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Combatting Co-Channel Interferers in a TDMA System Using Interference Estimates From Adjacent Frames

Erik Lindskog
Signal Processing Group, Uppsala University, Sweden
Presently visiting: Northeastern University
Department of Electrical and Computer Engineering
409 Dana, Boston, MA 02115, USA
Tel/Fax: +1-617-373-4029/8970 e-mail: lindskog@cdsp.neu.edu

Abstract

Suppression of co-channel interferers in a TDMA cellular phone system by the use of an antenna array is studied. Especially the problem of suppressing co-channel interferers that appear outside the training sequence is addressed. It is observed that if a co-channel interferer is present during the data sequence but not during the training sequence of a frame, then it will be present during the training sequence of an adjacent frame. The interferer plus noise spectrum of adjacent frames is utilized in order to suppress such co-channel interferers. The effect on the performance is illustrated with an example scenario for the GSM system.

1 Introduction

The problem addressed in this paper is the suppression of co-channels interferers in a TDMA system, by the use of an antenna array. The GSM system is used as an example. Here data is transmitted in frames, with a known training sequence in the middle. During the training sequence, an adaptive antenna at the receiver can be tuned in order to suppress co-channel interferers while receiving the desired signal. A problem however occurs with co-channel interferers that are not transmitting during the current training sequence but *are* transmitting during the current data sequence. See for example co-channel interferer 2 in Figure 2. An antenna array tuned to suppress the co-channel interferers present during the training sequence may then have a severely degraded performance. The idea in this paper is to use spatial spectral information from adjacent time frames in order to suppress such interferers.

2 Algorithm outlining

An indirect method is proposed for the tuning of the weights in the antenna array. First the FIR channels, $B_i(q^{-1})$, i=1,2,...,M, from the transmitted sequence to each of the antenna elements (se Figure 1), are identified. This identification is performed based on the data received during the training sequence.

The identified channels can be used for estimating the spatial spectrum of the desired signal while the residuals from the identification procedure can be used in order to estimate the spatial spectrum of the interference plus noise.

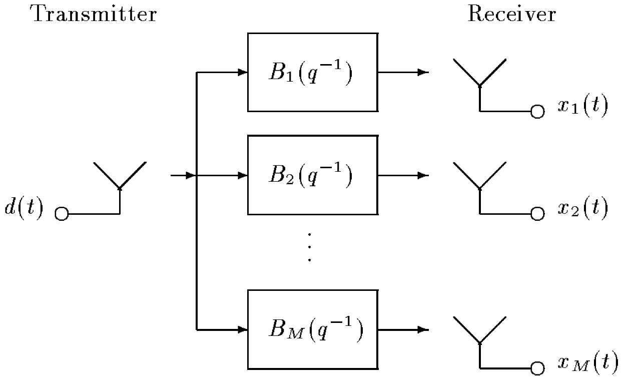

Figure 1: Discrete-time baseband channels, $B_i(q^{-1})$, from transmitted symbols, $d(t)$, to received signals $x_i(t)$.

If a co-channel interferer is not present during the current training sequence it will be present in an adjacent training sequence, see Figure 2. Antenna arrays are for practical reasons mainly of interest at the base stations. This scheme is therefore only considered for the base station receiver. In this case, the spatial spectrum of the interference plus noise during the adjacent training sequence will either be known from a previous identification procedure (receiving from a different mobile however) or will anyway have to be computed for the subsequent training sequence. When tuning the beamformer we can then do two different computations, one for the part of the data before the training sequence and one for the part of the data after the training sequence. For each tuning, we can use both the spatial spectrum of the interference computed during the current training sequence and the spatial spectrum of the interference computed during the adjacent training sequence. In this way, all co-channel interferers present during each data sequence part, will be accounted for.

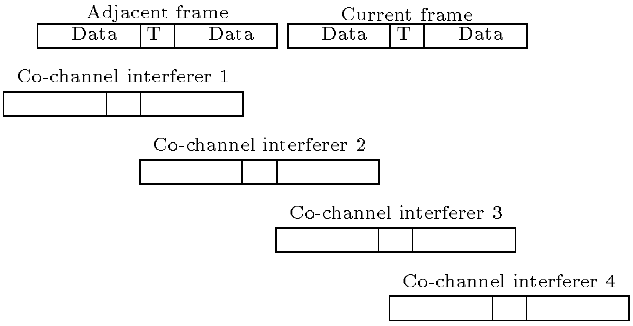

Figure 2: Adjacent time frames and example of locations co-channel interferers in GSM. The training sequence is denoted by "T" and the data sequences are denoted by "Data". Note the examples of locations for the co-channel interferers. Co-channel interferer 2 is, for example, not present during the current training sequence, but it is present during the first data sequence of the current frame.

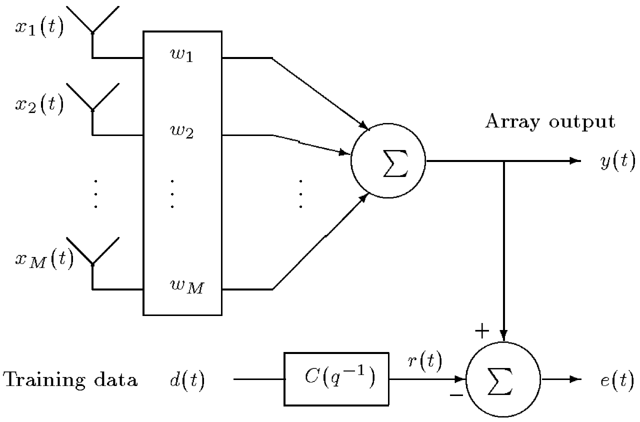

Figure 3: The antenna array and the weight coefficients to be tuned.

3 Algorithm details

The algorithm utilized is similar to the standard sample matrix inversion method, see for example [1]. Instead of estimating the covariance matrix of the received signal by means of the sample covariance matrix we will estimate it based on identified channels and the sample covariance matrices of the residuals for the current and the adjacent training sequences.

First the channels $B_i(q^{-1}) = b_{i,0} + b_{i,1}q^{-1} + \ldots + b_{i,nb} + q^{-nb}$, $i = 1, 2, ..., M$ to each of the antenna elements, are identified with the standard least squares method

$$B_i = (\Phi^H\Phi)^{-1}\Phi^H D \quad ,i = 1, 2, .., M \tag{1}$$

where

$$B_i = [b_{i,0}\, b_{i,1} \, \ldots \, b_{i,nb}]^T \tag{2}$$

and

$$\Phi = \begin{bmatrix} x_i(nb) & x_i(nb-1) & \ldots & x_i(0) \\ x_i(nb+1) & x_i(nb) & \ldots & x_i(1) \\ \vdots & \vdots & \vdots & \vdots \\ x_(N-1) & x_i(N-2) & \ldots & x_i(N-nb) \end{bmatrix}$$

$$D = [d_0\, d_1 \, \ldots \, d(N-1)]^T \tag{3}$$

The parameter N is the number of symbols in the training sequence. Here $N = 26$ is used. The antenna array with the beamformer coefficients are depicted in Figure 3. The beamformer weights, w_i, where

$$B_i = [b_{i,0}\, b_{i,1} \, \ldots \, b_{i,nb}]^T \tag{4}$$

and

$$\Phi = \begin{bmatrix} x_i(nb) & x_i(nb-1) & \ldots & x_i(0) \\ x_i(nb+1) & x_i(nb) & \ldots & x_i(1) \\ \vdots & \vdots & \vdots & \vdots \\ x_(N-1) & x_i(N-2) & \ldots & x_i(N-nb) \end{bmatrix}$$

$$D = [d_0\, d_1 \, \ldots \, d(N-1)]^T \tag{5}$$

The parameter N is the number of symbols in the training sequence. Here $N = 26$ is used. The antenna array with the beamformer coefficients are depicted in Figure 3. The beamformer weights, w_i, i=1,2,...,M, are tuned to optimally receive the reference signal

$$r(t) = C(q^{-1})d(t) \tag{6}$$

which is an estimate of the transmitted signal. The filter $C(q^{-1})$ models the GMSK modulation used in GSM, see [2] and [3]. Here, $C(q^{-1}) = 0.44i + 1.00q^{-1} - 0.44iq^{-2}$ is used. This filter is a simple model of the channel between transmitted symbols and received samples. The model includes the GMSK modulation (see [2]) and a model of a receiver filter as a 4th order Butterworth lowpass baseband filter. The correct model of the received samples after the GMSK modulation will vary depending on the chosen sampling instant. The chosen model represents only one particular sampling instant. Multipath propagation is not included in the model, i.e. the physical radio channel is modeled as a perfect channel with unit response. For a more detailed description of the model used see [3].

The beamformer weights are chosen as

$$w = (\hat{R}_{xx}^{-1}\hat{R}_{xr})^* \tag{7}$$

where $(\cdot)^*$ denotes elementwise complex conjugation, and

$$w = [w_1\, w_2 \ldots w_M]^T \; . \tag{8}$$

In (7), $\hat{R}_{xx}$ and $\hat{R}_{xr}$ are estimates of the covariance matrices

$$R_{xx} = E[x(t)x^H(t)] \tag{9}$$

$$R_{xr} = E[x(t)r^H(t)] \tag{10}$$

where

$$x(t) = [x_1(t)\, x_2(t) \ldots x_M(t)]^T \tag{11}$$

is the input to the array and $r(t)$ is the reference signal.

The covariance matrix estimate $\hat{R}_{xr}$ is formed based on the identified channels as

$$R_{xr} = BC^H \tag{12}$$

where

$$B = \begin{bmatrix} b_{1,0} & b_{1,1} & \cdots & b_{1,nb} \\ b_{2,0} & b_{2,1} & \cdots & b_{2,nb} \\ \vdots & \vdots & \vdots & \vdots \\ b_{M,0} & b_{M,1} & \cdots & b_{M,nb} \end{bmatrix} \tag{13}$$

and

$$C = [0.44\mathrm{i}\ 1.00\ \ -0.44\mathrm{i}\ 0\ \ldots\ 0] \tag{14}$$

The covariance matrix estimate $\hat{R}_{xx}$ is partitioned into a desired signal part, $\hat{R}_{ss}$, and an interference plus noise part, $\hat{R}_{nn}$,

$$\hat{R}_{xx} = \hat{R}_{ss} + \hat{R}_{nn}\ . \tag{15}$$

The part $\hat{R}_{ss}$ is formed from the identified channels as

$$R_{ss} = BB^H \tag{16}$$

while $\hat{R}_{nn}$ is formed from the residuals as

$$\hat{R}_{nn} = \frac{3}{4N}\sum_{t_c=1}^{N} n_c(t_c) n_c^H(t_c) + \frac{1}{4N}\sum_{t_a=1}^{N} n_a(t_a) n_a^H(t_a)\ . \tag{17}$$

Above, $n_c(t)$ represents the residuals for the current training sequence whereas $n_a(t)$ constitutes the residuals for the adjacent training sequence. The time indices t_c and t_a belong to the current and the adjacent training sequences respectively, while the parameter N is the number of symbols of the training sequence. Note that only one of the sums in equation 17 has to be computed for each training sequence. If an interferer is present during the current training sequence, it is more likely to be present during the current data sequence part in question, than if it was present during the adjacent training sequence. The contributions from the different training sequences are therefore weighted differently. Other weighting factors can certainly be considered.

For comparison we use a version of the algorithm that does not take into account the interference plus noise spectrum of the adjacent time frame. Here $\hat{R}_{nn}$ is simply

$$\hat{R}_{nn} = \frac{1}{N}\sum_{t_c=1}^{N} n_c(t_c) n_c^H(t_c) \tag{18}$$

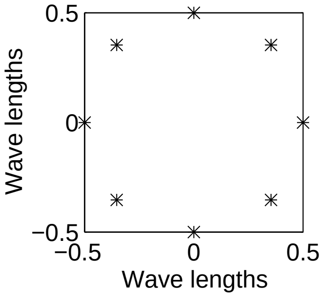

Figure 4: Antenna configuration.

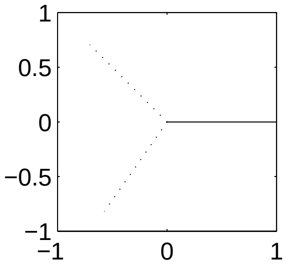

Figure 5: Desired signal (solid) and co-channel interferers present during the training sequence (dashed). The angle of incident of the adjacent frame interferer (dotted) is varied between -180 and 180 degrees.

4 Simulations

An example scenario is chosen in order to illustrate the behavior of the algorithms. The desired signal and co-channel interferers present during the training sequence is shown in Figure 5. The received desired signal and co-channel interferers are modeled by filtering binary symbols, ±1, through the three tap FIR filter $M(q^{-1}) = 0.44i + 1.00q^{-1} - 044iq^{-2}$. This filter is, as described in the previous section, a model of the channel between the transmitted symbols and the received samples. In reality, the true channel would vary depending on the sampling instant and possible intersymbol interference due to multipath propagation. This problem can, for example, be handled with the method described in [3].

The desired signal impinges onto the antenna from 0 degrees and two equal strength co-channel interferers impinge from -30 and -60 degrees respectively. A third co-channel interferer is used in the simulations. This co-channel interferer is thought of as not being present during the training sequence of the desired signal in question. It will either be present or not be present during the desired signals data sequence. It is however present during the training sequence of the adjacent frame. This co-channel interferer will be referred to as the adjacent frame interferer.

The SNR was 3 dB and the SIR was 0 dB, not counting the adjacent frame interferer. This co-channel interferer had a power twice that of the indi-

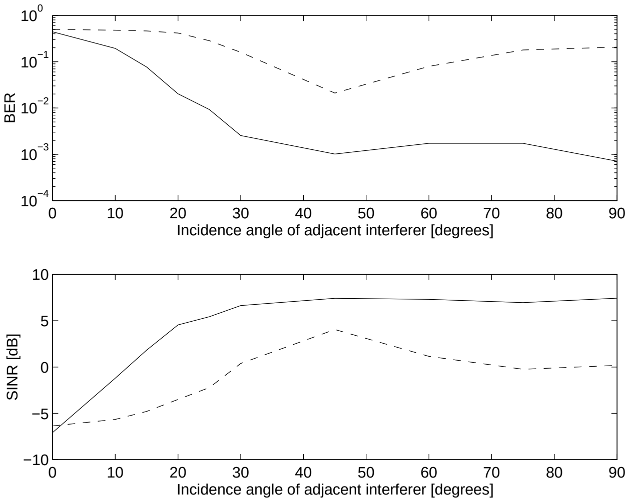

Figure 6: Adjacent frame interferer present during data sequence. BER for the MLSE and SINR after the beamformer. The algorithm using adjacent frame interference plus noise (solid) and the algorithm not using adjacent frame interference plus noise (dashed).

vidual interferers that were present during the current training sequence.

Two performance measures were evaluated for the algorithms. One was the BER for a MLSE after the beamformer, working with a three tap FIR channel. Another performance measure was the signal to interference and noise ratio (SINR) after the beamformer.

The algorithms were tested for two different cases. In the first case, the adjacent frame interferer was present during the whole current data sequence part, thus causing interference.

In the second case, the adjacent frame interferer was *not* present during the current data sequence part. This situation has to be considered because it is undesired that the algorithm using the adjacent interferer plus noise spectrum should be penalized too much in this case.

The performance was evaluated when the impinging direction of the adjacent time frame interferer was varied between -180 and 180 degrees. The resulting BER:s and SINR:s can be seen in Figures 6 and 7.

Examples of the antenna gain patterns for the two algorithms can be seen in Figures 8 and 9. Figure 8 shows a bad case for the algorithm that does not take the adjacent interferer into account. The adjacent frame interferer is amplified by the antenna array causing a low SINR. In Figure 9 the effect of taking the adjacent noise plus interferer spectrum into account can be seen. The adjacent frame interferer is now nulled out, resulting in a much better SINR.

In Figure 10, the BER and SINR can be seen for a simulation where only the angle of incidence of the adjacent frame interferer has been varied. The antenna gains for the two different algorithms are as in Figures 8 and 9. The adjacent frame interferer was

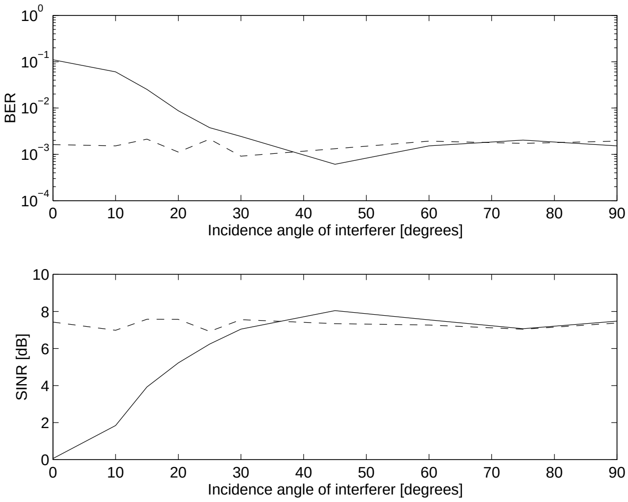

Figure 7: Adjacent frame interferer *not* present during data sequence. BER after the MLSE and SINR after the beamformer for the algorithm using adjacent frame interference plus noise (solid), dito *not* using adjacent frame interference plus noise (dashed)

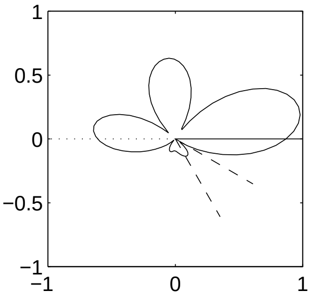

Figure 8: Antenna gain when the adjacent interferer is not accounted for. Solid line - Desired signal, Dashed lines - Co-channel interferers and Dotted line - Adjacent frame interferer.

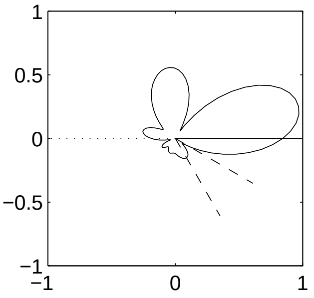

Figure 9: Antenna gain when the adjacent interferer is accounted for. Solid line - Desired signal, Dashed lines - Co-channel interferers and Dotted line - Adjacent frame interferer.

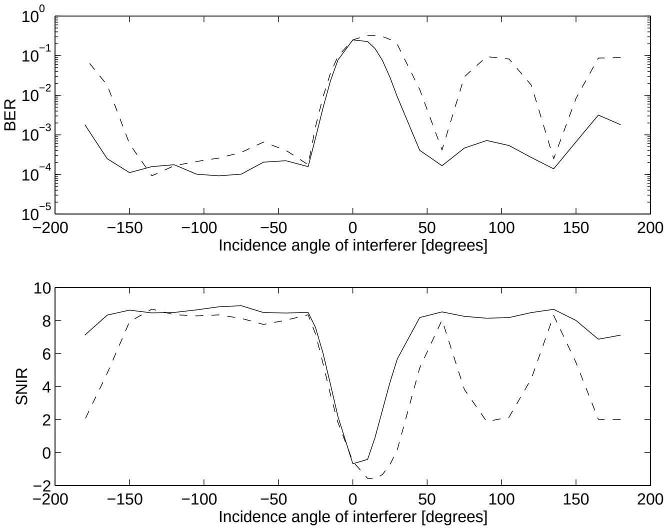

Figure 10: Adjacent interferer present during data sequence. BER after the MLSE and SINR after the beamformer for the algorithm using adjacent frame interference plus noise (solid), dito not using adjacent frame interference plus noise (dashed). The BER and SNIR was computed for the specific antenna gain patterns of Figures 8 and 9.

present during tha data sequence.

5 Results and Conclusion

As can be seen in Figure 6, for the scenario considered, using the interference plus noise spectrum from the adjacent time frame is advantageous if the interferer actually is present during the data sequence part of interest.

If, however, the interferer is *not present* during the data sequence part of interest, as in Figure 7. Then the proposed algorithm suffers a performance degradation when the angle difference between the accounted for, but not present, interferer and the desired signal is small. The reason for this is that the gain in the direction of the *desired* signal will then be somewhat reduced. However, if the angle difference between the impinging directions of the interferer and the desired signal is large, then we see from Figure 7 that one does not loose any performance by taking the not present interferer into consideration. This will however not be completely true in general. If the adjacent interferers accounted for are not present, and if the number of strong interferers to be nulled out are larger than the degrees of freedom of the antenna, then the proposed algorithm will likely suffer a performance degradation.

The general idea presented here can be applied to other beamforming schemes as long as they separate between signal and interference plus noise spectrums. For instance, one can use beamforming with the maximum SNR method. See for example [4].

Some improvement can be made of the method. When working on the first half of the data in the frame, the adjacent noise spectrum could be formed using the last part of the data sequence of the adjacent frame. The previously estimated data symbols would then be used instead of the training sequence. The same improvement cannot however be performed for the second half of the data in the frame.

References

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[3] E. Lindskog, “Making SMI-beamforming insensitive to the sampling timing for GSM signals,” in *Proceedings of the Sixth International Symposium on Personal, Indoor and Mobile Radio Communications*, Toronto, Canada, 27-29 September 1995, pp. 664–668.

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