EXPLOITING THE CYCLIC PREFIX FOR BEAMFORMING IN OFDM RECEIVERS

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ABSTRACT

Blind and semiblind antenna array algorithms are proposed for Orthogonal Frequency Division Multiplexing (OFDM) systems in the time domain. A blind technique is proposed for the maximization of the temporal Signal to Noise plus Interference Ratio (SNIR) in frequency-selective channels exploiting the Cyclic Prefix (CP). In OFDM systems, this maximization does not imply the minimization of the Bit Error Rate (BER). Thus, a semiblind method that takes advantage of both known training data and the CP is obtained, which provides the best results in terms of link quality. Simulations are conducted for Hiperlan/2 (HL/2) parameters, and a slight modification of its fixed pilot position is proposed, so as to improve the BER performance of the algorithms and to obtain estimates of the channel for the blind method.

1 INTRODUCTION

OFDM has been standarized as the modulation technique for several major standards comprising from broadcasting services to broadband Wireless Local Area Networks (WLAN) standards, among them ETSI BRAN HL/2 [1]. The performance of these systems can be considerably improved by the use of antenna arrays at the receiver side in a standard-compliant Single Input Multiple Output (SIMO) configuration, either in the time domain [2] or in the frequency domain [3]. Time domain algorithms applied before the FFT at the receiver improve the performance of post-FFT approaches [4], especially in channels with high delay spread.

OFDM standards usually provide known training data which may be used to compute the beamformer, often in form of a preamble. However, in highly variant scenarios, if the techniques are based exclusively on this data, there will be no reaction against non-stationary interferences that may appear during the burst. This is the case of WLAN systems, and the solutions could consist of implementing an adaptive algorithm [5] or using the blind method proposed in Section 3.1. By symbol by symbol updates, the semiblind algorithm proposed in Section 3.2 may also be suitable.

In this paper, a blind method that maximizes the SNIR at the output of the spatial filter is presented. Based on the redundancy introduced by the CP in the time domain, the method computes a broadband beamformer for frequency selective channels. The performance can be further optimized by the use of the known training data. Without loss of generality, if the interferences are considered to be fully synchronized with the user data, the combination of the traditional Sample Matrix Inversion (SMI) algorithm [6] during the preamble and the CP redundancy of the transmitted symbols within a burst can provide an important gain to the ultimate link quality (SMI-CP solution). While in the blind method the time diversity introduced by the CP is used to discriminate a desired signal against undesired signals as existing techniques in the literature (e.g. [7] and references therein), in the SMI-CP approach the CP redundancy is exploited to outperform existing conventional methods. The methods proposed in this paper compute a beamformer for the whole burst, but an adaptive design is also possible.

In the following, Section 2 presents the signal model and shows the receiver structure. Section 3 presents a blind array technique for the maximization of the SNIR and a semiblind method designed for OFDM systems with known training data (preamble). Both methods compute a broadband beamformer so as to deal with frequency selective channels [8]. Section 4 is devoted to practical HL/2 channel estimation issues while Section 5 describes the simulations and discusses the results immediately before the final conclusions.

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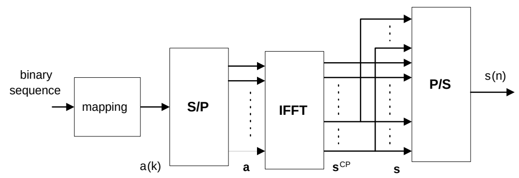

Figure 1: OFDM baseband transmitter (no channel coding)

2 SYSTEM MODELING

In the folowing, boldface capital letters refer to matrices and lowercase boldface letters refer to vectors. The operator $(\cdot)^*$ denotes conjugation, $(\cdot)^T$ transposition, and $(\cdot)^H = ((\cdot)^*)^T$. The $N \times N$ unitary Fourier matrix is denoted by $\mathbf{F}$ and $\mathbf{F}^H$ refers to the IFFT_N operation.

2.1 Signal Model

At the transmitter (Figure 1), the mapped symbols feed the Serial to Parallel (S/P) block to perform the IFFT_N, i.e. $\mathbf{s}^{CP}(m) = \mathbf{F}^H \mathbf{a}(m)$, where the vector $\mathbf{a}(m)$ has been filled

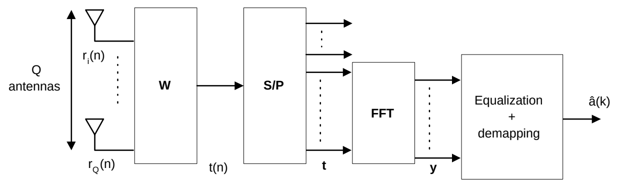

Figure 2: Time diversity receiver for HL/2

with the mapped symbols with the corresponding padding zeros at the unused subcarriers. Then, the transmitted OFDM symbol m in the time domain with a cyclic prefix of L samples is $\mathbf{s}(m)=[s(m,0)\,s(m,1)\ldots s(m,P-1)]^T$, where $s(m,n) = s(m,n+N-1)$, $0 \leq n \leq L-1$ and $s(m,L) = s^{CP}(m,0)$. The total number of samples is $P = N+L$.

Let $\mathbf{h}_q(m)=[h_q(m,0)\;\; h_q(m,1)\;\ldots\; h_q(m,K-1)]^T$ be the channel impulse response during OFDM symbol m of a given subchannel q, i.e. the channel from the single transmitter antenna to the receiver antenna q. The received OFDM symbol m in time domain can be expressed as

$$\mathbf{r}_q(m) = \mathbf{H}_q(m)\mathbf{s}(m) + \mathbf{v}_q(m), \tag{1}$$

where $\mathbf{v}_q(m)$ contains the contribution from the AWGN, the interference sources and also the Inter Symbol Interference (ISI) from the previous symbol, while $\mathbf{H}_q(m)$ is the $P \times P$ Toeplitz filtering matrix with first row $[h_q(m,0)\;\; 0\;\ldots\; 0]$ and first column $[\mathbf{h}_q(m)^T\;\; 0\;\ldots\; 0]^T$. The channel is considered time-invariant within a OFDM symbol.

2.2 Signal Structure

The insertion of a CP allows the transmitted OFDM symbol to be partitioned into P/L equal size blocks of L samples, where P is assumed to be a multiple of L. In turn, the received symbol can also be partitioned to exploit the temporal redundancy (see Section 3). Then, the received $L \times 1$ signal vector for the ith block during symbol m is expressed as

$$\mathbf{r}_q^i(m) = \mathbf{C}_q(m)\mathbf{s}^i(m) + \tilde{\mathbf{v}}_q^i(m) \qquad 0 \leq i \leq Nl, \tag{2}$$

where $\mathbf{C}_q(m)$ is the $L \times L$ Toeplitz filtering matrix with first row $[h_q(m,0)\;\; 0\;\ldots\; 0]$ and first column $[\mathbf{h}_q(m)^T\;\; 0\;\ldots\; 0]^T$, i.e. the upper left submatrix from $\mathbf{H}_q(m)$. In turn, $\tilde{\mathbf{v}}_q^i(m)$ contains the contribution from the noise and interferences, as well as the ISI from previous received signal block. Note that this ISI term could be significant. Finally, $Nl = \frac{P}{L} - 1$ is the index of the last block.

2.3 Receiver Structure

The time diversity receiver (Figure 2) performs the beamforming in the time domain as a filter bank, where each FIR filter for OFDM symbol m has M coefficients, i.e. $\mathbf{w}_q(m) = [w_q(m,0)\;\; w_q(m,1)\;\ldots\; w_q(m,M-1)]^T$. The filter vector $\mathbf{w}(m) = [\mathbf{w}_1(m)^T\;\; \mathbf{w}_2(m)^T\;\ldots\; \mathbf{w}_Q(m)^T]^T$ and $\mathbf{r}(m,n) = [\tilde{\mathbf{r}}_1(m,n)^T\;\; \tilde{\mathbf{r}}_2(m,n)^T\;\ldots\; \tilde{\mathbf{r}}_Q(m,n)^T]^T$ are used to compute output sample n as $t(m,n) = \mathbf{w}(m)^H\mathbf{r}(m,n)$. The received data vector containing the necessary samples to perform the filtering for antenna qth is $\tilde{\mathbf{r}}_q(m,n) = [r_q(m,n)\;\; r_q(m,n+1)\;\ldots\; r_q(m,n+M-1)]^T$. In the following, $\mathbf{r}^i(m,n)$ denotes the $MQ \times 1$ vector containing the $M-1$ samples of the Q antennas that follow sample n of the ith part in the received OFDM symbol m. An index of a sample exceeding $L-1$ refers to samples from next signal block, i.e. if $R > L-1$, $r_q^i(m,R) = r_q^{i+1}(m,R-L)$.

The output samples can be gathered up in a common vector $\mathbf{t}(m) = [t(m,0)\;\; t(m,1)\;\ldots\; t(m,P-1)]^T$. The cyclic prefix is disregarded by the matrix $\mathbf{G}$, therefore $\mathbf{y}(m) = \mathbf{FGt}(m)$. In the sequel, $\mathbf{y}_u(m)$ denotes the $Nu \times 1$ vector containing only the useful subcarriers. If the global response of the channel and the spatio-temporal filter is shorter than the cyclic prefix, the mapped symbols $a(m,k)$ can be recovered by simple equalization in the frequency domain, i.e. $\hat{a}(m,k) = y_u(m,k)/\hat{c}(k)$, where $\hat{c}(k)$ denotes the estimated total response coefficient for subcarrier k.

3 BLIND AND SEMIBLIND RECEIVERS

Both the blind and semiblind algorithms compute a beamformer for the whole burst of M_{pl} OFDM symbols, i.e. $\mathbf{w}(m) = \mathbf{w}$ for $0 \leq m \leq M_{pl}-1$. The former is a non data-aided method that takes advantage of the OFDM signal redundancy in time domain, whereas the latter relies on this redundancy and also on known training data.

3.1 Blind maximization of SNIR

In [7] a similar technique was applied for Frequency Diversity Spread Spectrum (FDSS) systems and frequency non-selective channels in the frequency domain (one filter tap per antenna). Here a wideband temporal beamformer is required for OFDM due to the channel frequency selectivity.

The cyclic prefix implies that first and last signal blocks of OFDM symbol m, i.e. $\mathbf{s}^0(m)$ and $\mathbf{s}^{Nl}(m)$, are equal. Exploiting the cyclic redundancy at the receiver side, see (2), a beamformer can be computed by the minimization of the Mean Square Error (MSE) between the output samples from the equal temporal transmitted blocks, according to

$$\mathbf{w} = \arg\min_{\mathbf{w}} E\left\{|\mathbf{w}^H\mathbf{r}^0(m,n) - \mathbf{w}^H\mathbf{r}^{Nl}(m,n)|^2\right\} \tag{3}$$
$$\text{subject to } \Re e\left\{E\left\{\mathbf{w}^H\mathbf{r}^0(m,n)\mathbf{r}^{Nl}(m,n)^H\mathbf{w}\right\}\right\} = \varphi.$$

The constraint is imposed to avoid the all-zeros solution. Since the noise and interferences are assumed to be uncorrelated, the optimum combiner in (3) will select only the signal that is exact in both time blocks, i.e. the desired signal. The problem in (3) is equivalent to the maximization of the output temporal SNIR defined as

$$\text{SNIR} = \frac{P_d}{P_{ni}} = \frac{\mathbf{w}^H\Phi_d\mathbf{w}}{\mathbf{w}^H\Phi_{ni}\mathbf{w}}, \tag{4}$$

where the matrices Φ_d and Φ_{ni} are the $QM \times QM$ covariance matrices from the desired signal and from the noise and interferences respectively. The maximization of (4) is equivalent to the maximization of the Rayleigh quotient $\frac{\mathbf{w}^H\Phi_d\mathbf{w}}{\mathbf{w}^H\Phi_r\mathbf{w}}$. In this case, Φ_r denotes the covariance matrix of the received signal, which is in fact much easier to estimate than Φ_{ni}. The solution is then given by the generalized eigenvector $\mathbf{w}$ associated to the maximum eigenvalue λ_{max} of

$$\Phi_d\mathbf{w} = \lambda_{max}\Phi_r\mathbf{w}. \tag{5}$$

The solution of (3) leads to a problem analogous to (5) where the covariance matrices shall be estimated by

$$\hat{\Phi}_d = \frac{1}{M_{pl}}\sum_{m=0}^{M_{pl}-1}\hat{\Phi}_d(m), \qquad \hat{\Phi}_r = \frac{1}{M_{pl}}\sum_{m=0}^{M_{pl}-1}\hat{\Phi}_r(m), \tag{6}$$

where

$$\hat{\Phi}_d(m) = \frac{1}{2L}\sum_{n=0}^{L-1}\Big(\mathbf{r}^0(m,n)\mathbf{r}^{Nl}(m,n)^H + \mathbf{r}^{Nl}(m,n)\mathbf{r}^0(m,n)^H\Big), \quad (7)$$

$$\hat{\Phi}_r(m) = \frac{1}{2L}\sum_{n=0}^{L-1}\Big(\mathbf{r}^0(m,n)\mathbf{r}^0(m,n)^H + \mathbf{r}^{Nl}(m,n)\mathbf{r}^{Nl}(m,n)^H\Big). \quad (8)$$

Although here only block solutions are considered, this method could be applied to shorter blocks than the whole user burst. In that case, interferences that appear within the burst and affect only some OFDM symbols could be rejected.

3.2 SMI-CP method

The SMI-CP semiblind beamformer is computed for the whole burst, but taking now also advantage from the preamble. The cost function is now a combination of (3) and the SMI cost in time domain, i.e.

$$J_{SBL} = J_{BLI} + J_{SMI}. \quad (9)$$

$$J_{BLI} = E\left\{|\mathbf{w}^H\mathbf{r}^0(m,n) - \mathbf{w}^H\mathbf{r}^{Nl}(m,n)|^2\right\}, \quad (10)$$

$$J_{SMI} = E\left\{|\mathbf{w}^H\mathbf{r}(n) - s(n)|^2\right\}. \quad (11)$$

SMI considers only the preamble data $s(n)$ as the known reference. Note also that during the preamble (11) the OFDM symbol index m is omitted for notation simplicity. Taking derivatives of (9) with respect to the spatial filter,

$$\mathbf{w} = (\Phi_{ni} + \Phi_R)^{-1}\,\mathbf{p}, \quad (12)$$

where $\Phi_{ni} = \Phi_r - \Phi_d$ is the noise and interference covariance matrix. Then Φ_r and Φ_d are estimated as in (6) while Φ_R and $\mathbf{p}$ refer to the preamble of N_{pr} samples

$$\hat{\Phi}_R = \frac{1}{N_{pr}}\sum_{n=0}^{N_{pr}-1}\mathbf{r}(n)\mathbf{r}(n)^H, \quad \hat{\mathbf{p}} = \frac{1}{N_{pr}}\sum_{n=0}^{N_{pr}-1}\mathbf{r}(n)s^*(n). \quad (13)$$

The main advantages of SMI-CP vs. SMI are:

1. According to a typical HL/2 user burst of 50 OFDM symbols, more samples are taken into account, thus stabilizing the estimations of the covariance matrices, resulting in a higher output SNIR [9].
2. If interferences appear after the preamble, they shall be cancelled by the SMI-CP beamformer.

In fact, if the covariance matrices are assumed to be ideal, it can be easily shown that for high SNIR, both solutions yield to the same beamformer. On the other hand, for low SNIR, the difference is only a scale factor, which has no importance due to equalization. The BER improvements showed in simulations are achieved due to practical issues, e.g. previous point 1.

4 CHANNEL ESTIMATION

As stated in Section 2.3, equalization is performed in the frequency domain. If a preamble is provided (denoted explicitly by subindex p) the channel estimates for used subcarrier k can be obtained by the averaging of the channel coefficients over the filtered M_p symbols of the preamble after the FFT,

$$\hat{c}(k) = \frac{1}{M_p}\sum_{m=0}^{M_p-1}\frac{y_{u_p}(m,k)}{a_p(m,k)} \qquad 0 \le k \le Nu-1. \quad (14)$$

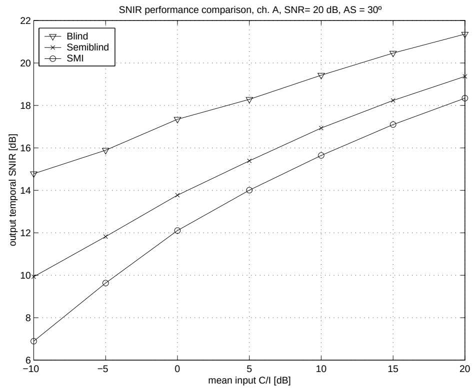

Figure 3: Output SNIR performance vs input C/I

As simulations are conducted in HL/2 environments, a slight modification of the pilot structure is needed to perform a pilot-based channel estimation, needed especially in the blind method. In HL/2 the pilots are fixed, located at subcarrier indexes $A = \{5, 19, 32, 46\} \subset \{0, .., 51\}$. So as to obtain the channel response, a moving pilot structure is proposed. The position of the pilots shall change at each transmitted OFDM symbol m of the burst as in DRM [10], e.g. $A(m) = \{0 + m_{13}, 13 + m_{13}, 26 + m_{13}, 39 + m_{13}\}$, where m_{13} indicates the remainder after the integer division of m and 13. The total response coefficient for subcarrier k is then averaged over the estimations collected during the burst. It is important to notice that the channel time variation shall be relatively slow. Note that this estimation is well-suited for channels with a long coherence time, which is the case.

5 SIMULATIONS

The simulated system corresponds to the HL/2 parameters, i.e. it consists of a $N = 64$ point IFFT/FFT, with a CP of $L = 16$ samples, thus $P = 80$ and $Nl = 4$. $Nu = 52$ are useful carriers, while the rest are padding zeros. Monte Carlo simulations have been conducted in a channel 'A' environment ($K = 9$) with a four-element uniform linear array at the receiver ($Q = 4$) with inter-element spacing of 0.5λ. Two randomly distributed interferences disturb permanently the desired user, causing a C/I range from -10 dB to 20 dB. The SNR is 20 dB. The terminals are moving at the maximum speed allowed in HL/2 scenarios, i.e. 3 m/s, while the angular spread is 30°. The selected signal mapping is QPSK and no channel coding is implemented. The number of filter taps is $M = 8$. Note that there are $M_p = 2$ OFDM symbols for the preamble, consisting of ($N_{pr} = 160$) time samples.

The blind method maximizes the temporal output SNIR, as it is shown in Figure 3. There, it can be seen that the SMI-CP method improves the performance of the SMI method as stated in Section 3.2. However, maximizing the temporal SNIR in OFDM does not minimize the BER of the system, see Figure 6. The cause for that is seen in Figures 4 and 5, obtained by a single channel realization. The Q subchannels suffer from deep fading around frequency index 25. Their legend includes the values of the average BER as well as the output temporal SNIR. The blind method maximizes then the SNIR in those subcarriers where the other algorithms have already a good performance in terms of BER (note that

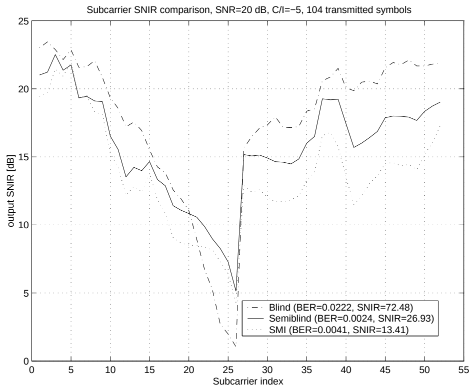

Figure 4: Output SNIR at the useful subcarriers

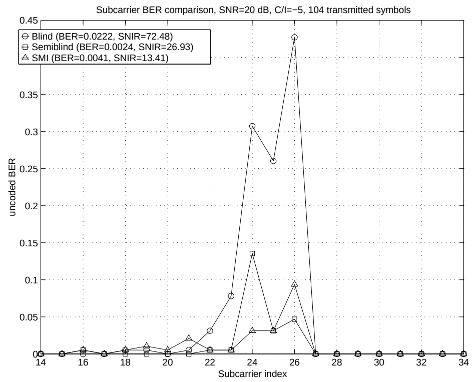

Figure 5: BER per subcarrier (partial)

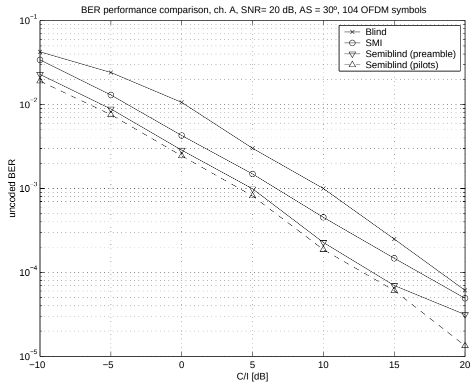

Figure 6: BER performance comparison

Figure 5 is a zoom around the subcarriers suffering from the deep fading). However, in subcarriers with low SNIR, the blind method incurs in a further degradation, setting a null in those subcarriers, which penalizes the average BER. Helped by the preamble, the SMI-CP method, and even SMI, avoid such a big degradation.

In terms of BER (Figure 6), the SMI-CP method outperforms the BER of both the SMI (more than 2 dB) and the blind technique. Its performance is further improved in more than 1 dB by the use of the moving pilots channel estimation technique, which gives also estimates for equalization in the blind method. With 104 transmitted OFDM symbols, this technique yields to better estimations than the ones obtained during the preamble. Note that this BER improvement is subject to the number of transmitted OFDM symbols during the burst. However, in the concrete case of HL/2, due to the long coherence time of the channel, channel estimates could be updated during several frames (5-10).

6 CONCLUSIONS

In this paper, a blind method that maximizes the output temporal SNIR using CP has been presented, which is especially suited for non-stationary scenarios. However, this temporal maximization does not imply the maximization of the SNIR in the frequency domain for all subcarriers, thus degrading the BER performance. A special moving pilots technique is proposed to obtain channel estimates for this blind method. The performance of this approach is then compared to the temporal SMI and the semiblind SMI-CP algorithms. Although the semiblind SMI-CP method increases the computational complexity, it outperforms the SMI (more than 2 dB) and the blind approach (more than 5 dB) in terms of BER. It is also shown that a 1 dB improvement is obtained by the moving pilots structure in semiblind SMI-CP method.

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