining $\hat{b}$ via (4), etc., until convergence is reached.

2.3. Approaches for equalization and interference cancellation

2.3.1. Joint MLSE

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

$$\tilde{\boldsymbol{S}}[k] = [\tilde{a}_1[k-1]\tilde{a}_2[k-1]\ldots\tilde{a}_1[k-q_h]\tilde{a}_2[k-q_h]] \tag{5}$$

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

$$\lambda[k] = \left| r[k] - \sum_{\kappa=0}^{q_h}\hat{h}[\kappa]\tilde{a}_1[k-\kappa] - j\hat{b}\sum_{\kappa=0}^{q_h}\hat{h}[\kappa]\tilde{a}_2[k-\kappa]\right|^2. \tag{6}$$

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.

2.3.2. 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 VAMOS without any change if legacy training sequences are transmitted. In the following, the mono interference cancellation (MIC) algorithm from [5–7] 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. 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

$$y[k] = \mathcal{P}_c\left\{\sum_{\kappa=0}^{q_p}p[\kappa]r[k-\kappa]\right\} \tag{7}$$

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

$$\mathcal{P}_c\{x\} = \frac{<x, c>}{|c|^2} = \frac{\mathrm{Re}\{xc^*\}}{|c|^2} \tag{8}$$

($<\cdot, \cdot>$: inner product of two vectors). It is shown in [7], 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 [7] 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 $jbh[0]$ 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 [7], suppression of the second user is possible without any loss in signal-to-noise ratio (SNR), and $\mathrm{SNR} = 2|h[0]|^2(\sigma_a^2/\sigma_n^2)$ 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 [7]. In this case, the minimum mean-squared error (MMSE) filter obtained adaptively is a solution which is optimally adjusted to the interference mixture. Given this and the fact that for TU channels the interference created by the other user of the same base station is close to orthogonal to the desired user in many cases, it can be expected that MIC performs better than joint MLSE in scenarios with additional interference from other cells.

2.3.3. MIC receiver with successive interference cancellation (S-MIC)

In interference limited scenarios, more sophisticated schemes than DARP Phase I receivers can be employed. In particular, one can make use of 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 via two separate MIC algorithms.

In a MIC receiver with successive interference cancellation (S-MIC), channel estimation according to Section 2.2 is performed first. If $\hat{b} \geq b_0$ (e.g. $b_0 = 0.8$), $a_2[\cdot]$ 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

$$r_c[k] = r[k] - j\hat{b}\sum_{\kappa=0}^{q_h}\hat{h}[\kappa]\hat{a}_2[k-\kappa], \tag{9}$$

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

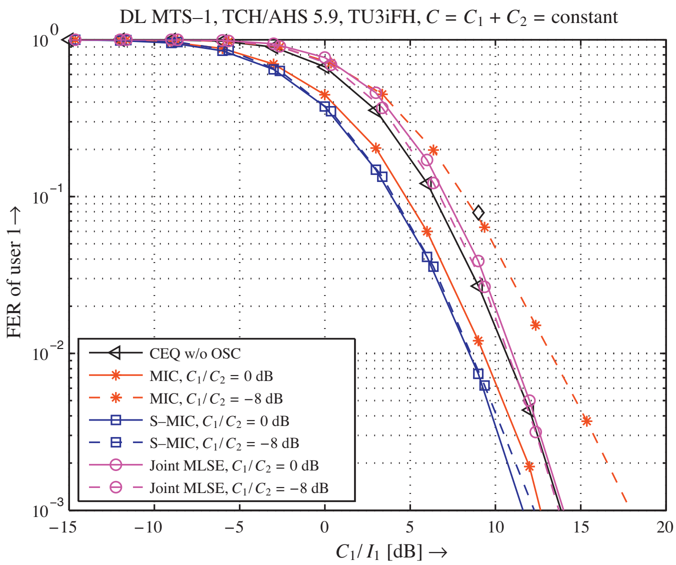

Fig. 1. FER of user 1 versus C_1/I_1 for different C_1/C_2 values for joint MLSE, MIC, and S-MIC. Also shown is the performance of a conventional GSM equalizer (CEQ) without second VAMOS user and the reference performance requirement (label '$\diamond$' at C_1/I_1 = 9 dB). MTS-1 scenario for external interference.

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]

It should be noted that MIC corresponds to the first level of VAMOS introduction [21] which stands for the phase where VAMOS signals are already transmitted by the base stations but still received by legacy SAIC enabled mobile terminals. In a later phase, referred to as the second level of VAMOS introduction, mobile terminals with receivers specifically tailored for VAMOS such as S-MIC will be available, and an even better network performance than for the first level can be expected.

2.4. Simulation results

For all numerical results presented in this section, 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 [10] and new VAMOS training sequence TSC0 from [12] 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 model from [22] has been used, where only a single cochannel interferer is present. A fixed-point receiver implementation with channel estimation and filter adaptation has been used in each case, and a timeslot based frequency offset compensation has been applied [23]. Receiver impairments such as phase noise and I/Q imbalance have been also taken into account, and typical values for a practical implementation have been selected, cf. [24].

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, I_1: average power of interferer) is shown for joint MLSE, MIC, and S-MIC. Results are given for different ratios of C_1/C_2 (C_2: average receive power of user 2). Also shown is the performance of a conventional GSM equalizer for a transmission without second VAMOS user and the reference performance requirement according to [10] for non-DARP capable mobile terminals (label '$\diamond$'). MIC exhibits a performance degradation for increasing C_2, but still meets the reference performance requirement. Due to the requirement of suppressing a strong second VAMOS user, MIC cannot mitigate the external interference in an optimum way. In contrast, S-MIC 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 almost independent of C_2, similar to S-MIC. However, there is a loss of about 2 dB compared to S-MIC because joint MLSE has no remedy against external interference. Therefore, S-MIC should be preferred in interference limited scenarios. Nevertheless, all considered receivers fulfill the performance requirement specified in [10], and in most cases an even better performance than for the conventional equalizer without a second VAMOS user (reference interference performance of GSM before introduction of SAIC) results.

3. Improved robustness to interference for tighter requirements

3.1. Modified decoding with L_p-norm metric

In order to enable tighter performance requirements for the case of higher-order modulation signals impaired by cochannel interference, various enhancements of receiver signal processing are conceivable. In the following, we consider exemplarily a modification of the channel decoder in order to adjust its branch metrics to a non-Gaussian total noise which typically occurs in interference-limited scenarios.

We first revisit a convolutionally encoded BPSK transmission over an additive white Gaussian noise (AWGN) channel with received signal

$$r[k] = Ac[k] + n[k]. \tag{10}$$

Here, $r[k]$ represents the real part of the complex-valued received signal,[2] A is a real-valued channel gain factor, $c[k]$ denotes the convolutionally encoded BPSK transmit sequence, $c[k] \in \{\pm 1\}$, and $n[k]$ stands for real-valued AWGN. Then, the metric for ML decoding of the convolutionally encoded transmit sequence can be written as

$$\Lambda(\tilde{c}[\cdot]) = \sum_{k=-\infty}^{+\infty} \left(\frac{\mathrm{LLR}[k]}{\mathrm{SNR}} - \tilde{c}[k] \right)^2, \tag{11}$$

where $\tilde{c}[k]$ denotes a hypothetical encoded symbol considered by the decoder, and

$$\mathrm{LLR}[k] = \ln \left(\frac{\Pr(c[k] = +1 | r[k])}{\Pr(c[k] = -1 | r[k])} \right) \tag{12}$$

is the log-likelihood ratio for symbol $c[k]$; $\mathrm{SNR} = A^2/\sigma_n^2$.

In channel decoding for GSM/EDGE with either GMSK or higher-order modulation, the metric in (11) is also applied, but LLR[k] is delivered by a soft-output equalizer, e.g. a BCJR algorithm, and there is a demapper and a deinterleaver between the equalizer and the channel decoder. Also, the factor of 1/SNR for scaling of the LLR's in (11) is often ignored because it does not affect the performance if the channel conditions are constant over the codeword.

The considered decoding problem is closely related to the decoding of bit-interleaved coded modulation (BICM) signals, because for a GSM/EDGE transmission with higher-order modulation, coding and modulation may be viewed as a kind of BICM. The decoding of BICM signals has been addressed e.g. in [25,26]. Here, a

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

[2] The imaginary part is not relevant for further processing.

Table 1
Gains in dB with respect to GSM specifications [10], obtained by employing sophisticated receiver algorithms [16] (Sens.: sensitivity; CCI: cochannel interference; ACI: adjacent channel interference).

Transmission scheme	Sens.	CCI	ACI	DTS1	DTS2
GMSK (AMR-NB)	3.0	9.0	13.0	7.5	2.5
8PSK (MCS 5–9)	4.0	3.5	9.0	–	–

similar metric as that in (11) results. However, for the case of a non-Gaussian total noise, the metric (11) is no longer optimum. Usually, the optimum metric cannot be determined anymore because the probability density function (pdf) of the total noise is unknown at the receiver. Nevertheless, it has been shown in [25] that significant performance gains can be obtained in the case of narrowband interference or impulsive noise, if an L_p-norm metric with proper choice of the positive parameter p ($p \neq 2$ in general) is adopted,

$$\Lambda_p(\tilde{c}[\cdot]) = \sum_{k=-\infty}^{+\infty} (\mathrm{LLR}[k] - \tilde{c}[k])^p, \tag{13}$$

where the optimum p can be determined by adaptive algorithms [26].

In GSM/EDGE, an impairment of the received signal by cochannel interference is typically present only within a part of a received burst for asynchronous networks. In this case, the effective *binary* channel observed by the decoder including soft-output equalization, demapping, and deinterleaving can be also viewed as a channel impaired by statistically independent impulsive noise, i.e., only some random bit positions are impaired by high levels of noise plus interference. It should be noted that this holds for arbitrary modulation formats. Typically, the impulsive noise can be characterized well by a Gaussian mixture [25]. Thus, it can be expected that the adoption of the L_p-norm metric with an adaptively optimized p yields significant performance gains.

This concept can be also applied to a transmission impaired by adjacent channel interference and to channel decoding in a VAMOS transmission.

3.2. Numerical results

In Table 1, it is shown which performance improvements can be achieved compared to the performance requirements of the current GSM standard if currently known sophisticated receiver algorithms with more precise calculation of soft output are used, cf. also [16]. The gains are valid with respect to the E_c/N_0 (sensitivity; E_c: average receive energy per encoded bit, N_0: power spectral density of underlying passband noise process) and C/I (interference limited scenarios; C: average receive power, I: average power of the total interference), respectively, specified for a certain FER in [10]. As interference limited scenarios, cochannel interference, adjacent channel interference, and the DARP test scenarios DTS1 and DTS2 [10] have been chosen. For GMSK modulation, AMR Narrowband (AMR-NB) speech coding has been considered, whereas for 8PSK an EGPRS transmission with modulation and coding schemes MCS5–MCS9 has been studied. Furthermore, all power delay profiles which are relevant for GSM/EDGE have been considered. The presented gains are worst-case gains with respect to all scenarios which have been investigated for each case. Also, the sensitivity gain is fully based on digital signal processing, consistently assuming a conservative noise figure of 8 dB of the analog receiver. Additional gains are deemed feasible with the concepts presented in Section 3.1. This is currently under investigation. Hence, there is much room for a tightening of the reference performance requirements of GSM/EDGE, enabled by the adoption of sophisticated signal processing techniques.

4. Transmission concepts for SPEED

4.1. OFDM with real-valued signal constellations

For SPEED, modulation and coding formats to be used in conjunction with OFDM transmission are to be specified. For scenarios with severe cochannel interference, a novel strategy for downlink OFDM transmission was presented in [27], which combines coded real-valued amplitude-shift keying (ASK) modulation and SAIC. In principle, real-valued ASK modulation suffers from a reduced power efficiency compared to QAM modulation because the constellation points are packed less densely in the complex plane. However, since only one real dimension is occupied per use of the complex channel, degrees of freedom which are unused for transmission are available for low-complexity interference suppression at the receiver using e.g. the MIC algorithm. The proposed scheme enables high downlink data rates already at low C/I values and is capable of exploiting large dominant-to-residual interference ratios (DIRs) contrary to conventional OFDM transmission with QAM modulation. A comparison to QAM transmission has shown that the novel scheme is superior with respect to bit error rate (BER) or FER after channel decoding for DIRs of at least 5 dB for the same spectral efficiency. For higher values of DIR, gains of up to 14 dB have been observed. The superior performance of the proposed scheme has been confirmed by a closed-form analysis of the raw BER in [27]. Therefore, by exploiting the additional degrees of freedom gained by using real-valued modulation one can more than compensate for the loss in power efficiency of ASK and enable high user data rates via a blind interference suppression scheme which does not require any explicit knowledge about the interferers and is moderate in terms of computational complexity.

4.1.1. System model

After an inverse DFT (IDFT) at the receiver, the discrete-time received signal for OFDM subcarrier μ of the ith receive signal block is given by

$$R_i[\mu] = H_i[\mu]A_i[\mu] + \sum_{\nu=1}^{J} G_{\nu,i}[\mu]B_{\nu,i}[\mu] + N_i[\mu], \tag{14}$$

where ν is the interferer index, $1 \leq \nu \leq J$. The discrete-time channel impulse responses for the desired signal and the interferer signals are assumed to be constant during the transmission of a data burst but may change randomly from burst to burst (block fading). The corresponding discrete frequency responses are $H_i[\mu]$ and $G_{\nu,i}[\mu]$ for the desired signal and the νth interferer, respectively. $A_i[\mu]$ and $B_{\nu,i}[\mu]$ denote the real-valued data symbols of the desired user and the νth interferer, respectively, at symbol time i and subcarrier frequency μ. The receiver noise is represented in the frequency domain by $N_i[\mu]$. For (14) it has been assumed that the OFDM symbols of the desired signal and the interferers are time-aligned, i.e., that the network is synchronous.

4.1.2. SAIC algorithm for OFDM

In the following, the SAIC approach of [5–7] based on complex filtering and subsequent projection of the filtered signal onto an arbitrary non-zero complex number c, cf. also Section 2.3.2, is adapted to the problem at hand. Because usually multiple interferers are present, the filter coefficients are optimized such that the variance of the difference between the signal after projection and the desired signal is minimized, i.e., the MMSE criterion is applied, guaranteeing interference suppression at minimum noise enhancement. As in an OFDM system the channel can be considered as flat for each subcarrier, the MIC algorithm is directly applicable to each subcarrier, and the required order of the complex filter for subcarrier μ is zero, i.e., it reduces to a scalar $P_i[\mu]$. We denote the real-valued output signal of the projection by $Y_i[\mu]$. The error signal, consisting of noise and residual interference, is given by

$$E_i[\mu] = Y_i[\mu] - A_i[\mu] = \mathcal{P}_c\{P_i[\mu]R_i[\mu]\} - A_i[\mu], \quad (15)$$

where $\mathcal{P}_c\{x\}$ denotes projection of x onto an arbitrary non-zero complex number c, cf. (8).

Adopting an MMSE approach, the associated cost function is defined as

$$J(P_i[\mu]) \triangleq \mathcal{E}\left\{(\mathcal{P}_c\{P_i[\mu]\cdot R_i[\mu]\} - A_i[\mu])^2\right\} \quad (16)$$

($\mathcal{E}\{\cdot\}$: expectation). Exploiting the fact that the cost function is convex we determine its minimum via the zeros of its derivative, resulting in

$$\frac{\partial J(P_i[\mu])}{\partial P_i^*[\mu]} = \Phi_{RR}[\mu]P_i[\mu] + \Phi_{R^*R}[\mu]P_i^*[\mu] - 2\,\varphi_{AR}[\mu] \stackrel{!}{=} 0. \quad (17)$$

The expressions in (17) are defined as

$$\Phi_{RR}[\mu] = \mathcal{E}\left\{R_i[\mu]R_i^*[\mu]\right\} = \sigma_a^2\cdot|H_i[\mu]|^2 + \sum_{\nu=1}^{J}\sigma_\nu^2\cdot|G_{\nu,i}[\mu]|^2 + 2\,\sigma_n^2, \quad (18)$$

$$\varphi_{AR}[\mu] = \mathcal{E}\left\{A_i[\mu]R_i^*[\mu]\right\} = \sigma_a^2\cdot H_i^*[\mu], \quad (19)$$

$$\Phi_{R^*R}[\mu] = \mathcal{E}\left\{(R_i^*[\mu])^2\right\} = \sigma_a^2\cdot(H_i^*[\mu])^2 + \sum_{\nu=1}^{J}\sigma_\nu^2\cdot(G_{\nu,i}^*[\mu])^2, \quad (20)$$

where σ_a^2 and σ_ν^2 $(1 \le \nu \le J)$ denote the variances of the desired signal and the νth interferer signal, respectively, and σ_n^2 is the variance of the inphase component of the rotationally symmetric noise $N_i[\mu]$. After straightforward calculations, we obtain the solution for the MMSE filter $P_i[\mu]$:

$$P_i[\mu] = 2\,\frac{\varphi_{AR}[\mu]\Phi_{RR}[\mu] - \varphi_{AR}^*[\mu]\Phi_{R^*R}[\mu]}{\Phi_{RR}^2[\mu] + |\Phi_{R^*R}[\mu]|^2}. \quad (21)$$

In a practical system, the MMSE solution can be approached by an adaptive algorithm.

In summary, we consider inclusion of modulation and coding schemes (MCSs) in SPEED which are based on ASK transmission and enable the application of a low-complexity SAIC algorithm. Such MCSs might be especially beneficial for scenarios with strong cochannel interference and are able to outperform OFDM transmission with PSK or QAM, provided that the dominant interferer uses also ASK modulation.

4.2. OFDM with unique words

Because there is still flexibility in the specification of SPEED, it is not mandatory to employ a conventional OFDM transmission, but also more recently developed concepts for OFDM can be considered. In [28], the potential of unique words used in OFDM symbols has been investigated. Here, instead of the conventional cyclic prefix (CP) of an OFDM symbol, a deterministic sequence, referred to as unique word, is used for the decoupling of adjacent OFDM symbols. The unique word is inserted at the beginning and at the end of each symbol and guarantees that an equivalent channel with cyclic convolution arises, similar to the case of a CP. Because the unique word is known in advance also at the receiver, it can be exploited for synchronization and channel estimation purposes.[3]

Furthermore, the approach in [28] is based on the introduction of redundant subcarriers, which is necessary in order to guarantee that at the end of each OFDM transmit symbol, the unique word appears without any interference from data-dependent parts. This introduction of redundancy can be also viewed as a kind of block coding using a complex field Reed-Solomon code [28]. It turns out that the transmit energy strongly depends on the positions of the redundant subcarriers. Therefore, a permutation matrix is employed which permutes the information carrying and redundant subcarriers in an optimum way. Finally, the generated word in the frequency domain may be viewed as a precoded version of the original information word, using an optimized precoding matrix $\boldsymbol{G}$. At the receiver, this precoding can be exploited by a linear MMSE receiver whose design is based on matrix $\boldsymbol{G}$ and the channel matrix. The results in [28] indicate that such a receiver is able to achieve a significant noise reduction by exploiting the correlation introduced by precoding the transmit signal with $\boldsymbol{G}$. Especially for weak channel coding, gains of up to 1 dB are possible compared to a conventional OFDM transmission [28].

In conclusion, we recommend the consideration of OFDM with unique words as a strongly competitive alternative for the standardization of SPEED. It should be noted that OFDM with unique words has no drawbacks in terms of spectral compactness compared to conventional OFDM.

5. Conclusions

GSM is still by far the most widely used mobile communications standard worldwide, and there are significant efforts underway to further enhance its performance and capacity. We have shed some light on the corresponding 3GPP work on VAMOS, tighter requirements, and SPEED and have shown that sophisticated digital signal processing techniques play a major role for the success of each of them. Hence, it can be expected that with the introduction of these concepts, again major progress can be made in the further development of GSM/EDGE in the near future, resulting in another increase in the spectral efficiency of the system. Therefore, it is expected that GSM/EDGE will continue to play a prominent role in mobile communications.

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