A Single Antenna Interference Cancellation Algorithm for GSM

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***Abstract*— In mobile communication networks, system capacity is often limited by cochannel interference. Therefore, receiver algorithms for cancellation of cochannel interference have recently attracted much interest. At the mobile terminal, algorithms can usually rely only on one received signal delivered by a single receive antenna. In this paper, a novel low–complexity single antenna interference cancellation (SAIC) algorithm for real–valued modulation formats referred to as mono interference cancellation (MIC) is introduced which is well suited for practical applications. By using this algorithm in the mobile terminals, capacity of GSM networks can be improved by up to 40-60 %.**

***Keywords*—Mobile communications, GSM, cochannel interference, equalization**

I. Introduction

In cellular mobile communication systems like the Global System for Mobile Communications (GSM), cochannel interference from cells using the same frequencies as the considered cell (frequency reuse) is an important capacity limiting factor. For systems with a low frequency reuse factor corresponding to high system capacity, cancellation of cochannel interference at the receiver is necessary in order to obtain a good performance. One class of interference cancellation algorithms is based on receive antenna diversity. However, in most cases, antenna diversity is only available at the base station but not at the mobile terminal due to cost and size limitations. Hence, for downlink transmission, single antenna interference cancellation (SAIC) algorithms are required. In this paper, a novel SAIC algorithm referred to as mono interference cancellation (MIC) is introduced which is applicable to general real–valued modulation formats and in particular to the Gaussian minimum–shift keying (GMSK) modulation used in GSM, which can be well approximated by filtered binary phase–shift keying (BPSK). The proposed SAIC algorithm, cf. also [1], employs complex–valued filtering and projection and is followed by trellis–based equalization. The required operations are well suited for a low–complexity software implementation. Furthermore, it is demonstrated that MIC has advantages compared to other SAIC algorithms proposed in the literature, e.g. [2], [3], [4].

II. MIC Algorithm

A. Basic Strategy

For derivation of the algorithm, we consider the case of a mobile terminal with a single antenna whose received signal is impaired by intersymbol interference (ISI), cochannel interference resulting from a single interfering base station, and additive white Gaussian noise. A transmission with linear modulation using real–valued coefficients is assumed for the desired signal as well as for the interferer. Thus, the discrete–time complex baseband received signal is given by

$$r[k] = \sum_{\kappa} h[\kappa]\, a[k-\kappa] + \sum_{\kappa} g[\kappa]\, b[k-\kappa] + n[k] \qquad (1)$$

($a[k]$, $b[k]$: real–valued data symbols of desired signal and interferer, respectively; $n[k]$: white Gaussian noise). $h[\kappa]$ and $g[\kappa]$ are the overall impulse responses corresponding to the desired and the interfering signals, respectively. The mobile channel is assumed to be (approximately) constant during each data burst.

For the algorithm, an arbitrary non–zero complex number c is selected and a corresponding number

$$c^{\perp} = \mathrm{Im}\{c\} - j\,\mathrm{Re}\{c\} \qquad (2)$$

($\mathrm{Re}\{\cdot\}$, $\mathrm{Im}\{\cdot\}$: real and imaginary part of a complex number, respectively) 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

$$y[k] = \mathcal{P}_c\Big\{ \sum_{\kappa} p[\kappa]\, r[k-\kappa] \Big\} \qquad (3)$$

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

$$\mathcal{P}_c\{x\} = \frac{< x,\, c >}{|c|^2} = \frac{\mathrm{Re}\{x\, c^*\}}{|c|^2} \qquad (4)$$

($< \cdot, \cdot >$: inner product of two vectors).

In the following, the choice of the filter transfer function $P(z) = \mathcal{Z}\{p[k]\}$ ($\mathcal{Z}\{\cdot\}$: z–transform) is addressed. Using the definitions

$$G_R(z) = \mathcal{Z}\{\mathrm{Re}\{g[k]\}\}, \quad (5)$$
$$G_I(z) = \mathcal{Z}\{\mathrm{Im}\{g[k]\}\} \quad (6)$$

and assuming knowledge of $G_R(z)$, $G_I(z)$ at the receiver, we propose to select $P(z)$ according to

$$P(z) = c^{\perp}\, Q(z)\, (G_R(z) - j\, G_I(z)), \quad (7)$$

where $Q(z) = \mathcal{Z}\{q[k]\}$ with arbitrary real–valued coefficients $q[k]$. Hence,

$$P(z)\, G(z) = c^{\perp} Q(z)\, (G_R^2(z) + G_I^2(z)). \quad (8)$$

All coefficients of the combined impulse response corresponding to $P(z)\, G(z)$ have the same phase as $c^{\perp}$. Therefore, also the filtered interference has the same phase as $c^{\perp}$ and vanishes after projection onto c, i.e., zero forcing with respect to the interferer is performed. This is possible because the interferer occupies only a subspace of the entire signal space. Furthermore, it can be shown that the useful signal after projection does not vanish if

$$\frac{H_R(z)}{H_I(z)} \neq \frac{G_R(z)}{G_I(z)} \quad (9)$$

($H_R(z) = \mathcal{Z}\{\mathrm{Re}\{h[k]\}\}$, $H_I(z) = \mathcal{Z}\{\mathrm{Im}\{h[k]\}\}$), and a high signal–to–noise ratio results if the ratios on both sides of (9) are clearly different. Thus, the feasibility of interference suppression by the proposed approach has been demonstrated. For practical purposes, we propose a modification of the given strategy which is introduced in the following and is particularly suited for an adaptive implementation which does not require explicit knowledge of the channel impulse responses $h[\kappa]$, $g[\kappa]$.

B. *Adaptive Implementation*

In an adaptive implementation of the MIC strategy, the FIR filter $P(z) = \sum_{k=0}^{q_p} p[k]\, z^{-k}$ should be chosen in such a way, that the signal after projection approximates a desired signal

$$w[k] = \sum_{\kappa=0}^{q_d} d[\kappa]\, a[k - k_0 - \kappa] \quad (10)$$

with real–valued coefficients $d[k]$ and a delay k_0 which are both free parameters for optimization. The filter coefficients are optimized for minimization of the error consisting of cochannel interference and noise, i.e., a minimum mean–squared error (MMSE) criterion is now applied instead of zero forcing (ZF) with respect to the interferer. This is because MMSE filters can be calculated more efficiently. In the MMSE optimum solution, the output signal of projection $y[k]$ is expected to be approximately free of cochannel interference, which is possible in principle according to the previous considerations, and has minimum noise enhancement. Intersymbol interference is not removed since it can be better taken into account by a subsequent trellis–based equalizer. In order to avoid the trivial solution in filter optimization, the additional constraint $d[0] = 1$ is adopted. It should be noted that this constraint is similar to that in optimization of the filters of decision–feedback equalization (DFE). Because the feedforward filter of a ZF–DFE produces white output noise, the noise component in $y[k]$ is approximately white for the optimum filters. The filter order q_d can be chosen for a tradeoff between performance and complexity of trellis–based equalization. Adaptive adjustment of the filter coefficients $p[\cdot]$ and $d[\cdot]$ may be performed according to the least–mean–square (LMS) algorithm. In order to calculate the desired signal of adaptation $w[k]$ (10), a training sequence which is known at the receiver has to be transmitted during a certain time interval. For example, in the GSM system each burst contains a training sequence of length 26 which may be used for filter adaptation. Adaptation may be continued thereafter switching to the decision–directed mode in which tentative decisions of the trellis–based equalization are employed. It is important to note that the algorithm performs blind adaptation with respect to the interferer because only the training sequence of the desired signal has to be known, contrary to joint detection approaches for interference cancellation [5].

For description of the LMS adaptation, the time–varying filter coefficient vectors

$$\boldsymbol{p}[k] = [\, p[0,k] \quad p[1,k] \quad \ldots \quad p[q_p,k]\,]^H \quad (11)$$

$$\boldsymbol{d}[k] = [\, d[1,k] \quad d[2,k] \quad \ldots \quad d[q_d,k]\,]^T \quad (12)$$

are defined ($(\cdot)^H$: Hermitian transposition, $(\cdot)^T$: transposition). The output signal of projection (3) can be represented as

$$y[k] = \mathcal{P}_c\{\boldsymbol{p}^H[k]\, \boldsymbol{r}[k]\} \quad (13)$$

with

$$\boldsymbol{r}[k] = [\, r[k] \quad r[k-1] \quad \ldots \quad r[k-q_p]\,]^T, \quad (14)$$

and the desired signal of adaptation (10) may be written as

$$w[k] = a[k-k_0] + \boldsymbol{d}^T[k]\, \boldsymbol{a}[k] \quad (15)$$

with

$$\boldsymbol{a}[k] = [\, a[k-k_0-1] \quad a[k-k_0-2] \quad \ldots \quad a[k-k_0-q_d]\,]^T \quad (16)$$

Using [6] for derivation of the LMS algorithm for minimization of the variance of the adaptation error $e[k] = \mathcal{P}_c\{\boldsymbol{p}^H[k]\, \boldsymbol{r}[k]\} - a[k-k_0] - \boldsymbol{d}^T[k]\, \boldsymbol{a}[k]$, we obtain the recursive update equations

$$\boldsymbol{p}[k+1] = \boldsymbol{p}[k] - \mu\, \frac{1}{2}\, c^*\, e[k]\, \boldsymbol{r}[k], \quad (17)$$
$$\boldsymbol{d}[k+1] = \boldsymbol{d}[k] - \mu\, e[k]\, \boldsymbol{a}[k], \quad (18)$$

where μ denotes the step size of adaptation [6]. For initialization, all–zero vectors may be chosen for $\boldsymbol{p}[0]$ and $\boldsymbol{d}[0]$. It should be noted that coefficient vector $\boldsymbol{d}[k]$ is real–valued because $e[k]$ and $\boldsymbol{a}[k]$ are real–valued. Although all choices for c result in the same performance, the choices $c = 1$ or $c = j$ are preferable regarding a simple implementation.

Alternatively, a recursive least–squares (RLS) algorithm may be selected for filter adaptation, or the optimum filters may be determined via direct minimization of the time–averaged squared error within a window of size K, $1/K\, \sum_{k=0}^{K-1} (y[k] - w[k])^2$.

III. Numerical Results

For performance evaluation, a GSM Adaptive Multi Rate (AMR) speech transmission with 12.2 kbit/s over a typical urban (TU) 50 channel (speed of mobile terminal: 50 km/h) without frequency hopping is considered first. For all simulations in this section, the GSM 1800/1900 MHz frequency band has been used, and a mobile station with typical hardware impairments being modeled in the simulations is assumed. The inphase/quadrature (I/Q) phase and gain imbalance are $3°$ and -0.5 dB, respectively. Furthermore, a DC offset of 8 dBc was added to the received signal, and Gaussian phase noise with $2.5°$ rms was present. Also, unless otherwise stated, a fixed–point DSP code implementation of MIC has been used in the simulations which can be directly applied in practice. Fig. 1 shows the frame error rate (FER) after decoding versus the carrier–to–interference ratio C/I_t. Here, C and I_t are the average received power of the desired signal and the total interference, respectively. In order to model practical situations sufficiently accurate, a scenario with 4 synchronous [5] interferers has been selected, where one of them dominates and has average power $I_1 = I_d$ while the remaining three interferers have equal average powers $I_2 = I_3 = I_4 = I_r/3$ (I_r: power of residual interference, $I_t = I_d + I_r$). MIC yields a gain of 12 dB compared to a standard receiver without interference cancellation [7] at FER $\approx 10^{-2}$ in case of a single interferer ($I_d/I_r \to \infty$). In comparison, the algorithm of [2] gains about 6 dB for a related scenario, cf. [2]. For $I_d/I_r = 8$ dB, MIC improves performance still by more than 5 dB whereas an improvement of 3 dB results for the quite pessimistic assumption $I_d/I_r = 0$ dB.

It should be noted that for all interference–limited scenarios of this section, also additive white Gaussian noise with $E_c/N_0 = 30$ dB was present (E_c: average received energy per coded bit, N_0: one–sided power spectral density of the underlying passband noise process).

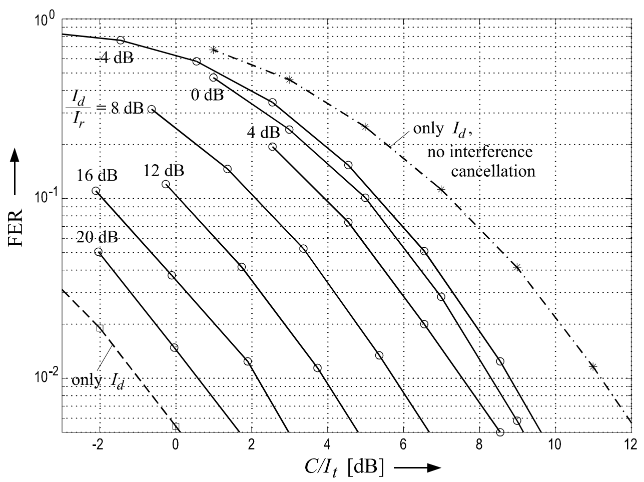

Fig. 1. FER of MIC versus C/I_t for various I_d/I_r (4 interferers, $I_t = I_d + I_r$). Dash–dotted line: standard receiver, only a single interferer. Dashed line: MIC, only a single interferer. TU 50 channel without frequency hopping, AMR 12.2 kbit/s.

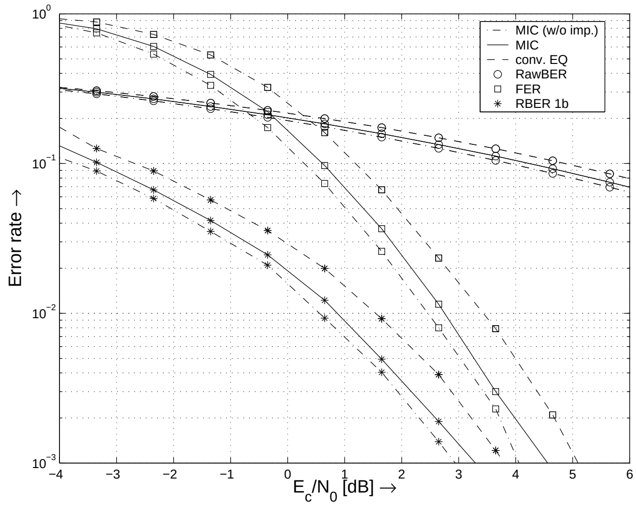

Fig. 2. RawBER ('o'), FER ('□'), and RBER ('$*$') versus E_c/N_0 for reference sensitivity test case. Solid lines: MIC receiver, with impairments; dash–dotted lines: MIC receiver, without impairments; dashed lines: standard receiver, with impairments. TU 3 channel with frequency hopping, AMR 5.9 kbit/s.

In practical applications, it is important that a receiver performs well not only in interference–limited but also in coverage–limited scenarios. Therefore, the reference sensitivity test case, characterized by a pure white Gaussian noise disturbance, is also investigated. Fig. 2 shows FER, the raw bit error rate (RawBER) before decoding, and the residual BER (RBER) for the Class 1b coded bits [7] versus E_c/N_0 for a GSM AMR speech transmission with 5.9 kbit/s over a TU 3 channel with ideal frequency hopping. The MIC receiver performs even slightly better than the standard receiver, which is due to optimized channel estimation and tracking used in the case of MIC and the capability of MIC to suppress DC offset as a special case of interference. Thus, the proposed MIC receiver entails no penalty for coverage–limited scenarios. In contrast, the SAIC algorithm of [3], [4] performs worse than the standard receiver in this case, cf. Fig. 6 of [3]. Also shown in Fig. 2 are results for MIC simulated without impairments which demonstrate that the loss due to impairments is quite small.

Furthermore, with respect to a practical implementation, a robust behavior to frequency offset is desirable. Due to Doppler shift and imprecision of local oscillators, temporary frequency offsets of 500 Hz or even more may be observed. In addition, because the desired signal and the interfering signals are sent by different base stations, different offsets for both signals are likely to occur which makes receiver design even more challenging. Fig. 3 shows RawBER, FER, and RBER for the Class 1b coded bits versus C/I for an AMR transmission with 5.9 kbit/s over a TU 3 channel with frequency hopping. A single interferer has been present. For the frequency offsets of both signals, three different scenarios are considered: 1) zero offset for desired signal (D) and interferer (I), 2) 200 Hz and 300 Hz offset for D and I, respectively, and 3) 400 Hz and 500

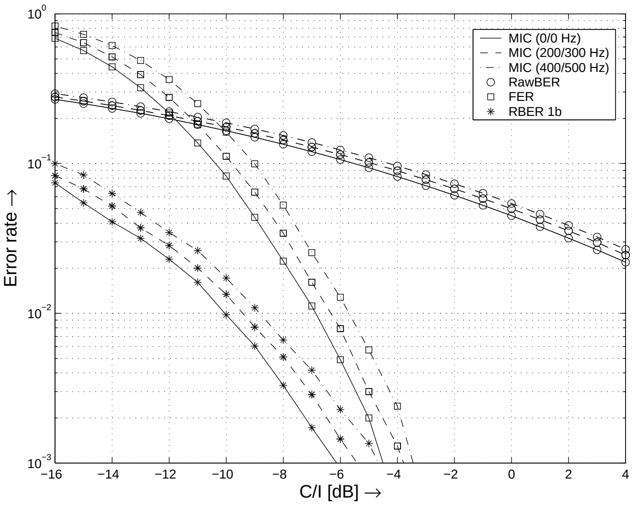

Fig. 3. RawBER ('o'), FER ('□'), and RBER ('∗') of MIC versus C/I for additional frequency offset. Solid lines: case 1), dashed lines: case 2), dash–dotted lines: case 3). TU 3 channel with frequency hopping, AMR 5.9 kbit/s.

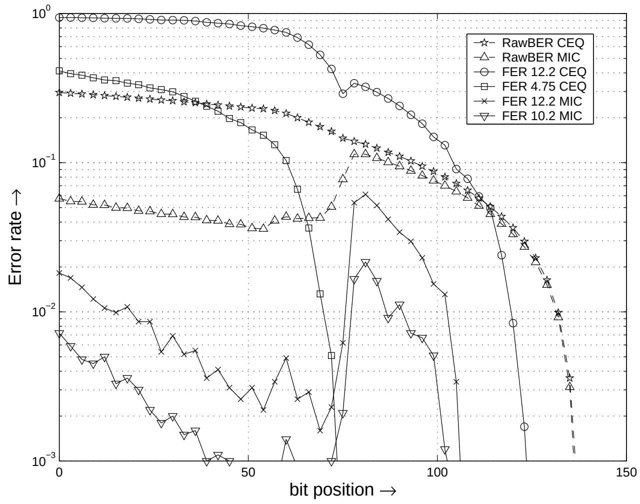

Fig. 4. RawBER (dashed lines) and FER (solid lines) versus offset in bit positions for asynchronous interference. AMR modes for MIC: 12.2 kbit/s and 10.2 kbit/s; AMR modes for conventional equalizer (CEQ): 12.2 kbit/s and 4.75 kbit/s. TU 50 channel without frequency hopping.

Hz for D and I, respectively. Outer loop offset control has been disabled, and only a frequency offset estimation/compensation algorithm embedded in MIC [8] has been active as a countermeasure against the offset. Fig. 3 shows that even for case 3), degradation due to frequency offset is only about 1.5 dB. Also in case 1), frequency offset compensation has been active as in all simulations for MIC. Hence, the proposed MIC strategy can be made quite robust to frequency offset by incorporating suitable extensions [8]. In particular, it is remarkable that time–slot–based frequency offset compensation for MIC still works for a C/I below 0 dB. It should be also noted that the adopted frequency offset compensation which has to use complex–valued signals requires a complex–valued filter $P(z)$ separated from projection.

Using the results shown in Fig. 3 for zero frequency offset, a further comparison of MIC to the SAIC algorithm of [3], [4] is possible. According to Fig. 3 of [4], which in principle holds for the same assumptions as Fig. 3 except for $E_c/N_0 = 100\,\mathrm{dB}$ in [4], $C/I \approx 7.5\,\mathrm{dB}$ is required for $\mathrm{RawBER} = 2 \cdot 10^{-2}$ for the scheme of [3], [4], whereas MIC needs only $C/I \approx 4\,\mathrm{dB}$ for the same target RawBER.

It is worth to be mentioned that adjacent channel interference can be suppressed by MIC in the same manner as cochannel interference, in contrast to SAIC approaches based on joint detection. This property of MIC becomes even more important when frequency reuse is further tightened. In this case, significant adjacent channel interference will come from neighboring sectors of the serving cell.

So far, synchronous networks have been assumed, resulting in a full overlapping of the bursts of desired signal and interferer. However, also asynchronous networks are still in operation, in particular in Europe. Therefore, performance of MIC for the asynchronous case is relevant, too. Fig. 4 shows RawBER and FER of MIC and the conventional equalizer (CEQ), respectively, for different AMR modes versus the offset between the bursts of desired signal and interferer in bit positions. A single interferer is assumed, and $C/I = -3\,\mathrm{dB}$. During shifting of the interferer, its received power is kept fixed, i.e., the interfering energy per time slot of the desired signal decreases. A TU 50 channel without frequency hopping has been selected. Zero offset corresponds to synchronous transmission, whereas an offset of 156 bits means that both bursts have disjoint time intervals. For CEQ, performance is depicted for AMR transmission with 12.2 kbit/s and 4.75 kbit/s, respectively, whereas for MIC, the AMR modes with 12.2 kbit/s and 10.2 kbit/s have been chosen. Please note, that in case of MIC and transmission rates below 10.2 kbit/s, error rates are lower than the range adopted for Fig. 4 and therefore are not shown here. Also for asynchronous interference, MIC performs significantly better than a conventional receiver. Roughly speaking, performance of MIC is optimum when the interferer starts at the beginning of the training sequence of the desired signal, corresponding to an offset of 61 bit positions. Here, the left half of the burst remains essentially error–free. Additionally, MIC can be adjusted approximately optimum to the right half because the interferer is present during the entire training sequence. As the interferer shifts further to the right, the performance deteriorates because the sample size for adaptation is no longer sufficiently high. The worst performance results when the beginning of the interferer lies close to the end of the training sequence. Further shifting again improves performance because the total collected interferer energy decreases, and less bit positions are contaminated by interference.

Test mobiles based on MIC are already available and have been used for extensive field trials in networks of different operators [9] in order to determine the network capacity for the case of MIC mobiles. The results demonstrate that significant capacity gains can be achieved by MIC. In [10], network capacity is numerically evaluated by simulations assuming an homogeneous network. For this assumption, capacity gains of about 40 % are reported in [10]. Also, non–homogeneous system simulations based on real network data have been conducted in [11] which show that even higher gains of up to 60 % can be achieved in real networks.

IV. CONCLUSIONS

A low–complexity strategy for single antenna interference cancellation referred to as mono interference cancellation (MIC) has been proposed. This novel technique is based on complex filtering, projection, and subsequent trellis–based equalization. Numerical results have demonstrated that MIC behaves quite robust in practical environments characterized for example by several simultaneous interferers and frequency offset. Furthermore, MIC is suitable for synchronous as well as asynchronous networks. The MIC algorithm has been already successfully employed in commercial field tests where it has been shown that it may yield capacity improvements of up to 40-60 % [5], [9], [10], [11]. Finally, we note that MIC might be also employed in different applications with real–valued modulation such as orthogonal frequency–division multiplexing (OFDM) combined with BPSK.

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