OFDM TRANSMISSION WITH SINGLE ANTENNA INTERFERENCE CANCELLATION

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ABSTRACT

Future mobile communications radio networks, e.g. 3GPP Long Term Evolution (LTE), will typically use an orthogonal frequency division multiplexing (OFDM) based air interface in the downlink. Furthermore, in order to avoid frequency planning, a frequency reuse factor of one is desirable. In this case, system capacity is limited by interference, which is particularly crucial for mobile terminals with a single receive antenna. For a high throughput, interference cancellation algorithms are required in the receiver. In this paper, a single antenna interference cancellation (SAIC) algorithm is introduced for amplitude–shift keying (ASK) modulation schemes used in coded OFDM transmission which achieves high gains in comparison to a conventional coded OFDM transmission employing quadrature amplitude modulation (QAM) in an interference limited scenario. Furthermore, an adaptive least–mean–square (LMS) and a recursive least–squares (RLS) SAIC receiver, respectively, are presented. We show that in particular the RLS solution enables a good tradeoff between performance and complexity and is robust even to multiple interferers and frequency synchronization errors.

1. INTRODUCTION

For next generation mobile communications air interfaces such as 3GPP Long Term Evolution (LTE) [1] orthogonal frequency division multiplexing (OFDM) will be adopted as basic transmission scheme in the downlink. In order to avoid frequency planning a frequency reuse factor of one is envisaged for 3GPP LTE. Hence, for receivers without interference suppression capabilities, transmission is possible only with relatively low data rates due to the resulting capacity limitation which contradicts the desire for high downlink data rates.

For this reason, interference cancellation and suppression has received high interest from both academia and industry and various contributions for OFDM transmission have been already made. For example, in [2], interference suppression for synchronous and asynchronous cochannel interferers is studied, where the proposed receiver employs an adaptive antenna array which performs minimum mean–squared error (MMSE) diversity combining. Related approaches based on receive diversity have been introduced e.g. in [3, 4].

Although multiple receive antennas are advantageous for cancellation of cochannel interference, due to cost and size limitations it is still a challenge to include more than one antenna into a mobile terminal. Therefore, single antenna interference cancellation (SAIC) algorithms have received significant attention in recent years, especially for transmission with single–carrier modulation, cf. [5, 6, 7, 8]. The advantages of SAIC for GSM radio networks were analyzed in [9], and in [6] it has been shown that GSM network capacity can be dramatically improved by SAIC. All mentioned approaches assume that the adopted modulation scheme can be interpreted as a scheme with real–valued data symbols, which holds e.g. for the Gaussian minimum–shift keying modulation of GSM.

OFDM transmission with real–valued data symbols has been studied in [10], and an equalizer for suppression of intercarrier interference resulting from the time variance of the mobile radio channel has been introduced which exploits the fact that the transmitted symbols are one–dimensional. However, cochannel interference and channel coding was not taken into account. In [11], a widely linear processing approach has been proposed for narrowband interference suppression and blind channel identification for OFDM transmission with real–valued symbols.

In this paper, we propose an SAIC algorithm for OFDM transmission, modifying the algorithm in [12, 6, 13] for single–carrier transmission suitably, referred to as mono interference cancellation (MIC). Real–valued amplitude–shift keying (ASK) modulation is assumed and additional channel coding is considered. Independent complex filtering with subsequent projection for interference removal is applied to each OFDM subcarrier. We present the analytical solution for the MMSE filter and also adaptive approaches which are based on the least–mean–square (LMS) and the recursive least–squares (RLS) algorithm, respectively, where it turns out that the latter is particularly suited for practical implementation.

In principle, real–valued ASK modulation has the drawback of being less power efficient than a corresponding complex quadrature amplitude modulation (QAM) scheme. However, since only one real dimension is used for data transmission additional degrees of freedom are available which can be exploited for interference cancellation at the receiver. We show that this advantage more than compensates for the loss in power efficiency and even significant gains are possible in an interference limited environment with respect to a conventional OFDM scheme using coded QAM modulation with the same spectral efficiency.

2. SYSTEM MODEL

In the considered scenario a mobile terminal with a single receive antenna is impaired by additive white Gaussian noise and J cochannel interferers representing surrounding base stations. The interferers are present on all subcarriers of the desired signal. Transmission is protected by convolutional coding with code rate R_c and bit-wise block interleaving over time and frequency with interleaving depth I_B. Subsequent linear modulation for the OFDM subcarriers uses real–valued coefficients for both desired and interferer signals which are assumed to employ the same modulation alphabet.

A rectangular pulse shaping filter is applied and a guard interval (GI) of sufficient duration such that intersymbol interference (ISI) can be avoided. A cyclic extension of the transmit sequence is contained in the GI, such that the corresponding discrete–time receive signal after removal of the GI can be represented by the cyclic convolution of the transmit sequence and the discrete–time channel impulse response. Thus, the discrete Fourier transform (DFT) coefficient of the ith receive signal block for subcarrier μ, $\mu \in \{0, 1, \ldots, N-1\}$ (N: DFT length) in equivalent complex baseband notation is given by

$$R_i[\mu] = H_i[\mu]A_i[\mu] + \sum_{j=1}^{J} G_{j,i}[\mu]B_{j,i}[\mu] + N_i[\mu] \ , \qquad (1)$$

where i and j, $1 \leq j \leq J$, are the OFDM symbol index and the interferer index, respectively. The discrete–time channel impulse responses comprising the effects of transmit filtering, the mobile channel and receive filtering for the desired signal and the interferer signals are assumed to be mutually independent and constant during the transmission of a data frame (corresponding to block fading). This results in discrete–frequency responses $H_i[\mu]$ and $G_{j,i}[\mu]$ for the desired signal and the interferers, respectively. $A_i[\mu]$ and $B_{j,i}[\mu]$ denote the independent, identically distributed (i.i.d.) real–valued data symbols of desired user and interferers, respectively, at time i and subcarrier frequency μ. The receiver noise is modeled by a white Gaussian process and is represented in frequency domain by $N_i[\mu]$. For (1), perfect frequency synchronization has been assumed and a symbol synchronous network, i.e., the symbol periods of useful signal and interfering signals are perfectly aligned.

To each subcarrier symbol $R_i[\mu]$, the SAIC algorithm described in the next section is applied before calculation of log–likelihood ratios of bits, deinterleaving and channel decoding.

For reference, we consider conventional OFDM transmission with convolutional coding, block interleaving and QAM modulation. Here, interference suppression is not employed as this can be accomplished only by highly complex methods for QAM and a single receive antenna such as multi–user detection algorithms or successive interference cancellation combined with successive decoding [14]. Therefore, in the QAM case a suboptimum detector with zero–forcing equalization of each subcarrier assuming ideal channel knowledge is applied. For the QAM scheme a lower code rate is applied in comparison to the ASK scheme in order to obtain the same spectral efficiency R (in bit/s/Hz), i.e., M^2–ary QAM transmission with code rate $R_c/2$ will be compared to M–ary ASK transmission with code rate R_c (M: size of ASK signal set).

In practice, additional frequency synchronization errors are present resulting from imprecision of local oscillators and Doppler shifts due to movement of the mobile terminal. In this case, the orthogonality among subcarriers no longer holds, and intercarrier interference results, cf. e.g. [15, 16]. In this paper, only static frequency synchronization errors are considered. In the following, the system model is extended correspondingly. We introduce the normalized frequency offsets ξ_s and ξ_j ($j \in \{1, \ldots, J\}$) of desired and interferer signals, respectively, with $\xi_. = \frac{\delta_.}{B/N}$, where $\delta_.$ denotes an absolute frequency offset in Hz and B is the system bandwidth.

In e.g. [17] the intercarrier interference caused by the carrier frequency offsets is determined. We assume $-1 < \xi_. < 1$ such that all possible frequency offsets are smaller than the subcarrier bandwidth B/N.

The average power of the intercarrier interference at subcarrier μ stemming from subcarrier $\bar{\mu}$ is given by [17] $\sigma_a^2|H_i[\bar{\mu}]|^2 \cdot \mathrm{sis}_N^2\left(\bar{\mu} - \mu + \xi_s\right)$ (σ_a^2: variance of $A_i[\mu]$) with

$$\mathrm{sis}_b(x) = \frac{1}{b}\frac{\sin(\pi x)}{\sin(\pi x/b)} \ , \ b \in \mathbb{N} \ , \ x \in \mathbb{R} \ . \qquad (2)$$

The cochannel interference power of interferer j with variance σ_j^2 at subcarrier μ contributed by subcarrier $\bar{\mu}$ is $\sigma_j^2|G_{j,i}[\bar{\mu}]|^2\,\mathrm{sis}_N^2(\bar{\mu} - \mu + \xi_j)$, and the resulting subcarrier CIR on subcarrier μ prior to SAIC can be expressed as

$$\begin{aligned}\mathrm{CIR}[\mu] = &\Big(\sigma_a^2|H_i[\mu]|^2 \cdot \mathrm{sis}_N^2\left(\xi_s\right)\Big)\Big/ \\ &\Big(\sum_{\bar{\mu}\in\mathcal{U}\setminus\{\mu\}} \sigma_a^2|H_i[\bar{\mu}]|^2 \cdot \mathrm{sis}_N^2\left(\bar{\mu} - \mu + \xi_s\right) \\ &+ \sum_{j=1}^{J}\sum_{\bar{\mu}\in\mathcal{U}} \sigma_j^2|G_{j,i}[\bar{\mu}]|^2 \mathrm{sis}_N^2(\bar{\mu} - \mu + \xi_j)\Big) \ ,\end{aligned} \qquad (3)$$

where $\mathcal{U}$ denotes the set of subcarriers where the desired user and the interferers are allocated.

3. INTERFERENCE CANCELLATION

In [12, 13, 6] an approach for SAIC was introduced for application in the GSM system where the cochannel interference is perfectly eliminated by complex filtering and subsequent projection of the filtered signal onto an arbitrary non–zero complex number c for the case of a single interferer. In the presence of multiple interferers 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., an MMSE criterion is applied guaranteeing interference suppression at minimum noise enhancement. The algorithm has been derived in [13, 6] for both flat fading and frequency–selective fading channels. As in an OFDM system the channel per subcarrier can be considered as flat, the version of the algorithm for flat fading is applicable to each subcarrier, and the required filter order of the complex filter for subcarrier μ is zero, i.e., a complex scalar $P[\mu]$ is employed for filtering. In [13], it has been shown that $P[\mu]$ can be selected such that the signal after projection is interference–free in the case of only one cochannel interferer. For several co–channel interferers, an MMSE solution for $P[\mu]$ is a suitable choice, which is derived in the following. We denote the real–valued output signal of projection by $Y_i[\mu]$. The error signal consisting of noise and residual interference is given by

$$E_i[\mu] = A_i[\mu] - Y_i[\mu] = A_i[\mu] - \mathcal{P}_c\{P[\mu]R_i[\mu]\} \ , \qquad (4)$$

where $\mathcal{P}_c\{x\}$ denotes projection of x onto an arbitrary non-zero complex number c and is given by

$$\mathcal{P}_c\{x\} = \frac{\mathrm{Re}\{x \cdot c^*\}}{|c|^2} \ , \qquad (5)$$

cf. [12, 13, 6] ($\mathrm{Re}\{\cdot\}$: real part of a complex number). Hence, the cost function of the MMSE approach is defined as

$$J(P[\mu]) \triangleq \mathcal{E}\left\{\left(\mathcal{P}_c\{P[\mu]\cdot R_i[\mu]\} - A_i[\mu]\right)^2\right\} \qquad (6)$$

($\mathcal{E}\{\cdot\}$: expectation operator). Exploiting the fact that $J(P[\mu])$ is convex we determine its minimum via the zeros of its derivative,

$$\frac{\partial}{\partial P^*[\mu]} J\left(P[\mu]\right) \stackrel{!}{=} 0 \ . \qquad (7)$$

For example, for $c = 1$ this results in

$$\frac{\partial J(P[\mu])}{\partial P^*[\mu]} = \Phi_{RR}[\mu]P[\mu] + \Phi_{R^*R}[\mu]P^*[\mu] - 2\,\varphi_{AR}[\mu] \stackrel{!}{=} 0 \quad (8)$$

($(\cdot)^*$: complex conjugation). The variables used in (8) are defined as

$$\begin{aligned}\Phi_{RR}[\mu] &= \mathcal{E}\left\{R_i[\mu]R_i^*[\mu]\right\}, &(9)\\ \varphi_{AR}[\mu] &= \mathcal{E}\left\{A_i[\mu]R_i^*[\mu]\right\}, &(10)\\ \Phi_{R^*R}[\mu] &= \mathcal{E}\left\{(R_i^*[\mu])^2\right\}. &(11)\end{aligned}$$

Eq. (8) can be solved for the MMSE solution $P[\mu]$ for each subcarrier. For adaptive adjustment of $P[\mu]$, the LMS and RLS algorithm, respectively, can be used. In this case, several OFDM training symbols $A_i[\cdot]$ are required. However, only the desired user's training symbols have to be known, and the algorithms perform blind adaptation with respect to the interference.

LMS Algorithm

After the training period the filter coefficients $P[\mu]$ are fixed and used for complex filtering in the current transmission frame, assuming that the channel is time–invariant during the transmission of such a frame. Using the normalized version of the LMS algorithm to allow for an adaptive LMS step size parameter we obtain the following update equation for the projection filter coefficients [18],

$$P_{i+1}[\mu] = P_i[\mu] + \frac{\tilde{\rho}}{M_x[\mu] + \epsilon} E_i[\mu] \cdot R_i^*[\mu] \ , \tag{12}$$

where $M_x[\mu]$ is the average power of the filter input signal $R_i[\mu]$,

$$M_x[\mu] = \mathcal{E}\left\{|R_i[\mu]|^2\right\} \ . \tag{13}$$

The parameter $\tilde{\rho}$ has to be chosen as $0 < \tilde{\rho} < 2$ to allow for convergence of the algorithm [18]. The variable $\epsilon \ll 1$ is a small real number required to avoid division by zero.

The convergence of the LMS algorithm is quite slow and therefore the algorithm is only suitable for long frames and low mobility of users.

RLS Algorithm

The major advantages of the RLS algorithm are an order of magnitude faster convergence than that of the LMS algorithm such that also time–variant channels can be tracked and that a lower excess MSE is obtained. As for the LMS algorithm, each subcarrier is treated independently and, hence, complexity scales linearly with the number of subcarriers. The input vector of the algorithm per subcarrier μ is defined as

$$\mathbf{U}_i[\mu] = [\mathrm{Re}\{R_i[\mu]\} \ \ -\mathrm{Im}\{R_i[\mu]\}]^T \ , \tag{14}$$

where $\mathrm{Im}\{\cdot\}$ is the imaginary part of a complex number and $(\cdot)^T$ denotes transposition. The a priori error signal of the RLS algorithm is defined as the difference of desired signal and the output of the projection of the filtered received signal,

$$\begin{aligned} E_i[\mu] &= A_i[\mu] - \mathrm{Re}\{P_{i-1}[\mu]R_i[\mu]\} \\ &= A_i[\mu] - \mathbf{U}_i^T[\mu]\mathbf{P}_{i-1}[\mu] \ , \end{aligned} \tag{15}$$

where $\mathbf{P}_{i-1}[\mu] = [\mathrm{Re}\{P_{i-1}[\mu]\} \ \mathrm{Im}\{P_{i-1}[\mu]\}]^T$. With definition of variables $\mathbf{U}_i[\mu]$, $E_i[\mu]$, and $\mathbf{P}_{i-1}[\mu]$, the RLS update equations given e.g. in [18] can be directly applied. In (15), $c = 1$ has been assumed without loss of generality[1].

4. SIMULATION RESULTS

In the following, a carrier frequency of 2 GHz is assumed, and $B = 7.68$ MHz. The DFT length is set to $N = 512$, and all subcarriers are used for transmission and are impaired by cochannel interference. The carrier–to–interference ratio (CIR) is given by C/I_t, where C and I_t are the average receive power of the desired signal and of the total interference, respectively. In order to model the interference structure of a cellular network appropriately, we took into account $J = 3$ cochannel interferers. One of the interferers dominates and has power I_d, whereas the other, residual interferers have equal average powers I_2 and I_3. The total power of the residual interference is $I_r = I_2 + I_3$, and $I_t = I_d + I_r$. The dominant–to–residual–interference ratio (DIR) is defined as I_d/I_r. The considered discrete–time channel impulse responses of desired signal and interferers have mutually uncorrelated Rayleigh fading taps with average tap powers according to an exponential power delay profile which is determined from the continuous power delay profile given in [19], $P(\tau) = e^{-\tau/\tau_0}$ for $0 \leq \tau \leq \tau_{\max} = 7\ \mu s$ and $P(\tau) = 0$ else, where $\tau_0 = 1\ \mu s$, by sampling with a sample spacing of $T_s = 130.2$ ns. A block fading model is adopted with random change of channel coefficients from frame to frame. Each frame consists of training blocks and data blocks, where each block comprises 7 OFDM symbols of duration $T = 512 \cdot T_s = 66.67\ \mu$s. Each data block is separately encoded.

The performance results for our proposed scheme are compared with results for convolutionally encoded QAM transmission according to Section II. In all cases, interleaving with depth $I_B = 32$ bits is applied. The constraint length of the used convolutional code is 9 for all code rates and schemes.

4.1. Performance in the Interference Limited Case

In the following, we consider interference limitation which is typical for a downlink scenario with a frequency reuse of one, and $E_b/N_0 = 30$ dB holds. Simulation results valid for both noise and interference limitation are discussed in [20] and omitted here due to space limitations. BLER results are shown for the analytical MMSE solution and the adaptive approaches using LMS and RLS algorithm, respectively. Results for the conventional scheme are only shown for a DIR of 0 dB. This is justified because performance of QAM transmission is approximately independent of DIR and only depends on CIR. From Figs. 1 and 2 we observe that with increasing DIR, performance of the transmission scheme with SAIC improves and significant performance gains result. The novel scheme outperforms the conventional transmission scheme for all DIRs for 8ASK transmission with $R_c = 1/2$. For 8ASK with convolutional coding with $R_c = 2/3$ the proposed scheme requires a DIR ≥ 5 dB to outperform the conventional transmission scheme. A DIR of 5 dB is a realistic value in practice assuming low shadowing correlations of different base stations. As a result of a higher diversity gain due to more powerful coding the slope of the QAM BLER curves is higher than that of the corresponding curves for ASK transmission, where a higher code rate was used to obtain the same spectral efficiency. Nevertheless, ASK outperforms QAM if a certain, reasonably low DIR is exceeded. For a DIR of 20 dB, i.e., a highly dominant cochannel interferer is present, 8ASK with $R_c = 1/2$ and $R_c = 2/3$ performs 14 dB and 10 dB better than QAM transmission, respectively.

Further analysis has shown that a training length of 21 OFDM symbols (corresponding to 3 blocks) is sufficient in order to obtain essentially the same performance with the adaptive RLS scheme as with the analytical MMSE solution for filter $P[\mu]$. This is due to the fact that the excess error induced by the RLS algorithm becomes small after a few training symbols already because of the filter length of one, resulting in almost coinciding curves for both solutions, cf. Figs. 1 and 2. For the LMS algorithm the training length had to be chosen about 10 times larger than for the RLS algorithm and still a performance loss in the order of approximately 1 dB for DIRs of 0 to 10 dB can be observed from Figs. 1 and 2. Therefore, in this case the LMS algorithm is impractical for scenarios with users

[1] In general, performance does not depend on c.

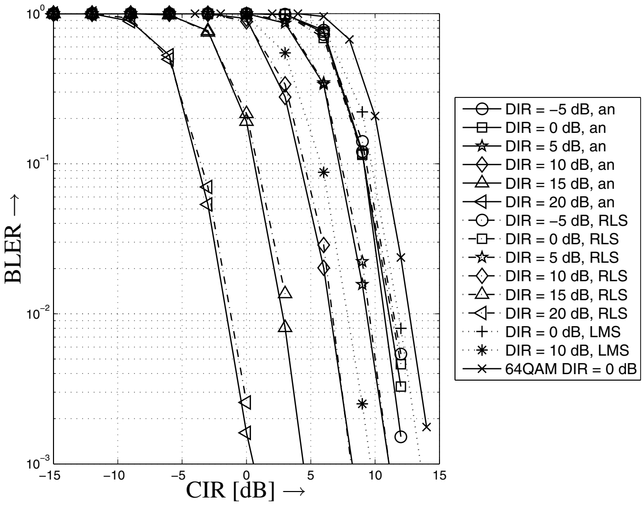

Fig. 1. BLER versus CIR for varying DIR. 8ASK with $R_c = 0.5$ and 64QAM with $R_c = 0.25$, $R = 1.5$ bit/s/Hz. "an" stands for the analytical MMSE solution.

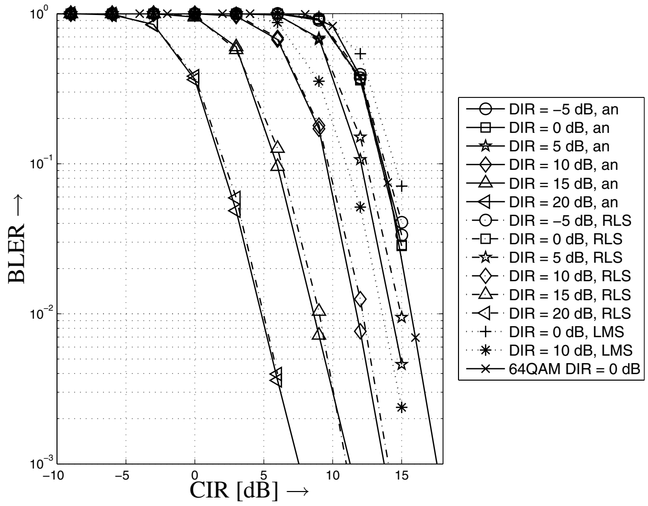

Fig. 2. BLER versus CIR for varying DIR. 8ASK with $R_c = 0.67$ and 64QAM with $R_c = 0.33$, $R = 2$ bit/s/Hz. "an" stands for the analytical MMSE solution.

of moderate–to–high mobility resulting in time–varying impulse responses.

4.2. Impact of Frequency Synchronization Errors

4.2.1. Small Frequency Offsets

We assume frequency offsets as in [13], i.e., a frequency offset of 400 Hz and 500 Hz is chosen for the desired signal and the dominant interferer, respectively, and no offsets are present for the residual interferers. The corresponding results are depicted in Fig. 3 for 8ASK transmission with code rates $R_c = 1/2$ and $R_c = 2/3$, respectively. The loss in comparison to perfect synchronization is below 1 dB for DIR values up to 15 dB, which demonstrates the robustness of the proposed scheme. In general, it can be observed that the performance loss increases with increasing DIR. This is a result of intercarrier interference induced by the frequency synchronization errors. The intercarrier interference inherently limits the maximum achievable effective DIR because now a higher number of effective interference terms is present for each subcarrier, cf. (3), and therefore particularly affects the receiver performance for high values of DIR. For example, the simulation results in Fig. 3 show a performance loss of approximately 2 dB for 8ASK transmission with $R_c = 2/3$.

4.2.2. Large Frequency Offsets

In [16], a frequency imprecision of 0.5 ppm is allowed. For a carrier frequency of $f_c = 2.0$ GHz this results in a maximum frequency offset of 1000 Hz. Assuming also Doppler shifts resulting from vehicular movement with a maximum speed of 200 km/h, an additional frequency offset of 370.4 Hz occurs. Therefore, as a worst case scenario we consider a frequency offset of 1370 Hz for the desired signal and an offset of -1370 Hz for the dominant interferer signal. The residual interferers are assigned frequency offsets of 500 Hz and 250 Hz, respectively. Simulation results for this case are illustrated in Fig. 4 for 4ASK transmission using the SAIC algorithm. As a reference, performance of the 4ASK scheme without frequency synchronization errors is also given in Fig. 4. The performance degradation is approximately 1.5 dB for a DIR of -5 dB and increases for increasing DIR due to intercarrier interference up to 3 dB for a DIR of 20 dB. Nevertheless, transmission performs reasonably well without the necessity to employ additional frequency synchronization algorithms, which are required for 8ASK in this case (results not shown).

5. CONCLUSIONS

A novel strategy for downlink OFDM transmission under presence of severe cochannel interference was presented, which combines convolutionally encoded real–valued ASK modulation with single antenna interference cancellation. Our scheme enables high downlink data rates already at low CIR values and is capable of exploiting increasing DIRs contrary to the conventional OFDM transmission scheme using QAM modulation. A comparison to the conventional approach has shown that for all modulation and coding schemes studied in this paper, the novel scheme is superior for DIRs of at least 5 dB in terms of error rate at the same spectral efficiency. For higher values of DIR gains of up to 14 dB are possible. Therefore, by exploiting the additional degrees of freedom gained by using real–valued modulation we can more than compensate for the loss in power efficiency of ASK. The adopted SAIC algorithm is blind with respect to the interference, i.e., it does not require any explicit knowledge about the interferer channels, and is moderate in terms of computational complexity. In addition, we have demonstrated that the proposed scheme is robust to errors in frequency synchronization.

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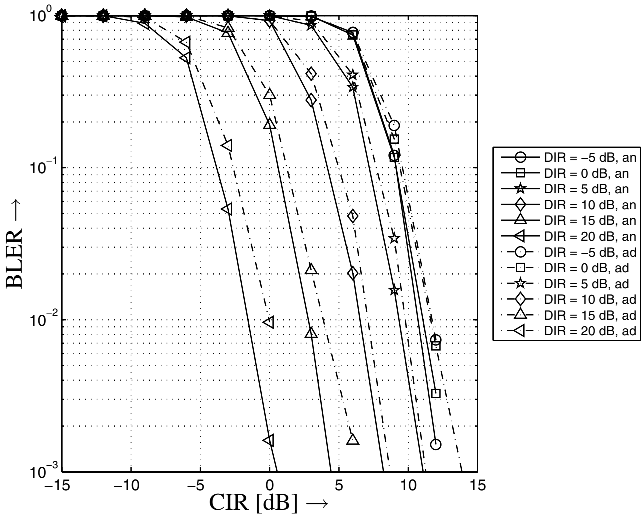

Fig. 3. BLER versus CIR for varying DIR. Small frequency offsets are present. 8ASK with $R_c = 0.5$, $R = 1.5$ bit/s/Hz. "an" stands for the analytical MMSE solution and absence of frequency offsets, "ad" stands for the adaptive RLS results.

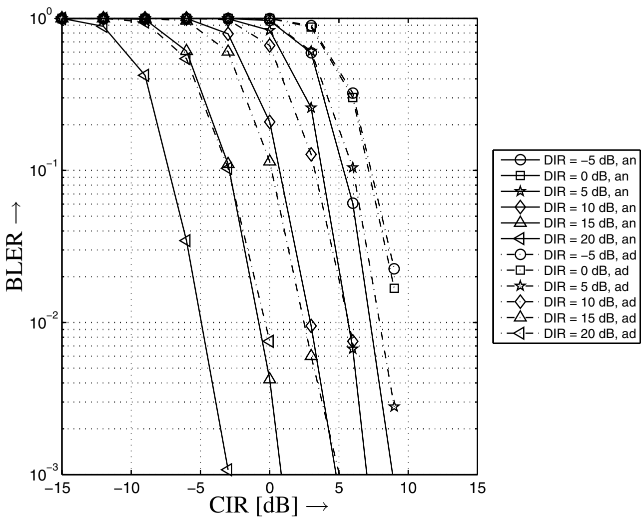

Fig. 4. BLER versus CIR for varying DIR. Large frequency offsets are present. 4ASK with $R_c = 0.5$, $R = 1$ bit/s/Hz. "an" stands for the analytical MMSE solution and absence of frequency offsets, "ad" stands for the adaptive RLS results.

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