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Self-reference Spatial Diversity Processing for Spread Spectrum Communications*

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Abstract **In this paper, the application of three blind (or self-reference) spatial diversity signal processing methods to *Spread Spectrum* (SS) communications is described. These methods do not require any kind of side information beyond knowledge of the signal structure, in contraposition to methods that depend on training sequences.**
Each self-reference method is specifically designed for a particular SS transmission scheme and uses a particular signal's structural information. All three methods, however, are derived under the common goal of finding the optimum beamformer, in the sense of maximum signal to interference-plus-noise ratio, in a blind manner.
The Code Reference Beamformer for Frequency Hopping systems is an approach based on the knowledge of the hopping sequence or code structure of the signal. This method is the best alternative for array signal processing on this particular SS scheme. Two recently proposed beamforming methods, which are based on the knowledge of the redundancy structure of the desired signal, are described under the common framework of efficiently using the inherent diversity – either in frequency or time – of two signalling formats, namely, Frequency Diversity (FD) SS and DS-CDMA. The former uses frequency diversity, while the latter uses time diversity. The diversity approach presented for FDSS seems to be the best choice at the moment. Regarding DS-CDMA, for which many beamforming algorithms have been developed in the literature, the self-reference beamforming is simply an alternative blind method that shows very good performance. These diversity approaches were compared via simulations to the standard Time Reference Beamformer (TRB) and showed a performance similar to it (with 10% of the bits as a training sequence) even without the need of any side information whatsoever.

Keywords **Spatial signal processing, Arrays, Beamforming, Blind methods, Self-reference, Diversity, Redundancy, Frequency hopping, CDMA, FDSS, Spread Spectrum**

1. Introduction

Array signal processing allows interference rejection according to the bearing angles or spatial signatures of the interfering signals by linearly combining the signals received at the antenna elements. The design of the beamformer requires some a priori knowledge about the desired signal in order to discriminate it from the interference. Depending on the character of this information, classical beamforming techniques can be classified into two types. The first one is *Time Reference Beamforming* (TRB), which needs a training sequence that in turn implies a loss of spectral efficiency and channel bandwidth utilization but yields robust methods. The second type is *Spatial Reference Beamforming* (SRB), which requires the knowledge of the direction of arrival (DOA) of the desired signal, the knowledge of the array geometry, and strict array calibration.

An alternative to requiring either a spatial or a time reference is *Code Reference Beamforming*, which can be included in the more general category of *blind beamforming*. These blind (or self-reference) algorithms exhibit good behavior in terms of robustness and spectral/channel efficiency. They are based on the minimization of an objective cost function which is defined based on certain inherent structural properties of the desired signal; hence, no side information, such as a training sequence or the DOA of the desired signal, is needed. This implies that self-reference spatial processing techniques can be applied to existing systems without the need of any modification of the emitter (e.g. the uplink channel of a cellular communication system [1]). An important feature of self-reference beamforming algorithms is that they are not dependent on channel properties or array calibration, which makes them robust methods. It is important to note, on the other hand, that each particular transmission scheme presents a different signal structure and, therefore, requires a specific particular blind method.

Many blind algorithms have been devised exploiting different properties of the signals. They can be divided into two main groups: *deterministic blind beamforming* (using the known structure of the signal) and *statistical blind beamforming* (using statistical properties of the source). All these techniques are also referred to as property restoral algorithms, because they force the signal estimates to exhibit certain properties that the actual signals are known to possess. For man-made signals, such as those encountered in wireless communications, these signal properties are often known with great accuracy, leading to robust algorithms.

Some of the most representative deterministic properties, *Constant Modulus* (CM) [2, 3] and *Finite Alphabet* (FA) (exploited by decision-directed methods [4]), are properties exhibited by many modulation schemes. Other

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signals, known as gated-like signals, have the property of not being present on the channel during a certain time (or at least being present with low power); this allows anticipative processing to estimate the parameters of the scenario free of the desired signal [5–7]. A particular example of this type of signal is Frequency Hopping, where knowledge of the frequency hopping sequence can be used for the anticipative processing [8–10] (see Sect. 2). Another property, the property of diversity, which refers to the existence of signal replicas along the time, frequency, code or space axis (usually to combat fading or jamming), provides the signal with a redundancy that allows optimum beamforming using the replicas as embedded self-references [11–13] (see Sect. 3).

Regarding statistical properties, the most important ones are probability density function fitting and statistical independence of the sources [14], higher order statistics (HOS) [15, 16], and cyclostationary properties [17]. It is important to note that blind statistical methods usually present bad convergence properties, yielding algorithms that are not very reliable.

Blind beamforming techniques lend themselves to block or direct implementation (usually based on a generalized eigenvector decomposition) and to adaptive algorithms (with a reduced computational load). Clearly, adaptive methods are especially suitable for tracking in non-stationary environments.

In this paper, three blind deterministic beamforming methods, based on the knowledge of the structure of the signal, are described for *Spread Spectrum* (SS) communication systems. Each one is specifically designed for a particular SS signalling scheme, but derived from a common framework of diversity aimed at the maximization of the Signal to Interference-plus-Noise Ratio (SINR). The performance of each method is verified via simulations. In Sect. 2, we describe the *Code Reference Beamformer* based on anticipative processing for slow *Frequency Hopping* SS (FHSS). Sect. 3 introduces beamforming techniques based on a general framework of diversity properties. Then, Sects. 3.1 and 3.2 present specific self-reference algorithms based on frequency diversity, for both fast FHSS and *Frequency Diversity* SS (FDSS), and based on time diversity, for *Direct Sequence* SS (DSSS), respectively. Finally, in Sect. 4, the conclusions are presented.

2. Code reference beamforming for slow FHSS

The addition of spatial diversity processing with an array to *Frequency Hopping Spread Spectrum* (FHSS) techniques provides a greater capacity for interference rejection. Existing beamformers for the reception of FH modulated signals are based on classical techniques of temporal or spatial reference beamforming. First, Acar and Compton [18] studied the adverse effects of FH in a time reference beamformer based on the LMS algorithm. Next, Bakhru and Torrieri [19–21] proposed the Maximin algorithm, a specific method for adaptive arrays using FH signals (assuming knowledge of the FH sequence) based on the spectral characteristics of the received signals. Finally, Eken [22] devised a modified sidelobe canceller for FH signals that needed a priori knowledge of the DOA of the desired signal. The main drawback of the previous approaches is the drop in SINR at hop instants. The reason for this behavior, as was pointed out in [18, 20], is that changes in signal frequency due to FH modulation are seen by the algorithm as changes in the DOA's, resulting in discontinuities of the adaptive algorithms' performance.

The anticipative processing algorithm proposed in [8–10] and denoted Code Reference Beamformer takes advantage of the knowledge of the FH sequence at the receiver, as in [20], and requires neither temporal nor spatial references. The method relies on the possibility of the measurement and estimation of the interference-plus-noise covariance matrix. This property has already been used by several authors in non FH contexts [5–7].

The proposed system, depicted in Fig. 1, is composed of two parallel processors; both processors begin by dehopping the received signal. The first one, called the anticipative processor, is used to obtain the interference-plus-noise covariance matrix, $\mathbf{R}_{in}$, before the desired signal is present. The second one, the on-line processor, performs the actual beamforming.

The beamforming part of the on-line processor is made up of two different stages, corresponding to the two terms of the optimum beamvector $\mathbf{w}_{opt} = \kappa \mathbf{R}_{in}^{-1} \mathbf{a}_d$, where κ is a scalar factor, and $\mathbf{a}_d$ is the steering vector of the desired signal. The first stage preprocesses the dehopped received snapshot, $\mathbf{x}_{ol}(t)$, using the inverse of the interference-plus-noise covariance matrix to obtain $\mathbf{x}(t) = \mathbf{R}_{in}^{-1} \mathbf{x}_{ol}(t)$. The second stage is devoted to properly weighting the preprocessed signal vector $\mathbf{x}(t)$ with the desired steering vector $\mathbf{a}_d$. Since $\mathbf{a}_d$ is assumed to be unknown, a blind es-

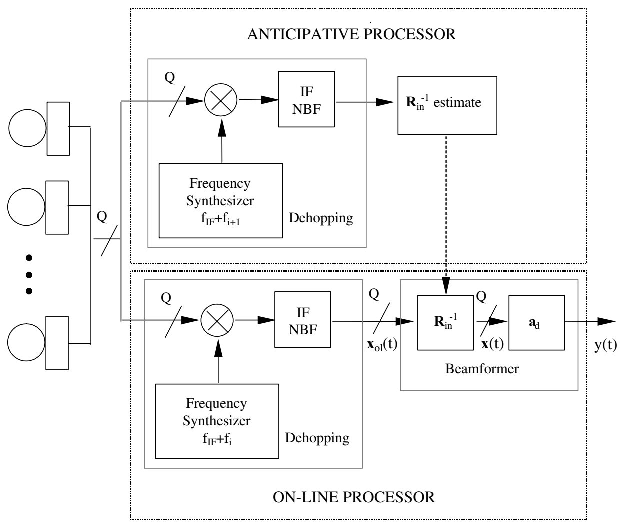

Fig. 1. Code Reference Beamformer.

timation of this steering vector is adaptively obtained by maximizing the output SINR [10]:

$$\mathbf{a}_d = \arg\max_{\mathbf{c}\in\mathbb{C}^Q} \left\{ \text{SINR} = \frac{\mathbf{c}^H\mathbf{R}_x\mathbf{c}}{\mathbf{c}^H\mathbf{R}_{in}^{-1}\mathbf{c}} - 1 \right\}, \tag{1}$$

where $\mathbf{R}_x = \mathbf{R}_{in}^{-1}\mathbf{R}_{x_{ol}}\mathbf{R}_{in}^{-1}$ is the correlation matrix of the preprocessed signal vector $\mathbf{x}(t)$, and Q is the number of antenna elements.

The solution to this problem statement is given by the generalized eigenvector of the matrix pencil $\left(\mathbf{R}_x, \mathbf{R}_{in}^{-1}\right)$ corresponding to the maximum generalized eigenvalue:

$$\mathbf{R}_x\mathbf{c} = \lambda_{\max}\mathbf{R}_{in}^{-1}\mathbf{c}\,, \tag{2}$$

where $(\lambda_{\max} - 1)$ is the maximum SINR. An adaptive implementation of the computationally expensive eigendecomposition can be found in [8, 10].

In mobile communications systems, due to the presence of fading on the received interferences (and the possible appearance/disappearance of interfering signals), the matrix $\mathbf{R}_{in}^{-1}$ estimated by the anticipative processor should be continuously modified in order to track the non-stationary scenario [10], becoming a discrete-time dependent $\mathbf{R}_{in}^{-1}(n)$. The inverse correlation matrix estimated by the anticipative processor is adequate only as an initialization of the system. Once it has been transferred to the on-line processor, and after few iterations to achieve convergence of the second stage, $\mathbf{R}_{in}^{-1}(n)$ is iteratively updated (see Fig. 2) using the following recursion:

$$\mathbf{R}_{in}^{-1}(n+1) = \gamma\mathbf{R}_{in}^{-1}(n) + (1-\gamma)\mathbf{m}(n+1)\mathbf{m}^H(n+1)\,. \tag{3}$$

Here $\mathbf{m}(n)$ is an estimate of the undesired component of the incoming snapshot pre-processed by the first stage of the beamformer [10]:

$$\mathbf{m}(n) = \mathbf{x}(n) - g(n)\tilde{s}(n)\mathbf{R}_{in}^{-1}\mathbf{w}\,, \tag{4}$$

where $\tilde{s}(n)$ is a noisefree regenerated signal that can be extracted from the beamformer output $y(n)$ and $g(n)$ is an adaptive coefficient continuously adjusted to minimize the

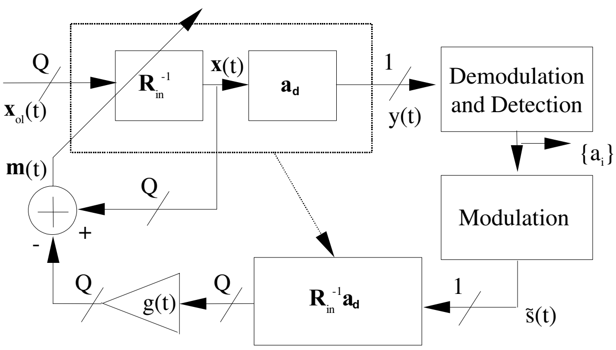

Fig. 2. First stage adaptation for frequency-non selective channels.

log-likelihood cost function $\mathbf{m}^H(n)\mathbf{R}_{in}^{-1}\mathbf{m}(n)$ (see [10] for details).

If the frequency synthesizer of the anticipative processor is not perfectly synchronized with the received signal, the estimated interference-plus-noise covariance matrix $\mathbf{R}_{in}$ will present a residual contribution from the desired signal. This situation, however, does not prevent the second stage from converging to the solution that maximizes the SINR at the array output. In [8] it was shown that the convergence rate with leakage of the desired signal into the covariance matrix can be maintained at the expense of an increment in the algorithm misadjustment. This is a key feature of the procedure's robustness.

In principle, the initial weighting vector used in the adaptive implementation is randomly chosen or fixed to present an isotropic gain pattern. Nevertheless, this initialization causes the problem, already observed in the literature, of SINR drops at frequency hop instants. This arbitrary initialization could be substituted by a good initial estimate of the steering vector of the desired signal with the aim of speeding up the convergence of the adaptive algorithm, hence reducing the SINR drops. This can be done by the use of focussing techniques, which convert the steering phases estimated in the previous hop to the actual one by simply multiplying them by the frequency ratio f_{i+1}/f_i. This allows a prediction of the optimum weighting vector for the next hop. Note that this procedure does not require previous knowledge either of the location of the sensors or of the desired direction of arrival. It is generally accepted, however, that in FH systems the frequency spacing is always set higher than the coherence bandwidth of the mobile channel. Therefore, two consecutive channels will appear as statistically independent, and the same will happen to the generalized steering vectors. This fact, along with the lack of linear dependence on frequency of the equivalent steering vector elements, might be expected to contribute to a degradation of the results achieved with the frequency focussing process. Nevertheless, simulation results show that it is still adequate to focus the second stage at the hop instants as a good initialization, since there is always enough residual spatial information that can be taken into account.

In order to show the effectiveness of the method, simulations were performed with a *uniform linear array* (ULA) of 4 sensors with half-wavelength separation with respect to the center frequency (900 MHz) of the hopping band. The following scenario was used: a Gaussian Minimum Shift Keying (GMSK) signal of interest (DOA of 20 degrees and $E_b/N_o = 15$ dB) spread over a 50% relative bandwidth (assuming slow FH) and two interfering users, also GMSK signals, with bearing angles of -30 and 40 degrees ($E_b/N_o = 15$ dB). Each user's flat-fading channel was generated with a Laplacian Power Angular Spectrum ray model with a power Azimuth Spread (AS) of 8 degrees [23]. The speed of the users was set to 120 km/h.

The instantaneous SINR evolution with and without adaptation of the first stage is depicted in Fig. 3 along with the optimum values (in the sense of maximum SINR using the true covariance matrices). It can be noted that

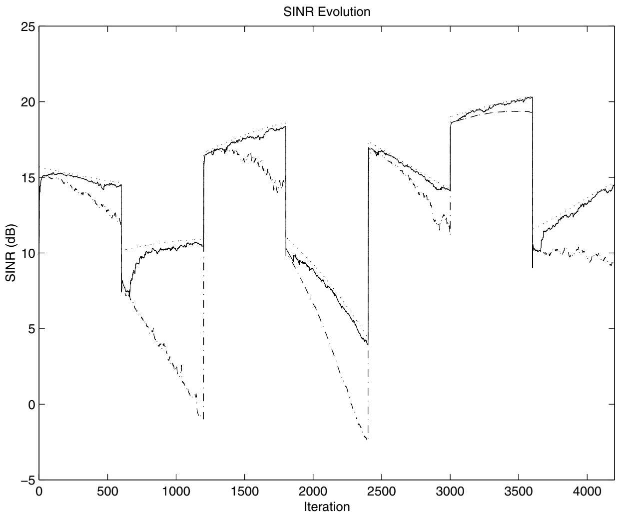

Fig. 3. Instantaneous SINR evolution of the Code Reference Beamformer for FH system using frequency focussing. Scenario: a desired signal with a DOA of 20° and two interfering users with DOAs of −30° and 40° (all users GMSK signals with $E_b/N_o = 15$ dB) received with an array of 4 sensors through a frequency non-selective mobile channel (v = 120 km/h). Dotted line: optimum value. Dash-dotted line: without adaptation of the first stage. Solid line: with adaptation of the first stage.

the adaptation of the first stage confers the ability to track the channel variations. A frequency focussing process has been used in both algorithms to reduce the SINR drops at hop instants.

3. Diversity-based beamforming for SS

Diversity-based beamforming combines the use of spatial diversity at the receiver with other types of diversity inherent to the signal of interest or to the channel.

Diversity is a concept closely related to multiplexing. It refers to the existence of replicas of the same signal, providing the signal structure with an inherent redundancy. Both terms can be lumped together into the more general idea of combining a number of signals – either the same signal or different signals from different users – in such a way that it is always possible to separate them out again. The combination of these sets of signals or diversity branches can be done in any of the support axes, namely: time, frequency, code and space.[1]

In the context of multiplexing, these four axes give rise to four types of multiplexing techniques:

i) *Time Division Multiple Access* (TDMA), the use of non-overlapping time slots for each user,
ii) *Frequency Division Multiple Access* (FDMA), the use of non-overlapping frequency bands for each user,
iii) *Code Division Multiple Access* (CDMA), the use of a different code (linearly independent and quasi-orthogonal) for each user, overlapping over the time and frequency axes, and
iv) *Space Division Multiple Access* (SDMA), where each user presents a different spatial signature.

In the same way, four kinds of diversity can be identified:

i) *Time Diversity* (TD), simple repetition of the signal in different non-overlapping time slots, e.g., a single user in DS-CDMA or the spontaneous phenomenon of multipath in frequency selective channels[2] (efficiently utilized by the RAKE receiver [24]),
ii) *Frequency Diversity* (FD), transmission of the same signal in different non-overlapping frequency bands, e.g., FDSS [25] and fast FH,
iii) *Code Diversity* (CD), transmitting the same signal with different codes[3] or spreading signatures (overlapping over the time and frequency axes) [26], and
iv) *Space Diversity* (SD), caused by the multipath phenomenon since each ray comes from a different direction of arrival (this diversity can only be exploited at the receiver by the use of antenna arrays).

Clearly, when diversity is used for transmission of a signal, the available power has to be distributed over the replicas.

Self-replicas can be seen as embedded self-reference signals that give rise to the important property of self-prediction, i.e., some replicas of the signal can be predicted using other replicas [11]. This property has also been called the self-coherence property, giving name to the Self-COherence REstoral algorithm (SCORE) [17], which was applied to cyclostationary signals.

Let us assume that there are two available signals, $u(t)$ and $v(t)$, with a correlated desired signal component, $d(t)$, and uncorrelated noise terms, $n_u(t)$ and $n_v(t)$:

$$u(t) = \alpha d(t) + n_u(t) \ , \tag{5a}$$

$$v(t) = d(t) + n_v(t) \ , \tag{5b}$$

where α is the normalized cross-correlation coefficient between $u(t)$ and $v(t)$. These two signals exhibit the key feature property of *exact self-prediction.* This means that it is possible to predict the desired component of one signal using the other one. In other words, we can define a prediction error that will be composed only of uncorrelated interference and noise. Therefore, this prediction error can be used as the objective function to minimize:

$$e(t) = u(t) - \alpha v(t) \ , \tag{6}$$

[1] Other support axes could also be considered, such as the polarization axis.

[2] Note that this diversity resulting from a repetition of time-delayed replicas of a signal is also commonly referred to as frequency diversity or path diversity in the literature. The former because in the frequency domain it presents a different fading for each frequency bin and the latter for obvious reasons.

[3] It is important to distinguish the term Code Diversity (replicas of a signal using different spreading codes) from the concept of Code Reference (use of the code structure of the signal as a reference).

where α is, in general, a complex constant. In the following, $\alpha = 1$ is assumed (case of equal-power and phase-aligned replicas) for simplicity of presentation.

3.1 Frequency diversity beamforming for FDSS and fast FHSS

The *Frequency Diversity Spread Spectrum* (FDSS) technique was proposed in 1996 [25] as a system to mitigate the problems of the common spread spectrum techniques, namely, vulnerability of Direct Sequence to bandlimited partial-time (burst) jamming and of Frequency Hopping to partial-band interference. FDSS can be seen as a simultaneous transmission of a signal over different non-overlapping frequency bands[4] (see Fig. 4), providing the global signal structure with the previously described property of exact prediction. A detailed comparison between FDSS and DS-CDMA can be found in [27].

A classical TRB can be used to perform a beamforming at the receiver. Nevertheless, since the property of exact prediction is clearly present in this particular SS scheme, we can use a self-reference approach to perform blind array processing without the need of side information. In this signalling scheme, there are N available frequency bands – diversity branches – that can be used. The approach taken in [11] consists on using the exact prediction between just two diversity branches. Note that the interference is assumed uncorrelated between the selected bands, which can be guaranteed if different chip spreading is used on each band (see Fig. 4). The potential benefit of using the N diversity branches jointly is not taken into account in the present paper (c.f. [28]). The underlying method is equivalent to the cross-SCORE algorithm [17], which yields exactly the same solution.

Note that the beamforming method described hereinafter is completely applicable to fast Frequency Hopping, which basically consists of a carrier hopping faster than the symbol period. This can be seen as replicas of a signal transmitted at different frequency bins at consecutive time slots.

In the case of frequency diversity, since the snapshots of the two diversity branches correspond to different carrier frequencies, which are assumed to undergo independent fading, it is clear that different beamvectors have to be used. Once the beamvectors for these two branches have been designed, their output signal can be used as a reference for the rest of the bands using a time reference approach.

The algorithm is based on the minimization of the prediction error – difference between the beamformer outputs – with a constraint to avoid the trivial solution:

$$\begin{aligned} &\min \ E\left\{|u(t)-v(t)|^2\right\} \\ &\text{subject to } \operatorname{Re}\left\{E\left\{u(t)v^*(t)\right\}\right\} = \phi_c \ , \end{aligned} \tag{7}$$

[4] The signal transmitted over each band is, in general, a (n,k) coded version of the original signal. The trivial repetition channel code (n,1) is considered here.

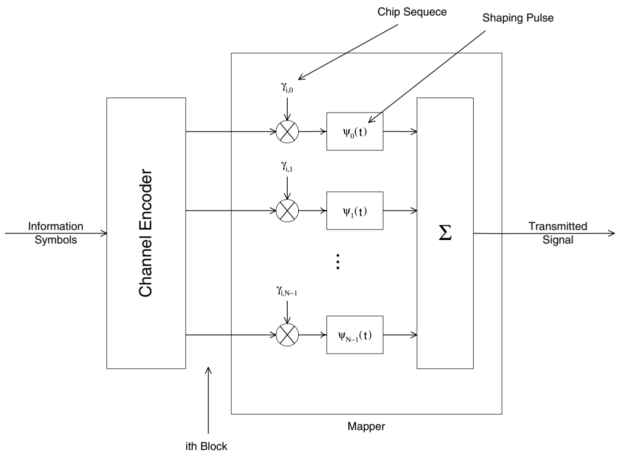

Fig. 4. Scheme of FDSS signal generation.

where $u(t)$ and $v(t)$ are the beamforming output signals corresponding to snapshots of two different bands: $u(t) = \mathbf{w}_u^H \mathbf{x}_u(t)$ and $v(t) = \mathbf{w}_v^H \mathbf{x}_v(t)$.

Equation (7) can also be seen as the maximization of the SINR:

$$\text{SINR} = \frac{\operatorname{Re}\left\{E\left\{u(t)v^*(t)\right\}\right\}}{E\left\{|u(t)-v(t)|^2\right\}} \ , \tag{8}$$

which is equivalent to the maximization of the following Rayleigh quotient:

$$\frac{2\operatorname{Re}\left\{E\left\{u(t)v^*(t)\right\}\right\}}{E\left\{|u(t)|^2+|v(t)|^2\right\}} = \frac{\mathbf{w}_u^H \mathbf{R}_{uv}\mathbf{w}_v + \mathbf{w}_v^H \mathbf{R}_{vu}\mathbf{w}_u}{\mathbf{w}_u^H \mathbf{R}_{uu}\mathbf{w}_u + \mathbf{w}_v^H \mathbf{R}_{vv}\mathbf{w}_v} \ , \tag{9}$$

where $\mathbf{R}_{ij} = E\left\{\mathbf{x}_i \mathbf{x}_j^H\right\}$. The eigenvectors that maximize the quotient of (9) are given by [11]:

$$\mathbf{R}_{uv}\mathbf{w}_v = \lambda_{\max}\mathbf{R}_{uu}\mathbf{w}_u \ , \tag{10a}$$

$$\mathbf{R}_{vu}\mathbf{w}_u = \lambda_{\max}\mathbf{R}_{vv}\mathbf{w}_v \ , \tag{10b}$$

and coincide with the solution given by the cross-SCORE algorithm [17].

Note that in the covariance matrices of the denominator of (9), $\mathbf{R}_{uu}$ and $\mathbf{R}_{vv}$, all the signals present in the scenario appear; whereas the numerator only contains the desired one, because it is the only correlated signal between both branches.

An adaptive version of the algorithm, called cross-coupled LMS, can be found in [11, 29]. Note that an optimal joint symbol detection, using the output signal of the N bands, can be performed taking into account the effect of the beamforming [11].

To test the described beamforming method, Monte-Carlo simulations were carried out with a uniformly spaced linear array of 7 sensors, using a half wavelength separation. The scenario was composed of a BPSK signal of interest (spread over a relative bandwidth of 50% with respect to $f_c = 900$ MHz) with a DOA of 15 degrees and 4 interfering BPSK signals present in all the bandwidth

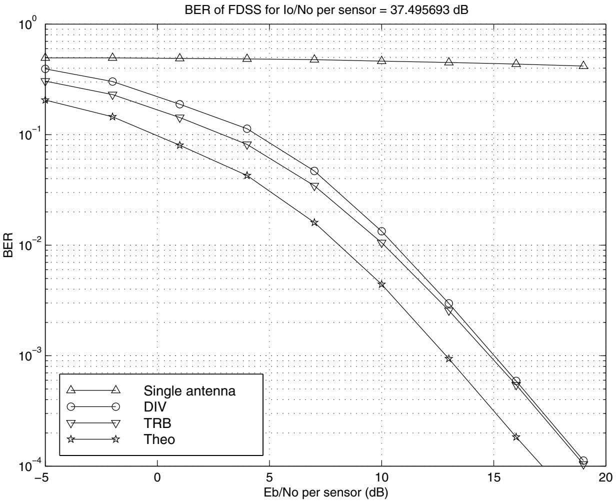

Fig. 5. BER for a receiver with an array of 7 elements of the classical TRB, the proposed diversity method and the minimum theoretical BER (for the array and the single sensor case) using FDSS. Scenario: desired signal with a DOA of 15° and 4 interfering signals with DOA's of 0°, −30°, 50° and 80° with an instantaneous I_0/N_0 per sensor of 37.5 dB.

(uncorrelated among the two selected bands) impinging from angles of 0, −30, 50 and 80 degrees (E_b/N_0 of 30 dB, 20 dB, 40 dB and 20 dB) respectively, with a total $I_0/N_0 = 37.5$, with I_0 the global interference power per sensor. To model the angular spread of the steering vectors, a Laplacian Power Angular Spectrum with a power azimuth spread of 8 degrees was used according to [23]. The temporal dispersion of the channel was generated with a Pedestrian model as specified by ETSI [30] assuming a speed of 3 km/h.

In Fig. 5, the *Bit Error Rate* (BER) of the proposed method can be seen compared to that of a classical TRB (10% as a training sequence) and to the minimum theoretical achievable BER – either using an array or a single antenna. Obviously, the TRB performance depends on the percentage of the bits used as a training sequence. From the simulation results, it can be easily seen that the blind method performs similarly to a TRB with 10% training sequence even though it needs no side information and therefore presents a higher spectral and bandwidth utilization.

3.2 Time diversity beamforming for DS-CDMA

The property of exact prediction can also be used in DS-CDMA communication systems by realizing how, within any symbol, each chip can be predicted from any other – assuming knowledge of the spreading code.

The straightforward parallelism of a signal replicated along the frequency axis would be a symbol repeated along the time axis. Actually, a single symbol of duration T_s can be interpreted as an implicit repetition of subsymbols – chips – of shorter period T_c (see Fig. 6a–b).

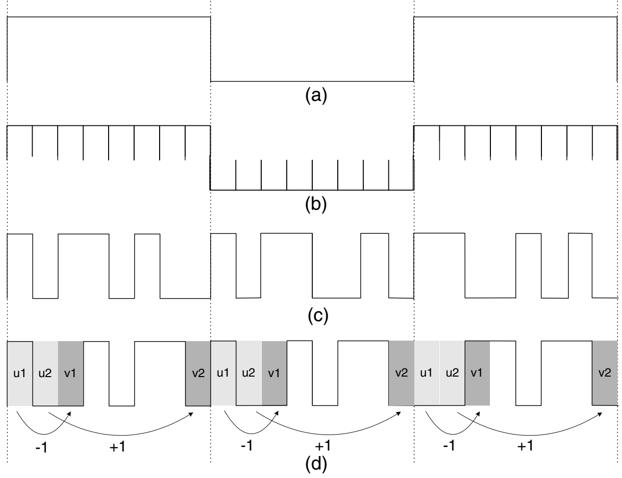

Fig. 6. Example of Time Diversity in DS-CDMA: (a) original bit sequence, (b) bit sequence interpreted as a diversity repetition code, (c) spread bit sequence using a standard PN, and (d) spread bit sequence using a fixed redundancy structure (redundant PN).

Nevertheless, in order to be able to have multiple users overlapping in time and frequency, they have to possess a redundancy structure more sophisticated than a simple repetition scheme. It is compulsory, therefore, to provide the replicas with a specific pattern for each user, i.e., a temporal signature or chip spreading code (see Fig. 6c). This way, it will be feasible to overlay users, giving rise to the well-known DS-CDMA technique.

The direct application of the previously described algorithm in the frequency diversity context consists of taking a couple of chips within one symbol – two temporal branches – and use then the exact prediction property. It is noteworthy that a unique beamvector $\mathbf{w}$ is needed in this case, because narrowband is assumed and the signals are considered to be stationary along a symbol period. It is possible, however, to perform a much better estimation of the desired signal covariance matrix $\mathbf{R}_d$ by exploiting the prediction property among the multiple replicas. Assuming that there are N_c chips per symbol, $\left(N_c^2 - N_c\right)/2$ prediction error or cross-correlation terms can be used instead of just two [12], although not all of them will be linearly independent, but rather just $(N_c - 1)$ [31].

To describe the algorithm, we first define a block snapshot by grouping the snapshots columnwise within a symbol period $\mathbf{X}(t) = [\mathbf{x}_1(t), \mathbf{x}_2(t), \cdots, \mathbf{x}_{Nc}(t)] \in \mathbb{C}^{\mathrm{QxN}_c}$, where Q is the number of sensors, N_c the number of chips per symbol (spreading factor), and $\mathbf{x}_k$ the snapshot corresponding to branch k, defined as:

$$\mathbf{x}_k(t) = \sqrt{p_d} s_{d,k}(t)\mathbf{a}_d + \sum_{i=1}^{N_I} \sqrt{p_i} s_{i,k}(t)\mathbf{a}_i + \mathbf{n}_k(t) , \quad 1 \le k \le N_c , \tag{11}$$

where N_I is the number of interfering signals, $s_{d,k}(t)$ ($s_{i,k}(t)$) refers to desired (ith interfering) temporal signal with power p_d (p_i) in branch k, $\mathbf{a}_d$ ($\mathbf{a}_i$) denotes the desired (ith interfering) steering vector, and $\mathbf{n}_k(t)$ represents the noise component of the snapshot in branch k.

The first step consists of *polarizing* the block snapshot so that the desired signal is fully correlated whereas neither the interferences nor the noise are (assuming long PN sequences, i.e., time-varying codes). This is achieved by multiplying the block snapshot by the synchronized spreading sequence of the desired user (PN synchronization is assumed), $\mathbf{c}(t) \in \mathbb{C}^{N_c x1}$:

$$\mathbf{X}^{pol}(t) = \mathbf{X}(t) \cdot diag(\mathbf{c}(t)) \ , \tag{12}$$

where $diag(\mathbf{v})$ is a diagonal square matrix containing vector $\mathbf{v}$ along its main diagonal.

In the *polarized* snapshots, the desired signal is fully correlated: $s_{d,k}^{pol}(t) = s_d(t)$, whereas the interfering users still remain uncorrelated: $E\{s_{i,k}^{pol}(t)s_{j,l}^{pol,*}(t)\} = \delta_{ij}\delta_{kl}$, which can be guaranteed due to the time-varying nature of the spreading codes. Since the desired signal is the only one correlated among the time branches, $\mathbf{R}_d$ and $\mathbf{R}_x$(received signal covariance matrix) can be estimated as [12]:

$$\mathbf{R}_d = \frac{1}{N_c(N_c-1)} \sum_{l=1}^{N_c} \sum_{\substack{m=1 \\ m \neq l}}^{N_c} E\left\{\mathbf{x}_l^{pol}(t)\mathbf{x}_m^{pol,H}(t)\right\} \tag{13}$$

$$\mathbf{R}_x = \frac{1}{N_c} \sum_{k=1}^{N_c} E\left\{\mathbf{x}_k^{pol}(t)\mathbf{x}_k^{pol,H}(t)\right\} \ , \tag{14}$$

which can be compactly expressed as

$$\mathbf{R}_d = \frac{1}{N_c(N_c-1)} E\left\{\mathbf{X}^{pol}(t)\tilde{\mathbf{I}}\mathbf{X}^{pol,H}(t)\right\} \ , \tag{15}$$

$$\mathbf{R}_x = \frac{1}{N_c} E\left\{\mathbf{X}^{pol}(t)\mathbf{X}^{pol,H}(t)\right\} \ , \tag{16}$$

where the matrix $\tilde{\mathbf{I}} \in \mathbb{C}^{N_c x N_c}$ is defined as $\tilde{\mathbf{I}} = \mathbf{1} - \mathbf{I}$, being $\mathbf{1}$ the all-one matrix and $\mathbf{I}$ the identity matrix. The final formulas for the estimation of the desired and total covariance matrices (15) and (16), which have been derived from a viewpoint of chip-level cross-correlation properties using the key feature of exact prediction, are equivalent to those found by Suard et al. in [7] from a matched filtering perspective.

Nevertheless, this method based on exact prediction properties requires time signature or PN synchronization. In some adverse situations, spatial filtering is needed before complete synchronization is achieved, so that beamforming helps the synchronization stage. For these cases, it is necessary to devise an alternative diversity scheme able to work without global PN synchronization.

The alternative scheme proposed in [13] works by introducing an explicit fixed redundancy structure over the time-varying spreading codes – long PN sequences – contained in each symbol. Therefore, only symbol timing is needed (a requirement far more relaxed than the global long PN sequence synchronization needed in the previous method). The underlying idea consists of generating a modified PN sequence so that it presents a fixed structure within each symbol. This can be done by the use of a fixed redundancy matrix $\mathbf{G}$ applied on a PN-frame basis:

$$\mathbf{rpn}_q = \mathbf{G}^T \mathbf{pn}_q \ , \tag{17}$$

where the original PN frame, $\mathbf{pn}_q \in \mathbb{C}^{\mathrm{Nx1}}$, is transformed into the redundant and longer PN frame, $\mathbf{rpn}_q \in \mathbb{C}^{\mathrm{Mx1}}$, using the redundancy matrix, $\mathbf{G}^T \in \mathbb{C}^{\mathrm{MxN}}$, with $\mathrm{R} \triangleq (\mathrm{M}-\mathrm{N})$ the number of redundant chips. The redundancy structure considered herein is simply the repetition of certain chips with a possible change of sign so that the finite alphabet is kept (as can be seen from Fig. 6d where the matrices used have been defined according to Fig. 7). The redundancy matrix has to be properly designed so that the desired auto- and cross-correlation properties of the original PN sequences are not destroyed (see [31] for details). Furthermore, if different users have to coexist, the redundancy structures for all of them have to be jointly designed so that each presents a zero cross-correlation with respect to any other user's structure. To this end, extensive search methods have to be used and, therefore, small values of R are easier to handle (c.f. [31]).

The receiver will make use of a parity check matrix, $\mathbf{H} \in \mathbb{C}^{\mathrm{MxR}}$ to define an error $\mathbf{e}_q \in \mathbb{C}^{\mathrm{Rx1}}$:

$$\mathbf{e}_q = \mathbf{H}^T \mathbf{y}_q = \left(\mathbf{H}_u^T - \mathbf{H}_v^T\right) \mathbf{y}_q = \mathbf{u}_q - \mathbf{v}_q \ , \tag{18}$$

where $\mathbf{y}_q \in \mathbb{C}^{\mathrm{Mx1}}$ contains the desired modulated RPN signal plus noise and interferences (defined on a frame basis). The matrices $\mathbf{H}_u$ and $\mathbf{H}_v$ can be seen as matrices that extract two different sets of diversity elements, $\mathbf{u}$ and $\mathbf{v}$, from the redundant vector, $\mathbf{y}$ (see Fig. 7). Note that the previously described method using the implicit redundancy of spreading codes is a particular case of this general scheme with the specific redundancy matrix $\mathbf{G} = \mathbf{I}$ (with no explicit redundancy, i.e., $\mathrm{R} = 0$ and $\mathrm{M} = \mathrm{N}$).

In the same way as for the frequency diversity scheme, the beamforming algorithm for time diversity is based on the maximization of the SINR:

$$\mathrm{SINR} = \frac{\mathrm{Re}\left\{E\left\{\mathbf{u}_q^T \mathbf{v}_q^*\right\}\right\}}{E\left\{\left\|\mathbf{e}_q\right\|^2\right\}} \ , \tag{19}$$

$$\mathbf{G}^\mathrm{T} = \begin{bmatrix} 1 & 0 & 0 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 & 0 & 0 \\ -1 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 0 & 0 & 1 \\ 0 & 1 & 0 & 0 & 0 & 0 \end{bmatrix} \qquad \mathbf{H}_\mathrm{u}^\mathrm{T} = \begin{bmatrix} 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 & 0 & 0 & 0 & 0 \end{bmatrix} \qquad \mathbf{H}_\mathrm{v}^\mathrm{T} = \begin{bmatrix} 0 & 0 & -1 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 \end{bmatrix} \qquad \mathbf{H}^\mathrm{T}\mathbf{G}^\mathrm{T} = \mathbf{0}$$

Fig. 7. Example of redundancy and check-parity matrices $\mathbf{G}$ and $\mathbf{H}$.

which is equivalent to the maximization of the following Rayleigh quotient:

$$\frac{2\mathrm{Re}\left\{E\left\{\mathbf{u}_q^T\mathbf{v}_q^*\right\}\right\}}{E\left\{\left\|\mathbf{u}_q\right\|^2+\left\|\mathbf{v}_q\right\|^2\right\}}=\frac{\mathbf{w}^H\left(\mathbf{R}_{uv}+\mathbf{R}_{vu}\right)\mathbf{w}}{\mathbf{w}^H\left(\mathbf{R}_{uu}+\mathbf{R}_{vv}\right)\mathbf{w}}\,, \tag{20}$$

where $\mathbf{R}_{ij}=E\left\{\mathbf{X}_q\mathbf{H}_i\mathbf{H}_j^H\mathbf{X}_q^H\right\}$ and $\mathbf{X}_q$ is a block snapshot matrix containing the snapshots of the whole frame columnwise.

Note that only the desired signal is present in the numerator of (20) because it is the only correlated signal in both branches (recall that, for this to be true, the $\mathbf{G}$'s of all the users must be jointly designed to present a zero cross-correlation with respect to any other user structure [31]). The denominator of (20) contains all the signals present in the scenario. Thus, the desired and global covariance matrices can be defined as $\mathbf{R}_d \propto (\mathbf{R}_{uv}+\mathbf{R}_{vu})$ and $\mathbf{R}_x \propto (\mathbf{R}_{uu}+\mathbf{R}_{vv})$.

Once the matrices $\mathbf{R}_d$ and $\mathbf{R}_x$ have been estimated, either using the implicit or explicit redundancy structure, the beamvector that maximizes the SINR is given by the generalized eigenvector $\mathbf{w}$ corresponding to the maximum generalized eigenvalue:

$$\mathbf{R}_d\mathbf{w}=\lambda_{\max}\mathbf{R}_x\mathbf{w}\,, \tag{21}$$

yielding $\mathrm{SINR}_{\max}=1/(\lambda_{\max}^{-1}-1)$. Adaptive implementations of the computationally expensive calculation of the generalized eigenvector can be found in [10, 13].

To evaluate and compare the methods presented using the implicit and explicit redundancy, Monte-Carlo simulations were performed with a uniformly spaced linear array of 7 sensors, using a separation of half wavelength. A simple and suboptimum single-user detector consisting of a matched filter was used after beamforming. The scenario involved was a BPSK signal of interest with a DOA of 15 degrees and 4 interfering BPSK signals impinging from angles of 0, −30, 50 and 80 degrees (E_b/N_o of 30 dB, 20 dB, 40 dB and 20 dB) respectively, with a total $I_0/N_0=25.6$, with I_0 the global chip-level interference power per sensor. Gold sequences of length $N_c=1023$ and $\mathrm{SF}=31$ were utilized (for the RPN case, $\mathrm{R}=2$ was used). To model the angular spread of the steering vectors, a Laplacian Power Angular Spectrum with a power azimuth spread of 8 degrees was used according to [23]. The temporal dispersion of the channel was generated with a Pedestrian model as specified by ETSI [30] with a speed of 3 km/h.

In Fig. 8, the performance of the proposed methods is shown in terms of BER. The uncoded BERs for both the implicit (PN-based) and explicit (RPN-based) time-redundancy methods, for the classical TRB (10% as a training sequence) aided with a decision directed (DD) approach during the data transmission, and the minimum theoretical BER – for both the array and single antenna case – are depicted. It can be seen how the PN-based method outperforms the TRB and is just about 1 dB worse than the theoretical minimum BER. The RPN-

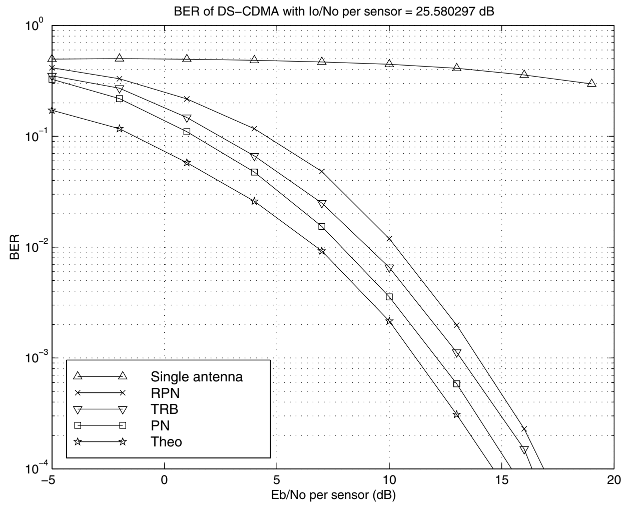

Fig. 8. Uncoded BER of the different methods (TRB+DD using 10% as training sequence, PN-based, RPN-based) along with the theoretical minimum BER – for the array and the single antenna case. The scenario was involved of a desired signal with DOA of 15° and 4 interfering signals with DOA's of 0°, −30°, 50° and 80° (E_b/N_o of 30 dB, 20 dB, 40 dB and 20 dB) respectively. $N_c=1023$ and $\mathrm{SF}=31$.

based method is clearly inferior to those two methods. The difference of performance between the PN and RPN based methods is due to the number of cross-correlation terms per symbol used ($\mathrm{R}=2$ for RPN vs. 465 (30 independent) for PN); it has to be remembered, however, that the RPN approach allows beamforming before global PN synchronization. It can be concluded from the results that the choice of the algorithms has to be based on a trade-off between performance and the amount of side information to be transmitted – with the consequent reduction of spectral efficiency and bandwidth utilization.

4. Conclusions

In this paper, three blind spatial diversity array signal processing methods for Spread Spectrum (SS) techniques have been described. Each method was specifically designed for a particular SS signalling scheme.

The *Code Reference Beamformer* using anticipative processing was derived for slow FHSS to maximize the SINR. Its performance was shown via simulations in a non-stationary environment. It was able to track the scenario in an adaptive fashion, with a performance, in terms of SINR, quite close to the optimum theoretical value. This self-reference beamforming method appears to be the best alternative for array signal processing applied to slow FHSS.

Two diversity-based beamforming algorithms were described for the frequency and time axes and were derived under a common framework of redundancy properties. They were shown to yield maximum SINR and were compared via simulations to the classical TRB. These self-reference beamformers, despite the lack of side information, achieved a performance quite similar to TRB. For FDSS, the described self-reference method seems to be the best choice to perform beamforming at the receiver. For DS-CDMA, however, there is a wide range of available algorithms in the literature and the presented self-reference method is simply a blind alternative; a choice of algorithm would depend on a trade-off between performance and amount of side information.

The beamforming methods presented herein are based on structural inherent properties of the signals and do not need any external reference or side information; therefore, they present a good spectral efficiency and channel bandwidth utilization. They do not make any assumption regarding the steering vectors or spatial signatures of the signals and, consequently, they are robust and do not require array calibration.

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