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EMPLOYING DUAL ANTENNA INTERFERENCE SUPPRESSION TECHNIQUES IN GSM 8PSK DOWNLINK

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ABSTRACT
In 3GPP GERAN \(^1\), the performance specification of dual-antenna interference suppression techniques will be a subject of the Downlink Advanced Receiver Performance (DARP) phase II work item. In this study, we apply a filter-based single antenna interference suppression technique for downlink GSM 8PSK modulation assuming the terminal is equipped with two antennas. We evaluate the performance using a correlation-based spatial channel model and show the potential of suppressing the co-channel interference even for high correlation factors. Besides, a performance improvement can be observed if the phase angles of the correlation factors for the desired user and interferer are different in highly spatially correlated channels.

I. INTRODUCTION

In TDMA mobile radio systems like GSM/EDGE, the service area is divided into cells and the frequency resources are repeated in remote cells. Theoretically, the network capacity can be increased by decreasing the reuse factor. But this would increase the amount of interference from neighboring cells operating on the same channel, so called Co-Channel Interference (CCI). Although CCI can be reduced by some techniques like discontinuous transmission, power control, frequency hopping, etc. which have been included in the GSM standard, it remains the limiting factor for network capacity, unless CCI suppression is done.

Today the mobile terminal is usually equipped with only one receive antenna. This makes downlink CCI a particular challenging task. In [1, 2], Single Antenna Interference Cancellation (SAIC) techniques including filter-based approaches and multiuser detection techniques suitable for TDMA systems were widely discussed. For GMSK modulation, real-valued processing [3, 4] can be applied to obtain spatial diversity. The decoupled linear CCI suppression/trellis-based equalization SAIC technique encountering real-valued processing and temporal over-sampling offers a powerful solution for the GSM system with respect to computational complexity and performance [4, 5]. However, GSM 8PSK modulation occupies two dimensions per transmit symbol, thus real-valued processing is not applicable to obtain spatial diversity. Moreover, due to the small excess bandwidth in GSM, oversampling does not improve performance as well. Existing solutions for 8PSK single antenna interference suppression employing multiuser detection techniques with joint reduced-state sequence estimation have been discussed e.g. in [1, 6]. However, multiuser detection requires channel information of both desired user and interferer which is usually not available in GSM. Moreover, if the number of interferers grows, the computational complexity will become very high for practical implementation.

It is well known that interference can be better suppressed by means of antenna arrays [7, 8, 9]. In this paper, we assess the performance of a straightforward extension of SAIC [1, 5, 8] using decoupled linear filtering/trellis-based equalization similar as in [4]. In contrast to the single antenna based solution, the scheme under consideration applies complex-valued processing of the signals of two receive antennas. The linear pre-filter design is based on the MMSE criterion to maximize the Signal to Interference plus Noise Ratio (SINR) of the desired user’s signal. At the same time the channel order of the output of the prefilter is truncated, such that a full-state 8PSK equalizer employing 64 states can be used. Furthermore, we show the performance results for spatially correlated channels where correlation factors are generally complex [10]. We also show that huge gains can be achieved by the underlying scheme even for high absolute values of correlation factors. Finally, it is shown that in highly spatially correlated scenario, the phase difference between the correlation factors (corresponding to the difference of the direction of arrival of the desired signal and interference) at the receive antennas plays an important role for interference suppression.

The outline of this paper is as follows. In Section II, we describe the signal and spatially correlated channel models. Section III describes the decoupled linear CCI suppression/trellis-based equalization scheme applied in GSM 8PSK with two receive antennas. Section IV presents the simulation results. Conclusions are drawn in Section V.

II. SIGNAL AND CHANNEL MODELS

Throughout this paper, complex baseband notation is used. Consider a linear, dispersive and noisy mobile radio system for the downlink where \( J \) co-channel signals are received by an antenna array with \( M \) elements. The sampled received signal at the \( m \)-th (\( 1 \leq m \leq M \)) antenna can be modelled as

\[
y_m(k) = \sum_{j=0}^{J} \sum_{l=0}^{L} h_{m,j,l}(k) a_j(k - l) + n_m(k),
\]

where \( y_m(k) \) is the \( k \)-th symbol rate output sample at the \( m \)-th receive antenna. \( h_{m,j,l}(k) \) is the \( l \)-th channel coeffi-

\(^1\)GERAN is responsible for the specification of the radio access part of the GSM/EDGE system within the 3rd Generation Partnership Project (3GPP).
cient of the $j$-th ($0 \leq j \leq J$) co-channel user observed at the $m$-th receive antenna at time index $k$, which includes the transmit filter, physical channel, the receive filter and sampling rate. Note that some of the channel coefficients can be chosen as zeros, therefore, all co-channel users are assumed to have the same maximum channel memory $L$ with $h_{m,j}(k) = [h_{m,j,0}(k), h_{m,j,1}(k), \ldots, h_{m,j,L}(k)]$. The term $n_m(k)$ represents the $k$-th sample of the white Gaussian noise at the $m$-th receive antenna. The $k$-th i.i.d. (independent identically distributed) data symbol of the $j$-th user is denoted as $a_j(k)$, which is randomly drawn from an M-ary alphabet, here 8-ary. Moreover, $k$ stands for the time index and $K$ is number of symbols per burst. We assume channel coefficients to be time invariant within one burst, so the time index $k$ can be dropped for convenience. Without loss of generality, it is assumed here $j = 0$ refers to the desired user and $1 \leq j \leq J$ refers to the interferers in the context of interference suppression. For simplicity, we assume that only one dominant interferer is present and two receive antennas are available in the system. In practice, the signal received at the two receive antennas are usually correlated. Therefore, algorithms should be evaluated by taking the correlation factors into account. In [10], correlation factors of two types of propagation scenarios were investigated:

a) The transmitting Base Transceiver Stations (BTS) of the desired user and interferer are spaced far apart and illuminate completely different scatterers, i.e., the interference appears as inter-site interference.

b) The BTS of the desired user and interferer are closely spaced and illuminate the same scatter as for a multi-sector BTS causing intra-site interference.

In the following, focus is put on the propagation scenario type a) which happens more often in practice. In such a scenario, only correlation factors between the received signals coming from the same transmitter should be considered, which is called receive correlation in this context, while correlation factors between signals of different data sources observed at the receive antenna (transmit correlation) can be assumed zero. The receive correlation factor of two $l$-th tap channel coefficients between receive antennas $m$ and $m'$ is given by

$$\rho_{mm',l}^{Rx} := \frac{E[h_{m,j,l}h_{m',j,l}^{*}]}{\sigma_l^2}$$  (2)

where $0 \leq |\rho_{mm',l}^{Rx}| \leq 1$ for all $l$. The quantity $\sigma_l^2$ stands for the variance of $h_{m,j,l}$ irrespective of $m$ or $j$ by assuming all the transmission links experience the same physical environment. For frequency selective channels, the receive correlation factor is defined as

$$\rho_{mm'}^{Rx} := \frac{E[h_{m,j}h_{m',j}^{*}]}{\sigma^2} = \sum_{l=0}^{L} \sigma_l^2 \rho_{mm',l}^{Rx}$$  (3)

where $\sigma^2$ is the sum of the all variances $\sigma_l^2$ for $0 \leq l \leq L$ and it is normalized to one. Within the scope of this work, we assume

The CCI is suppressed by the linear filter. At the same time, the linear filter can also be designed to shorten the impulse response of the desired user channel.

- The intersymbol interference (ISI) caused by the desired user channel is suppressed by the nonlinear detector (such as trellis-based or tree based).

By separating the task of CCI suppression and ISI equalization, the linear filter can use all its degrees of freedom to minimize CCI while the subsequent nonlinear detector performs ISI equalization, which is optimum if only ISI is present in the input of the equalizer. By shortening the overall channel impulse response, the number of states of the nonlinear detector (e.g. Viterbi equalizer) is reduced, hence, the complexity reduces
correspondingly. In practice, the linear filter can be easily implemented as an add-on to the existing equalizer.

The optimization criterion for linear filter design can be chosen to maximize the Signal-to-Noise Ratio (SNR) or to minimize the Mean Squared Error (MSE). This will generally lead to solving the eigenvalue problems which may be too complex for practical implementation. Similar to [1], a prefilter can be designed without solving the eigenvalue problem. It results in minimizing the following MMSE cost function:

\[ C := E\{ w^H y - \hat{w}^H \hat{a}_0 - a_0(k - k_0) \} \]

where the vector \( y \) contains the received samples over a period of \( N \) symbols from the dual-antenna outputs and is defined as \( y = [y_1(k), y_1(k - 1), \ldots, y_1(k - N + 1), y_2(k), y_2(k - 1), \ldots, y_2(k - N + 1)]^T \). The vector \( \hat{w} \) contains the linear filter coefficients and it is denoted as \( \hat{w} = [\hat{w}_0, \hat{w}_1, \ldots, \hat{w}_N, \hat{w}_{20}, \hat{w}_{21}, \ldots, \hat{w}_{2N}]^T \). The hypotheses of the overall channel coefficients of the desired user before the equalizer is defined as \( \hat{h}^H = [h_{w1}, \ldots, h_{wL}] \), whose memory is \( L' \leq L \). The parameter \( k_0 \) is the decision delay. The vector \( \hat{a}_0 := [a_0(k - k_0 - 1), \ldots, a_0(k - k_0 - L')] \) is the data sequence of the desired user. Note that the MMSE design criterion (5) is the same as maximizing the SNR but the first tap of the overall channel impulse response for the desired user seen by the consecutive nonlinear detector is normalized to one. By taking the Wirtinger partial derivatives with respect to \( \hat{w}^H \) and \( \hat{w}^H \), the optimum filter coefficients can be obtained as

\[ \hat{w}^H = \hat{R}^{-1} \hat{R}^{H} \hat{y} \hat{a}_0 \]

where \( \hat{R}_{yy} = E\{yy^H\} \) is the autocorrelation matrix of the received symbols. \( \hat{R}_{ya} = E\{\hat{a}^H y\} \) is the cross-correlation of the desired user signal with the received symbols. \( \hat{R}_{ya} = E\{\hat{a}_0(k - k_0)^H y\} \) is a cross-correlation vector. The overall channel impulse response for the desired user seen by the consecutive nonlinear detector is

\[ \hat{h}^H = \hat{w}^H \hat{R}^H \hat{y} \hat{a}_0 \]

Conventionally, \( \hat{R}_{yy}, \hat{R}_{ya} \) and \( \hat{R}_{ya} \) are estimated using the training sequence of the desired user. However, the estimation is poor due to the limited length of the training sequence. For i.i.d transmitted data, \( \hat{R}_{yy}, \hat{R}_{ya} \) and \( \hat{R}_{ya} \) only depend on the channel coefficients and noise variance. Usually they can be better estimated if reliable channel estimates are available [1]. We restrict here our scope for the case that channel estimates are only available for the desired user, which is more realistic in practice. The estimates of \( \hat{R}_{yy}, \hat{R}_{ya} \) are then computed using channel coefficients of the desired user and the noise variance.

Due to the complex-valued signal constellation of 8PSK, real-valued processing [12] cannot be applied to obtain the spatial diversity. It is perceivable that the correlation of the antenna branches will influence performance of this decoupled linear CCI suppression/trellis-based equalization.

**IV. SIMULATION RESULTS**

In this section, we evaluate the performance of decoupled CCI suppression/trellis-based equalizer in conjunction with a dual-receive antenna for GSM 8PSK modulation using computer simulations. We consider for the downlink one desired user and one or two co-channel interferers in the system. The Typical Urban (TU) GSM channel profile (3GPP TS 45.005) [13] is used for all the users and their BTS are assumed to be placed far apart, hence interferer model in Figure 1 applies. A root raise cosine (RRC) filter is assumed as the receive filter and symbol rate sampling is performed. In addition, we adopt the block fading assumption for the simulation and the fading between bursts are assumed to be independent. Furthermore, all co-channel signals are assumed to be synchronous. The channel estimates for the desired user are obtained by correlating the received signal with the known training sequence code. No channel knowledge is assumed for the CCI. The overall channel impulse response for the desired user is truncated to \( L' = 2 \) taps unless otherwise stated, which is supplied to the subsequent 64 states Viterbi equalizer.

First of all, we assume the special case that receive correlation factors are zero, i.e., \( \rho^2 = 0 \) for all users. Figure 3 shows the raw Bit Error Rate (BER) of the desired user in the presence of one or two equally strong interferers, respectively in a spatially uncorrelated TU0 (0 km/hour) GSM channel model. We assume a signal to noise ratio of 30dB which is suitable for GSM to operate on 8PSK. The BER curve is plotted against the Carrier to Interference Ratio (CIR). As a reference, performance results for the conventional single user Maximum Likelihood Sequence Estimator (MLSE) with 512 states comprising one antenna (full-state single user Viterbi equalizer) are also included. At a BER level of 0.01 with one interferer in the system about 18dB gain can be observed compared to the conventional receiver. If two equally strong interferers are present, about 7.5dB gain can be obtained. It should be noted that the performance of the conventional receiver will be almost the same whether one or two equally strong interferers are present.
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Figure 4: Performance of decoupled CCI suppression/trellis-based equalizer with a dual-receive antenna on TU0 channel for GSM 8PSK. Receive correlation factors are assumed the same for desired user and interferer with $\rho_{Rx} = 0.1, 0.5$ and 0.9 respectively. Correlative channel estimation for desired user, synchronous GSM network, SNR=30dB.

Next, we investigate the effect of receive correlation factors on the performance of the decoupled linear CCI suppression/trellis based-equalizer.

First, the same real value for the receive correlation factor is assumed for the desired user and interferer in the simulation (phase difference between the receive correlation factors of the desired user and interferer is zero), i.e., $\rho_{Rx}(1)=\rho_{Rx}(2) = \rho_{Rx}$, where $\rho_{Rx}$ is chosen to be 0.1, 0.5 and 0.9 to represent the low, medium, and high correlation case, respectively. The raw BER of the desired user is plotted in Figure 4 for the above mentioned correlation factors against the conventional receiver. It can be seen that, as the receive correlation factor $\rho_{Rx}$ increases, the raw BER of the desired user also increases. Moreover, the raw BER performance degrades slower when $\rho_{Rx}$ increases from low to medium value than it does when $\rho_{Rx}$ increases from medium to high value. However, even for high receive correlation factor of $\rho_{Rx} = 0.9$, about 13dB gain can be obtained against the conventional receiver at a BER level of 0.01.

Second, complex correlation factors are considered. In this simulation, the absolute value of the receive correlation factors of the desired user and interferer remains to be the same, i.e., $|\rho_{Rx}| = |\rho_{Rx}(1)| = |\rho_{Rx}(2)|$, where $|\rho_{Rx}|$ is selected to be 0.1, 0.5 and 0.9. The phase differences of the correlation factors of the desired user and the interferer are assumed to be $0^\circ$, $45^\circ$, $90^\circ$ and $180^\circ$, respectively. The simulation results show that the raw BER decreases with increasing phase difference for receive correlation factors of $|\rho_{Rx}| = 0.9$ and $|\rho_{Rx}| = 0.5$. For $|\rho_{Rx}| = 0.9$, at a raw BER level of 0.01, the improvement between $0^\circ$ and $180^\circ$ phase difference is about 11dB as shown in Figure 5. For $|\rho_{Rx}| = 0.5$, the improvement reduces to around 1dB as shown in Figure 6. For $|\rho_{Rx}| = 0.1$, no improvement can be observed which is not depicted here.

Note that the phase angle of the correlation factor corresponds to the direction of arrival of the signal at the receive antennas [14]. The simulation results show that the worst case for interference suppression occurs when the signals of the desired user and interferer arrive at the receive antennas with the same direction. Moreover, especially at high correlation, interference suppression is improved when the signals of the desired user and interferer arrive from different directions.

V. CONCLUSIONS

We apply a decoupled linear CCI suppression/trellis-based equalization scheme using a dual-receive antenna in the GSM 8PSK downlink. We evaluate the performance using a correlation-based spatial channel model and show the potential gain of suppressing the co-channel interference even for highly spatially correlated channel. Additionally, the performance of the proposed scheme further improves if the phase angles of the correlation factors for the desired user and interferer are different. Such a situation occurs when their signals arrive from different directions.

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REFERENCES


Figure 6: Raw BER of decoupled CCI suppression/MLSE64 equalizer with a dual-receive antenna. The desired user and interferer have the same absolute correlation factor of $|\rho_{Rx}| = 0.5$ with varying phase difference $\Delta \arg(\rho_{Rx}) = 0^\circ, 45^\circ, 90^\circ$ and $180^\circ$.


