

A Single Antenna Interference Cancellation Algorithm for Increased GSM Capacity*

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Abstract

In mobile communications 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 letter, a 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. Field trials in commercial GSM networks using prototype terminals with the proposed MIC algorithm have demonstrated that the novel concept may yield capacity improvements of up to 80 %. The underlying principle is also beneficial for adjacent channel interference and receivers with multiple antennas. Furthermore, in coverage-limited scenarios, there is no performance degradation compared with conventional receivers.

Keywords: Mobile communications, GSM, cochannel interference, equalization.

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1 Introduction

In cellular mobile communications 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. In case of multiple receive antennas, solutions for cochannel interference cancellation are well known [1]. 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 letter, 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 MIC algorithm, cf. also [2], 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. It is demonstrated that MIC has advantages compared to other SAIC algorithms proposed in literature, e.g. [3, 4].

This letter is organized as follows. In Section 2, the system model is introduced. In Section 3, the MIC strategy is presented for flat fading and frequency-selective fading channels. An adaptive implementation of MIC is discussed in Section 4. Some numerical results are given in Section 5.

2 System Model

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 symbol-spaced discrete-time received signal in equivalent complex baseband notation is given by

$$r[k] = \sum_{\kappa=0}^{q_h} h[\kappa] a[k - \kappa] + \sum_{\kappa=0}^{q_g} g[\kappa] b[k - \kappa] + n[k]. \quad (1)$$

Here, $a[k]$ and $b[k]$ denote the real-valued data symbols of the desired signal and the interferer, respectively (variance of both symbols: σ_x^2). $h[\kappa], \kappa \in \{0, 1, \dots, q_h\}$ and $g[\kappa], \kappa \in \{0, 1, \dots, q_g\}$ are the discrete-time causal overall impulse responses including the effects of transmit filtering, the mobile communications channel, and receiver input filtering corresponding to the desired and the interfering signals, respectively (q_h, q_g : channel orders). The mobile communications channel is assumed to be (approximately) constant during one transmitted data burst, which is justified e.g. for the GSM system, but may vary randomly from burst to burst due to frequency hopping. The Gaussian noise process $n[\cdot]$ with variance σ_n^2 is assumed to be rotationally symmetric and white. The latter holds for a square-root Nyquist continuous-time receiver input filter like a square-root cosine filter as it is employed in many GSM receivers [5].

3 MIC Strategy

3.1 Flat Fading Channel

As a special case of the given channel model, we consider first a discrete-time flat fading channel, $h[0] \neq 0, h[\kappa] = 0, \kappa \neq 0, g[0] \neq 0, g[\kappa] = 0, \kappa \neq 0$, resulting in $r[k] = h[0] a[k] + g[0] b[k] + n[k]$.

For the following, an arbitrary complex number c with $|c| = 1$ ($|\cdot|$: magnitude of a complex number) is selected and a corresponding number $c^\perp = \text{Im}\{c\} - j \text{Re}\{c\}$ ($\text{Re}\{\cdot\}, \text{Im}\{\cdot\}$: real and imaginary part of a complex number, respectively) is generated. c and c^\perp may be interpreted as mutually orthogonal vectors in two real-valued dimensions, $\langle c, c^\perp \rangle = 0$, where the inner product $\langle x, y \rangle$ between two complex numbers x and y is defined as $\langle x, y \rangle = \text{Re}\{x y^*\}$.

Let us multiply the received signal with a factor $p = c^\perp g^*[0]/|g[0]|$ ($(\cdot)^*$: complex conjugation), yielding

$$\begin{aligned} \tilde{r}[k] &= p \cdot r[k] = c^\perp \frac{g^*[0]}{|g[0]|} (h[0] a[k] + g[0] b[k] + n[k]) \\ &= c^\perp \frac{g^*[0]}{|g[0]|} h[0] a[k] + c^\perp |g[0]| b[k] + c^\perp \frac{g^*[0]}{|g[0]|} n[k]. \end{aligned} \quad (2)$$

The signal component pertaining to cochannel interference may be now considered as a two-dimensional vector with a direction given by c^\perp . Hence, the interference can be eliminated by projecting $\tilde{r}[k]$ onto a two-dimensional vector corresponding to c . Denoting the coefficient of projection of a complex number x onto c by $\mathcal{P}_c\{x\}$,

$$\mathcal{P}_c\{x\} = \frac{\langle x, c \rangle}{|c|^2} = \text{Re}\{x c^*\}, \quad (3)$$

we obtain with Eqs. (1) and (3)

$$\mathcal{P}_c\{\tilde{r}[k]\} = \frac{\text{Re}\{g^*[0] h[0] c^\perp c^*\}}{|g[0]|} a[k] + \frac{\text{Re}\{g^*[0] c^\perp c^* n[k]\}}{|g[0]|}. \quad (4)$$

Here, $\text{Re}\{c^\perp c^*\} = 0$ has been used. Because of $c^\perp c^* = -j$, Eq. (4) can be further simplified to

$$\mathcal{P}_c\{\tilde{r}[k]\} = \frac{\text{Im}\{g^*[0] h[0]\}}{|g[0]|} a[k] + \frac{\text{Im}\{g^*[0] n[k]\}}{|g[0]|}. \quad (5)$$

In particular, the signal after projection does not depend on the direction given by c .

The signal-to-noise ratio (ratio of variance of useful signal and variance of noise) after interference cancellation is given by

$$\text{SNR} = 2 \frac{\text{Im}^2\{g^*[0] h[0]\}}{|g[0]|^2} \frac{\sigma_x^2}{\sigma_n^2}. \quad (6)$$

Here, it has been used that $n[k]$ is rotationally symmetric, i.e., its variance with respect to any direction in the complex plane is given by $\sigma_n^2/2$. It can be easily shown that

$$0 \leq \text{SNR} \leq 2 |h[0]|^2 \frac{\sigma_x^2}{\sigma_n^2}, \quad (7)$$

where $\text{SNR} = 0$ if $\arg\{g[0]\} = \arg\{h[0]\}$ or $\arg\{g[0]\} = \arg\{h[0]\} + \pi$ and $\text{SNR} = 2 |h[0]|^2 \sigma_x^2 / \sigma_n^2$ if $\arg\{g[0]\} = \arg\{h[0]\} \pm \pi/2$ ($\arg\{\cdot\}$: argument of a complex number). For a mobile channel with random coefficients both extreme cases occur with probability zero. Hence, the proposed algorithm eliminates the interferer perfectly (zero forcing with respect to interferer) but entails a certain loss in signal-to-noise ratio depending on the relation of the arguments of the channel coefficients of desired signal and interferer.

3.2 Fading ISI Channel

For fading ISI channels multiplication with a scalar factor p prior to projection onto c is replaced by filtering using a filter with transfer function $P(z) = \mathcal{Z}\{p[k]\}$ ($\mathcal{Z}\{\cdot\}$: z -transform, $p[k]$: complex-valued impulse response of filter) and a corresponding output signal $\tilde{r}[k] = p[k] \star r[k]$ (\star : convolution). Using the definitions

$$G_R(z) = \mathcal{Z}\{\text{Re}\{g[k]\}\}, \quad G_I(z) = \mathcal{Z}\{\text{Im}\{g[k]\}\} \quad (8)$$

and assuming knowledge of $G_R(z)$, $G_I(z)$ at the receiver, the transfer function $P(z)$ is chosen as

$$P(z) = c^\perp Q(z) (G_R(z) - j G_I(z)), \quad (9)$$

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

$$P(z)G(z) = c^\perp Q(z) (G_R^2(z) + G_I^2(z)). \quad (10)$$

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., again zero forcing with respect to the interferer is performed. This holds for any factor $Q(z)$ in Eq. (9) with real-valued coefficients $q[k]$. Hence, the $q[k]$ represent degrees of freedom in filter design which may be utilized for example for additional noise suppression. The filtered received signal after projection is given by

$$\mathcal{P}_c\{\tilde{r}[k]\} = \mathcal{Z}^{-1}\{Q(z)(H_R(z)G_I(z) - H_I(z)G_R(z))\} \star a[k] + n_p[k]. \quad (11)$$

Here, $\mathcal{Z}^{-1}\{\cdot\}$ denotes inverse z -transform and

$$H_R(z) = \mathcal{Z}\{\text{Re}\{h[k]\}\}, \quad H_I(z) = \mathcal{Z}\{\text{Im}\{h[k]\}\}. \quad (12)$$

Hence, the overall impulse response including projection is real-valued. $n_p[k]$ is the real-valued noise after filtering and projection, which in general is colored and has variance

$$\sigma_{n_p}^2 = \frac{\sigma_n^2}{2} \int_{-1/2}^{1/2} |Q(e^{j2\pi f})|^2 |G_R(e^{j2\pi f}) - jG_I(e^{j2\pi f})|^2 df. \quad (13)$$

According to Eq. (11), the useful signal after projection does not vanish if

$$\frac{H_R(z)}{H_I(z)} \neq \frac{G_R(z)}{G_I(z)}, \quad (14)$$

and a high signal-to-noise ratio results if the ratios on both sides of (14) are clearly different. In principle, it would be possible to optimize the coefficients of $Q(z)$ for maximum signal-to-noise ratio. However, with this approach, it is difficult to control the length of the overall impulse response including projection.

A general problem of the presented strategy is that the interferer impulse response $g[k]$ which has to be known is difficult to estimate with good accuracy at the mobile station because the training sequence of the interferer is not available. Only blind approaches are applicable which are usually complex and not robust. However, even if channel estimation was feasible, the solution for filter $P(z)$ (Eq. (9)) might be quite sensitive to channel estimation errors. Furthermore, a colored noise process may result after projection which is undesirable for subsequent trellis-based equalization. Therefore, for practical purposes, we prefer the modified approach

of the next section which, in contrast to the approach in this section, in particular lends itself to adaptive implementation without requiring explicit channel knowledge. Nevertheless, the previous considerations have proven that perfect interference cancellation via filtering and projection is in principle feasible, although a receiver based on Eq. (9) is not directly suited for implementation.

4 Adaptive Implementation of MIC

4.1 Cost Function

For an adaptive implementation of the MIC strategy of Section 3.2, a suitable cost function depending on the FIR filter $P(z) = \sum_{k=0}^{q_p} p[k] z^{-k}$ has to be considered. For this, it should be taken into account that it is desirable that the signal after projection approximates a signal

$$w[k] = \sum_{\kappa=0}^{q_d} d[\kappa] a[k - k_0 - \kappa] \quad (15)$$

with real-valued coefficients $d[\kappa]$ and a delay k_0 which are both free parameters for optimization. Therefore, the filter coefficients may be optimized for minimization of the variance of the difference between the output signal after projection and $w[k]$, i.e., the cost function is selected as

$$J(p[0], p[1], \dots, p[q_p]) \triangleq E \{ (\mathcal{P}_c\{\tilde{r}[k]\} - w[k])^2 \} = E \left\{ \left(\mathcal{P}_c \left\{ \sum_{\kappa=0}^{q_p} p[\kappa] r[k - \kappa] \right\} - w[k] \right)^2 \right\} \quad (16)$$

($E\{\cdot\}$: expectation). Hence, a minimum mean-squared error (MMSE) criterion is applied instead of the zero-forcing (ZF) criterion of Section 3, and the total error consisting of interference and noise is considered. Intersymbol interference is again not removed since it can easily be dealt with in 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 $\mathcal{P}_c\{\tilde{r}[k]\}$ is approximately white for the optimum MMSE filters. The filter order q_d can be chosen for a tradeoff between performance and complexity of trellis-based equalization.

4.2 Filter Adaptation

Adaptive adjustment of the filter coefficients $p[\cdot]$ and $d[\cdot]$ for minimization of the cost function (16) may be done according to the least-mean-square (LMS) algorithm. In order to calculate

the desired signal of adaptation $w[k]$ (Eq. (15)), 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. 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 [6].

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

$$\mathbf{p}[k] = [p[0, k] \quad p[1, k] \quad \dots \quad p[q_p, k]]^H \quad (17)$$

$$\mathbf{d}[k] = [d[1, k] \quad d[2, k] \quad \dots \quad d[q_d, k]]^T \quad (18)$$

are defined ($(\cdot)^H$: Hermitian transposition, $(\cdot)^T$: transposition). The output signal after projection is $\mathcal{P}_c\{\mathbf{p}^H[k] \mathbf{r}[k]\}$ with

$$\mathbf{r}[k] = [r[k] \quad r[k-1] \quad \dots \quad r[k-q_p]]^T, \quad (19)$$

and the desired signal of adaptation (Eq. (15)) may be written as $w[k] = a[k-k_0] + \mathbf{d}^T[k] \mathbf{a}[k]$ with

$$\mathbf{a}[k] = [a[k-k_0-1] \quad a[k-k_0-2] \quad \dots \quad a[k-k_0-q_d]]^T. \quad (20)$$

Using [7] for derivation of the LMS algorithm for minimization of the variance of the adaptation error $e[k] = \mathcal{P}_c\{\mathbf{p}^H[k] \mathbf{r}[k]\} - a[k-k_0] - \mathbf{d}^T[k] \mathbf{a}[k]$, we obtain the recursive update equations

$$\mathbf{p}[k+1] = \mathbf{p}[k] - \mu \frac{1}{2} c^* e[k] \mathbf{r}[k], \quad (21)$$

$$\mathbf{d}[k+1] = \mathbf{d}[k] - \mu e[k] \mathbf{a}[k], \quad (22)$$

where μ denotes the step size of adaptation [7]. For initialization, all-zero vectors may be chosen for $\mathbf{p}[0]$ and $\mathbf{d}[0]$. It should be noted that the coefficient vector $\mathbf{d}[k]$ is real-valued because $e[k]$ and $\mathbf{a}[k]$ 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 based on $J(\cdot)$ 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 , $1/K \sum_{k=0}^{K-1} (\mathcal{P}_c\{\tilde{r}[k]\} - w[k])^2$. Because these algorithms can be derived in a similar manner as their counterparts for conventional adaptive MMSE filtering [7], details are omitted here. For an implementation in a GSM receiver, they may be preferred to the LMS algorithm because the training sequence employed in GSM is

relatively short and the LMS algorithm is known to converge slowly for colored input processes (both desired signal and interference may be significantly colored). Nevertheless, also the LMS algorithm may be applied if adaptation is extended beyond the training sequence, switching to the decision-directed mode. If only training symbols are used in adaptation, convergence to spurious local minima does not occur, because the underlying cost function $J(\cdot)$ is convex.

Finally, we note that it is straightforward to generalize all considerations of this section to a fractionally-spaced filter $p[\cdot]$, for example with $T/2$ -spaced processing.

5 Numerical Results

For all numerical results, a fixed-point digital signal processor (DSP) code implementation of MIC has been used in the simulations which may be directly applied in a mobile phone. A $T/2$ -spaced filter $p[\cdot]$ has been employed, with two T -spaced polyphase component filters of order $q_p = 3$. Furthermore, $q_d = 2$ and $k_0 = 3$ is valid, and the filter coefficients have been determined with the algorithm which directly minimizes the time-averaged squared error of the training symbol positions in each burst, solving a linear system of equations, cf. Section 4. In general, for increasing filter orders q_p and q_d , the adaptive solution will diverge more and more from the optimum MMSE solution because the sample size for adaptation is fixed, whereas performance of the MMSE solution improves. It turns out that the selected filter orders guarantee a good compromise between good approximation of the MMSE solution with adaptation and sufficiently low minimum mean-squared error of the optimum solution for the relevant GSM channel profiles.

For performance evaluation, we consider a GSM Adaptive Multi Rate (AMR) speech transmission with 5.9 kbit/s over a typical urban (TU) 3 (speed of mobile terminal: 3 km/h) channel with ideal frequency hopping. The GSM 1900 MHz frequency band is used. We assume a mobile station with typical hardware impairments being modeled in the simulations. The error vector magnitude (EVM) [8] is 6.5 % root mean square (rms), caused by an inphase/quadrature (I/Q) phase and gain imbalance of 3° and -0.5 dB, respectively, and Gaussian phase noise with 2.5° rms. Furthermore, a DC offset [9] of 8 dBc has been added to the received signal. For the desired signal and for the interferer, there has been a frequency offset of 200 Hz and 300 Hz, respectively, between transmitter and receiver. The receiver has been enhanced by suitable countermeasures against the mentioned impairments, such as the frequency offset estimation/compensation algorithm in [10].

In the following, we compare performance of a mobile terminal with MIC and subsequent

equalization with that of a terminal with soft-output trellis-based equalization with 16 states without interference cancellation as it is used in standard GSM mobiles [11, 12]. Both receivers have been realized as fully quantized 16 bit fixed-point implementations on a DSP. In each of the following cases, results for the raw bit error rate (RawBER) before decoding, the frame error rate (FER) after decoding, and the residual BER (RBER) for the Class 1b coded bits [11] are given.

First, the cochannel interference test case specified in [13] is considered (a single random GMSK modulated cochannel interferer). Fig. 1 shows error rates versus the carrier-to-interference ratio C/I . Here, C and I are the average received power of the desired signal and the interferer, respectively. The results indicate that more than 11 dB can be gained by MIC compared to a standard receiver at $\text{FER} \approx 10^{-2}$ as well as $\text{RBER} \approx 10^{-2}$. In comparison, the gains of the SAIC algorithm proposed in [3] are limited to about 6 dB for a similar scenario, cf. Fig. 6 of [3].

According to Fig. 2, similar gains can be obtained for the adjacent channel interference test case specified in [13]. For all GSM receivers, an adjacent channel protection (ACP) of at least 18 dB is desirable [13]. Here, an ACP of 18 dB means that performance in case of cochannel interference and in case of adjacent channel interference with an 18 dB higher average interference power are equal. A high ACP is important especially for systems with low reuse factor, because adjacent channels then might be often used in neighboring cells or other sectors of the same base station. This potentially results in strong adjacent channel interference due to near-far effects, as has been shown in field trials [14]. According to Figs. 1 and 2, an ACP of 18 dB is approximately fulfilled for MIC.

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 [13], which is characterized by a white Gaussian noise disturbance, is also investigated. Fig. 3 shows the corresponding error rates versus E_c/N_0 (E_c : average received energy per coded bit, N_0 : one-sided power spectral density of the underlying passband noise process)¹. The MIC receiver performs even slightly better than the standard receiver, which has several reasons. Channel estimation and tracking is not directly comparable (DFE filter calculation vs. direct channel estimation), and a different frequency offset estimation algorithm has to be used in case of MIC, cf. [10]. It should be noted that the considered conventional receiver is a commercial solution, which has been also used e.g. for obtaining some of the results in [15]. It fulfills the

¹For all considered interference-limited scenarios, $E_c/N_0 = 30$ dB is valid.

GSM performance recommendations [13], although it may be possible to enhance its performance further. Also, our results for the conventional receiver agree with results presented in [4]. In summary, the proposed MIC receiver entails no penalty for coverage-limited scenarios. In contrast, the SAIC algorithm of [4] performs worse than the standard receiver in this case although performance is similar to that of MIC for interference-limited scenarios, cf. Fig. 6 of [4].

In order to model practical situations more accurately than in the GSM test cases, we consider now a scenario with 4 synchronous [6] interferers, where one of them dominates and has average power I_d while the remaining three interferers have equal powers $I_r/3$ (I_r : power of residual interference). An AMR speech transmission with 12.2 kbit/s over a TU50 (speed of mobile terminal: 50 km/h) channel without frequency hopping is assumed for the following. Fig. 4 shows FER after decoding versus the carrier-to-interference ratio C/I_t . Here, I_t is the average received power of the total interference, $I_t = I_d + I_r$. Please note that also for a single interferer, performance is worse than for the previously considered scenario because less error protection is present in AMR 12.2 kbit/s and no frequency hopping is applied. Again, MIC yields a gain of 12 dB compared to the standard receiver without interference cancellation at $\text{FER} \approx 10^{-2}$ in case of a single interferer ($I_d/I_r \rightarrow \infty$). 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. Also here, all results are valid for fixed-point arithmetic in the receiver and additional hardware impairments.

6 Conclusions

A single antenna interference cancellation algorithm for real-valued modulation formats has been presented which is able to cancel a single cochannel interferer perfectly. For the GSM cochannel and adjacent channel interference test cases, high gains compared to a standard receiver have been demonstrated, whereas no performance degradation can be observed for coverage-limited scenarios. In practice, several interferers are present, but usually one of them dominates. The algorithm has been already successfully employed in commercial field tests where it has been shown that it may yield capacity improvements of up to 80 % [6, 14, 16]. These capacity gains have been confirmed by simulation results presented in [17]. A generalization to $N_R > 1$ receive antennas is possible, cf. [2]. In this case, $2N_R - 1$ interferers can be cancelled perfectly.

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Figures

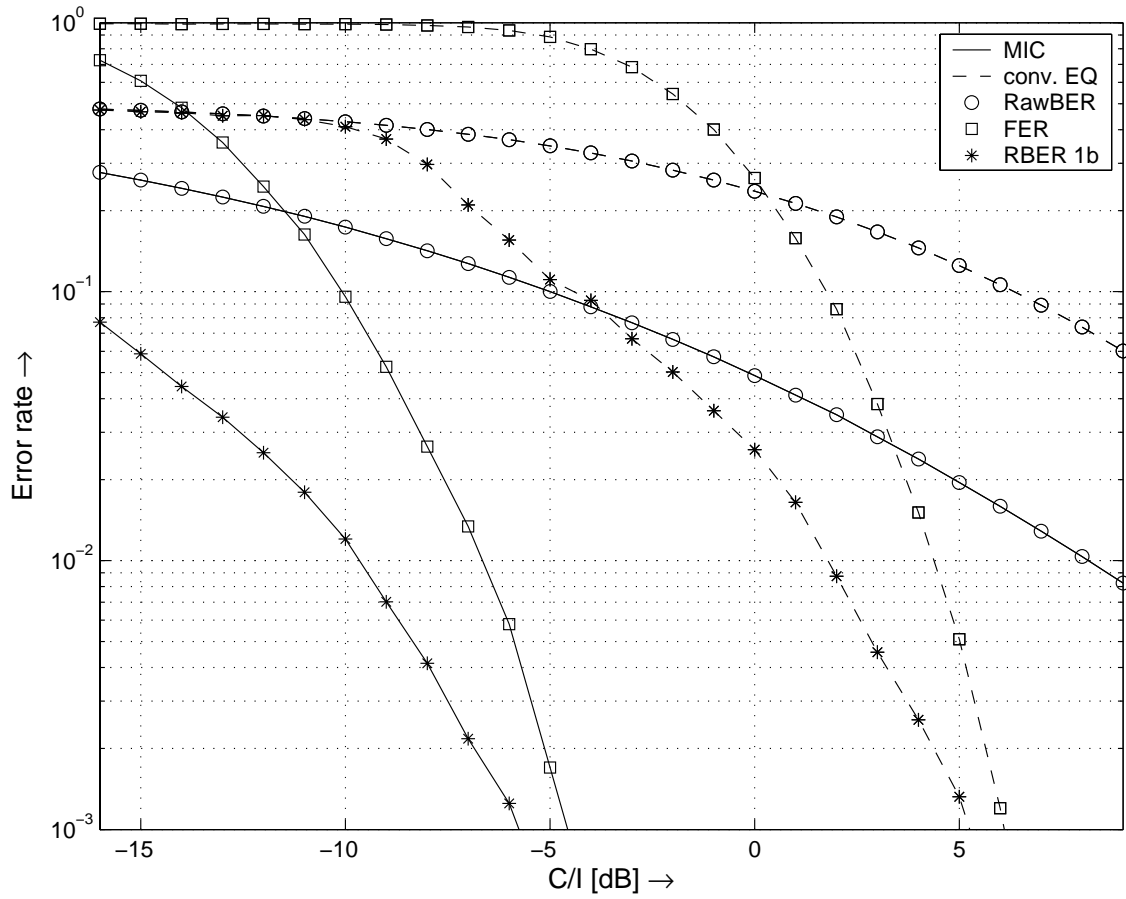


Figure 1: RawBER ('o'), FER ('□'), and RBER ('*') versus C/I for single cochannel interferer test case. Solid lines: MIC receiver; dashed lines: standard receiver. (AMR 5.9 kbit/s speech transmission, TU3 channel, frequency hopping.)

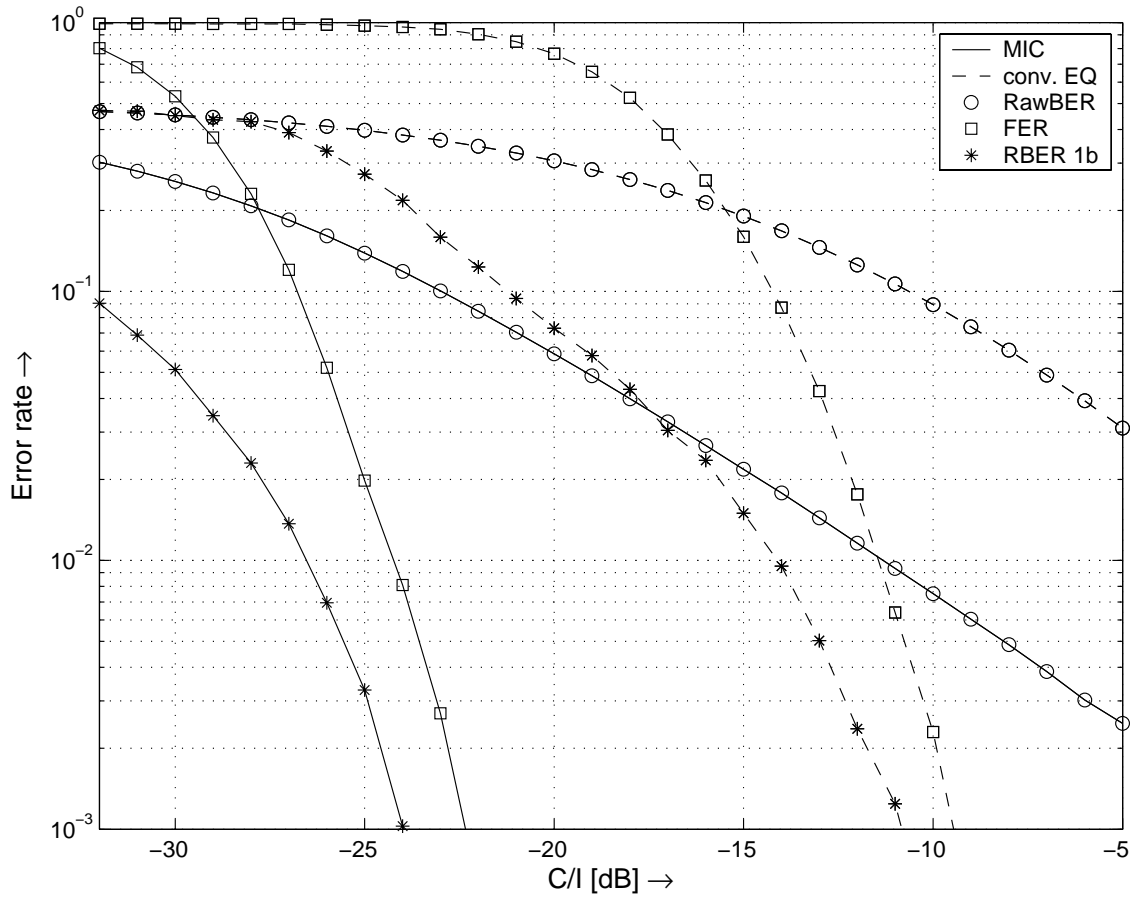


Figure 2: RawBER ('o'), FER ('□'), and RBER ('*') versus C/I for single adjacent channel interferer test case. Solid lines: MIC receiver; dashed lines: standard receiver. (AMR 5.9 kbit/s speech transmission, TU3 channel, frequency hopping.)

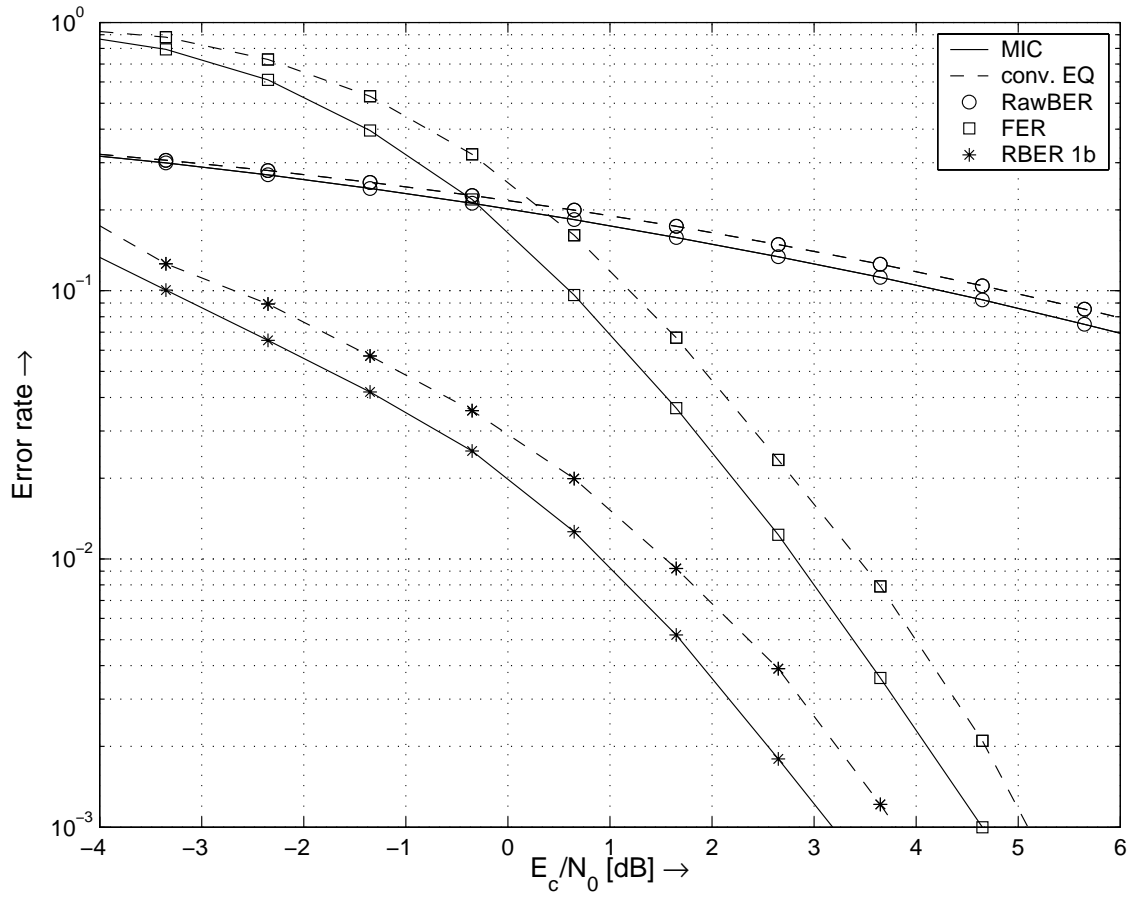


Figure 3: RawBER ('o'), FER ('□'), and RBER ('*') versus E_c/N_0 for reference sensitivity test case. Solid lines: MIC receiver; dashed lines: standard receiver. (AMR 5.9 kbit/s speech transmission, TU3 channel, frequency hopping.)

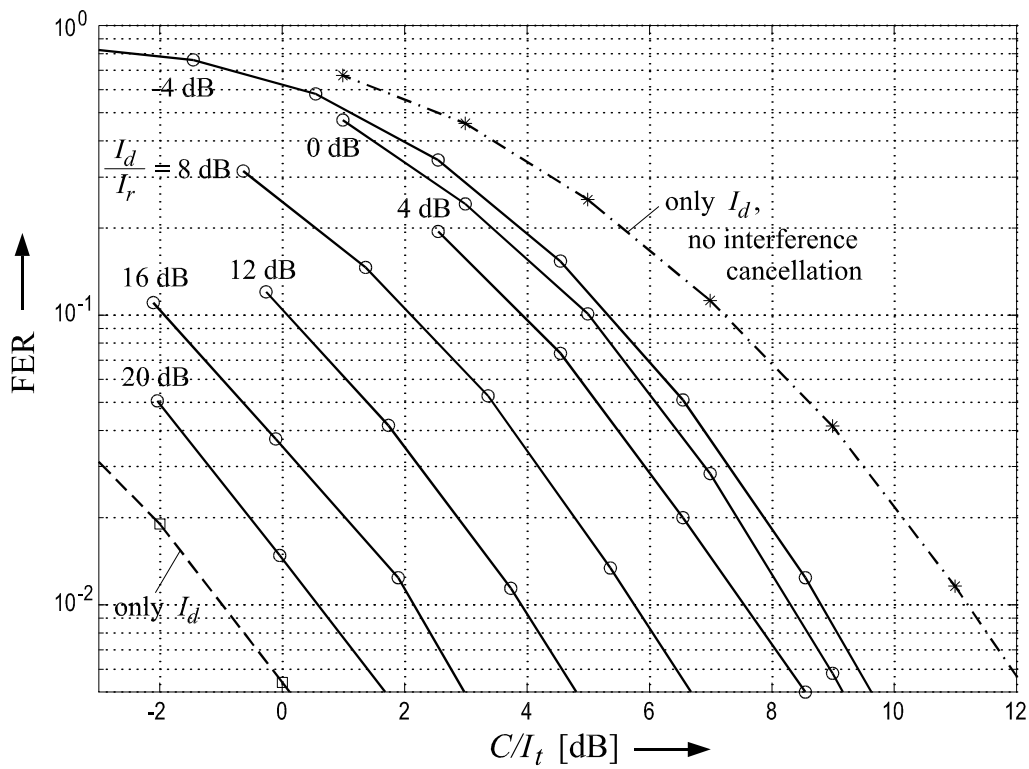


Figure 4: FER of MIC versus C/I_t for various I_d/I_r (4 interferers, $I_t = I_d + I_r$). Dash-dotted line: standard receiver, only a single interferer. Dashed line: MIC receiver, only a single interferer. (AMR 12.2 kbit/s speech transmission, TU50 channel, no frequency hopping.)