A constant power algorithm for partial-band interference rejection in frequency-hopping communication systems*

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ABSTRACT

We consider the rejection of partial-band interferences in frequency-hopping communication systems using an array of sensors. A simple yet effective method is proposed based on a constant power algorithm. The principle behind this method is that partial-band interferences contribute to power variations in the received signal, hence the idea to constrain the output power of the array to be constant in order to reject interferences. It is shown that the algorithm converges in a reasonably low number of hops. Additionally, it achieves a nearly optimal signal to interference and noise ratio without requiring any information about the location of the desired user.

1 Introduction

In satellite communication systems, frequency-hopping is recognized as an efficient means for secure communications [1]. In contrast to direct-sequence code-division multiple access its requirements in terms of synchronization are less stringent. However, particularly for military applications, powerful interferers aiming at disturbing the communications are major cause of performance degradations. This is specially the case with a partial band jammer, i.e. an interference which occupies a fraction of the total bandwidth. Since the signal of interest (SOI) is hopping, it will hop in and out of the jamming band, resulting in the interference being present during some hop intervals and absent in others. These power variations at the receiver are detrimental to its performance e.g. in terms of bit error rate.

Adaptive array antennas are recognized as an efficient way to combat interferers in various communication systems [2, 3]. Interference rejection using an array of sensors have been extensively studied and numerous adaptive algorithms for antenna arrays have been already proposed. Specific attention has been paid recently to spread-spectrum systems but mainly for direct-sequence spread-spectrum systems. In contrast, adaptive algorithms which take into account the specificities of frequency-hopping spread spectrum (FH-SS) communication systems are much fewer in number. [4] studies the effect of frequency hopped signals onto a LMS adaptive array. Nulling of a following jammer is dealt with in [5]. In [6], a control scheme is proposed. At each hop, the next hop is observed in order to estimate the jammers covariance matrix. However, this structure increases the hardware complexity of the RF front-end part. A similar technique is presented in [7] which also results in a complicated hardware configuration due to the utilization of many filters. Finally, the algorithm proposed in [8] requires the location information of the desired signal.

In this paper, we consider the problem of partial-band interference rejection using an array of sensors. In contrast to previous studies, we exploit in advantage the specificities of FH-SS systems. More precisely, we use the fact that the power of the SOI is constant while that of partial-band interferences varies from hop to hop. This enables us to derive a simple algorithm whose principle is to maintain the output power of the array constant, hence the name constant power algorithm (CPA).

2 System model

We consider environments in which there exists one desired signal and powerful interferers. It is assumed that DOAs (Direction of Arrival) of each incoming signal are all different. Figure 1 describes how the desired signal and interferers are distributed over the frequency and time axes. The signal is hopping within a bandwidth B centered around some frequency f_c. At each hop, the desired signal occupies a bandwidth B_h with center frequency f_h. Therefore, the number of frequency slots for the desired signal N_{slot} is defined as $N_{slot} = B/B_h$. T_h denotes the hopping duration. Within the bandwidth B, it is assumed that L partial band interferences are present, each of them occupying a bandwidth $B_{j,\ell}$ ($\ell = 1, \cdots, L$). As the signal of interest hops, the signal to interference plus noise ratio (SINR) varies and it degrades when the desired signal and interferers overlap

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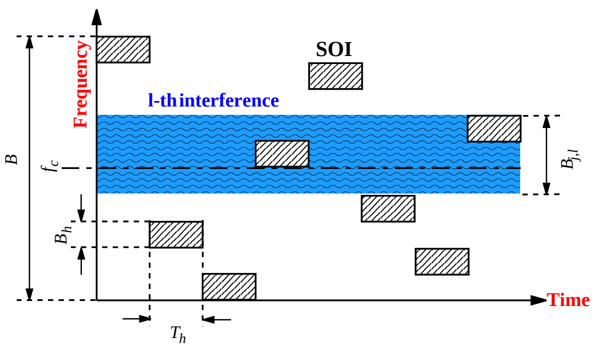

Figure 1: Distribution of desired signal and interferers over the time and frequency axes.

over the frequency axis. Finally, let

$$\eta_\ell = \frac{B_{j,\ell}}{B} \tag{1}$$

denote the fraction of bandwidth that is jammed by the ℓ-th interference.

We assume a M element uniform linear array. The signals received by elements are first down-converted through mixers that are driven by a local oscillator. The local oscillator provides signals whose frequency is dependent on the hopping sequence so that the received signals are de-hopped if the synchronization between the hopping sequence at the receiver and the received signals is precisely achieved. Then, the de-hopped signals are sampled with the sampling rate $f_s = 1/T_s$ samples per second. It follows that $N_s = T_h/T_s$ snapshots can be obtained in each hop. Let θ_d and $\theta_{j,\ell}$ $(\ell = 1, \cdots, L)$ denote the DOA of the desired signal and L interferers, respectively. The DOAs are measured clockwise from the boresight of the array antenna. For later use, let $\boldsymbol{\theta}_j = [\theta_{j,1}, \cdots, \theta_{j,L}]^T$ be the vector of interferences DOAs.

Let us consider the h-th frequency hop. After dehopping, the sampled received signals can be written in vector form as

$$\boldsymbol{x}_h(n) = \boldsymbol{a}_h(\theta_d)s_h(n) + \boldsymbol{A}_h(\boldsymbol{\theta}_j)\boldsymbol{i}_h(n) + \boldsymbol{\xi}_h(n) \tag{2}$$

where $n = 1, \cdots, N_s$ is an integer-valued variable that specifies the sampling time nT_s and the subscript $_h$ stands for the hop index. In (2), $\boldsymbol{a}_h(\theta)$ is the array response to a source with center frequency f_h impinging from direction θ and is given by

$$\boldsymbol{a}_h(\theta) = \begin{bmatrix} 1 & e^{j\frac{2\pi\Delta}{\lambda_h}\sin\theta} & \cdots & e^{j\frac{2\pi\Delta}{\lambda_h}(M-1)\sin\theta} \end{bmatrix}^T \tag{3}$$

where $\lambda_h \triangleq c/f_h$ is the wavelength corresponding to frequency f_h and Δ is the inter-element spacing. $s_h(n)$ is a desired signal whose symbol rate is $f_{sym} = 1/T_{sym}$ bits per second. $\boldsymbol{A}_h(\boldsymbol{\theta}_j) = [\boldsymbol{a}_h(\theta_{j,1}), \cdots, \boldsymbol{a}_h(\theta_{j,L})]$ is a matrix whose columns are formed by the steering vectors of the interferences and $\boldsymbol{i}_h(n) = [i_{h,1}(n), \cdots, i_{h,L}(n)]^T$ is the vector of the L interference signals. Finally, $\boldsymbol{\xi}(n)$ is assumed to be zero-mean complex-valued white Gaussian noise.

The output of the antenna array is obtained as a linear combination of the signals received on each antenna element and can be written as

$$y_h(n) = \boldsymbol{w}^H \boldsymbol{x}_h(n) \tag{4}$$

where $\boldsymbol{w}$ is a weight vector $(M \times 1)$ to be determined. Observe that we consider a single weight vector for the whole band B and not a different weight vector for each hop.

3 Derivation of the CPA method

Before presenting the CPA method, a few remarks are in order. It is well-known that the optimal solution for $\boldsymbol{w}$, i.e. the solution that achieves the maximum signal to interference and noise ratio (SINR) is given by [3]

$$\boldsymbol{w}_h^{\text{opt}} = \boldsymbol{C}_h^{-1}\boldsymbol{a}_h(\theta_d) \tag{5}$$

where $\boldsymbol{C}_h$ is the interference and noise covariance matrix on the h-th hop. However, (5) requires a few conditions that are not met in practice for FH-SS systems. The solution in (5) requires that measurements free of the SOI are available, which is likely to be seldom the case in communication systems. Using the covariance matrix of the signal and interferences in lieu of $\boldsymbol{C}_h$ in (5) usually results in a loss of performance. Also, the formulation in (5) assumes that a different vector is used in each hop. This results in higher computational complexity. Finally, (5) assumes that the DOA of the signal of interest is known which is unrealistic in the framework considered herein.

Therefore, we turn to the derivation of another method which can handle the problems raised by the specific framework considered. Our idea is to use the fact that the signal of interest has a constant power while that of partial-band interferences varies with time due to frequency hopping. Therefore, we look for a weight vector that results in a constant power at the output of the array. Doing so, the algorithm is likely to place nulls towards the interferers since only the latter contribute to power variations. Accordingly, the signal of interest is likely not to be modified since its power is already constant. To summarize, our approach consists of finding the weight vector $\boldsymbol{w}$ that minimizes

$$Q(\boldsymbol{w}) = \mathcal{E}\left\{|P_h - \alpha|^2\right\} \tag{6}$$

where

$$P_h = \frac{1}{N_s}\sum_{n=1}^{N_s} y_h^*(n)y_h(n) \tag{7}$$

is the power of the incoming signals during one hop duration T_h and α is a constant. $\mathcal{E}\{.\}$ stands for statistical expectation. $\boldsymbol{w}$ is obtained iteratively by applying the steepest descent method [9] as

$$\boldsymbol{w}(h+1) = \boldsymbol{w}(h) - \mu \boldsymbol{\nabla}_{\boldsymbol{w}}(h) \quad (8)$$

$$\boldsymbol{\nabla}_{\boldsymbol{w}}(h) = 2\left[P_h(h) - \alpha\right] \Phi(\boldsymbol{w}(h)) \quad (9)$$

$$\Phi(\boldsymbol{w}(h)) = \frac{2}{N_s} \sum_{n=1}^{N_s} \boldsymbol{x}_h(n) \boldsymbol{x}_h^H(n) \boldsymbol{w}(h) = \frac{2}{N_s} \sum_{n=1}^{N_s} \boldsymbol{x}_h(n) y_h^*(n) \quad (10)$$

where h is an integer-valued variable that specifies the hop index. μ is a constant that controls the convergence and stability of the CPA. It should be pointed out the above procedure (8)-(10) results in a computationally efficient algorithm. Additionally, we stress the fact that it does not use (or require) any information about the signal of interest, such as its location.

4 Simulation results

In this section, simulation results are presented to assess the performance of our method. We consider a 8 element array with omni-directionnal antennas. The inter-element spacing is set to $\lambda_c/2$ where $\lambda_c = c/f_c$ is the wavelength corresponding to f_c. In all simulations, a BPSK-modulated signal with differential coding is transmitted as the desired signal from $\theta_d = 20°$. The received signals are de-hopped and band-limited by receiver filters, the roll-off filter (roll-off factor 0.5) and converted to discrete signals with a sampling rate $f_s = 2f_{sym}$ that results in 100 snapshots in each hop. The hopping sequence is generated randomly with a uniform distribution of the instantaneous band in 100 frequency slots. The ratio B/f_c is equal to 0.04. For the CPA algorithm, the parameters μ and α are selected as 0.00095 and 4, respectively. Moreover, the initial weight $\boldsymbol{w}(0)$ is set to $\boldsymbol{w}(0) = \begin{bmatrix}0.1 & 0 & \cdots & 0\end{bmatrix}^T$. In all figures, the output SINR is computed as

$$SINR = \frac{1}{N_{slot}} \sum_{h=1}^{N_{slot}} \frac{P_s |\boldsymbol{w}^H \boldsymbol{a}_h(\theta_s)|^2}{\sum_{\ell=1}^{L} P_{j,\ell} |\boldsymbol{w}^H \boldsymbol{a}_h(\theta_{j,\ell})|^2 + P_n \|\boldsymbol{w}\|^2} \quad (11)$$

where P_d and P_j are the power of the desired signal and interferers included in the input signal, respectively. Similarly, P_n is the noise power in the output signal. Note that the output SINR is computed as the average SINR over the N_{slot} possible frequency hops.

We begin with examining the array beampattern obtained with the CPA method. In a first example, a wide-band interferer is impinging from $\theta_j = -30°$ and occupies 50 % of total bandwidth B, i.e. $\eta = 0.5$. The signal to noise power ratio (S/N) and signal to interference power ratio (S/I) are set to 10dB and -10dB at each element, respectively. Both signals are received by the 8-element linear array so that the maximum output SINR is $10\log_{10} 8 + 10 \simeq 19$dB corresponding to the case the desired signal is combined among elements while all interferers are perfectly cancelled over the space and frequency axes. Figure 2 shows the obtained antenna pattern after 1000 iterations of the CPA algorithm under the environment mentioned above. It is shown that a peak and a null are directed toward each direction θ_d and θ_j. The SINR of the output signal is 18.6dB which is close to the optimum 19dB. This shows that the CPA has converged to a solution which is nearly optimal.

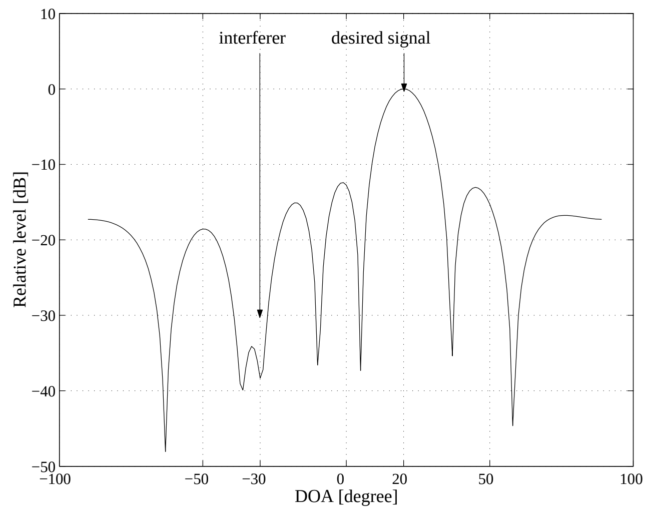

Figure 2: Antenna beampattern after 1000 iterations. (S/N=10dB, S/I=-10dB, $\theta_d = 20°$, $\theta_j = -30°$).

In a second example, we consider two interferers incoming from $\theta_{j,1} = -30°$ and $\theta_{j,2} = 50°$ in addition to the desired signal from $\theta_d = 20°$. S/I$_1$ and S/I$_2$ are set to -10dB while $\eta_1 = \eta_2 = 0.5$ where the subscript $\ell = 1, 2$ corresponds to the ℓ-th interferer. It should be noted that there is no frequency slot without interferer since the two interferers do not overlap in frequency in this example. Figure 3 shows the antenna pattern after 1000 iterations. It can be observed that a peak is generated at the DOA of the desired signal while two nulls are generated toward the DOAs of the interferers even though the whole band is contaminated by the interferers. The SINR of the output signal received by this antenna pattern is 18.2dB. Again, this is close to the optimum value.

Figure 4 compares the normalized convergence curves of the evaluation function $Q(\boldsymbol{w})$ for the two cases described above. Notice that it takes a longer time for the CPA to converge when there are two interferers than when there is only one. However, the convergence is achieved for a rather small number of hops, typically a few hundreds.

We now study the influence of η onto the output SINR. We consider the case of one desired signal and one interferer with S/N=10dB, S/I=-10dB. The DOAs

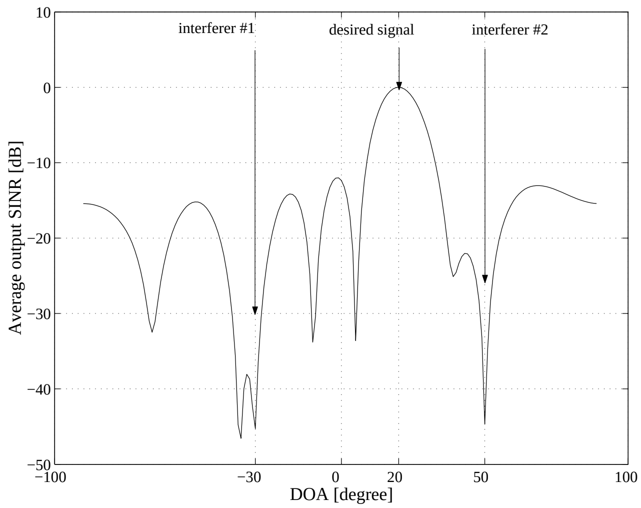

Figure 3: Antenna beampattern after 1000 iterations.(S/N=10dB, S/I$_1$=S/I$_2$=-10dB, $\theta_d = 20°$, $\theta_{j,1} = -30°$, $\theta_{j,2} = 50°$).

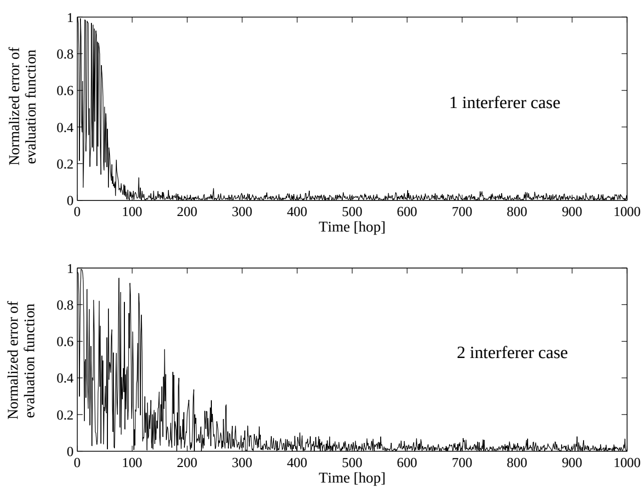

Figure 4: Convergence curves with one or two intereferences.

of the desired signal and interferer are now generated randomly in the range $[-90° < \theta_d, \theta_j < 90°]$. However the angle difference between θ_d and θ_j is kept as $14° < |\theta_d - \theta_j|$ which corresponds to the main-lobe beamwidth of an 8-element linear array. For each simulation 100 Monte-Carlo trials [with different θ_d and θ_j] are run and, for each trial, the SINR is computed according to (11) with $\boldsymbol{w}$ obtained after 1000 iterations of the CPA algorithm. The SINR is then averaged over the 100 trials. Figure 5 examines the influence of η. The average output SINR is seen to degrade as η increases. This is logical since the bandwidth occupied by the interference increases with η. Nevertheless, it should be pointed out that the CPA algorithm succeeds to achieve an average 17dB SINR for η up to $\eta = 0.9$.

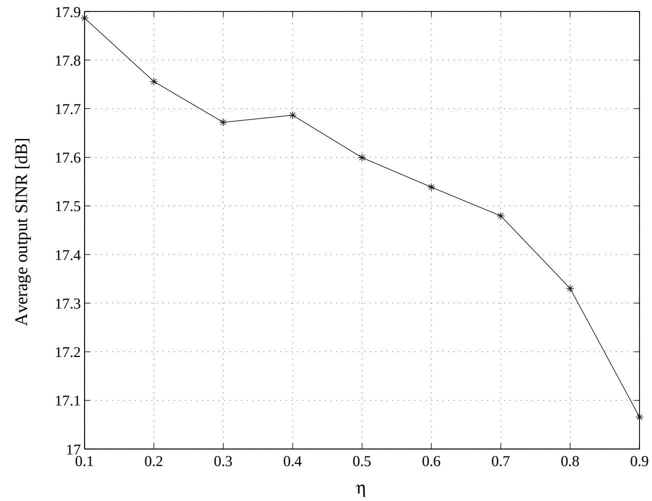

Figure 5: Average output SINR versus η.

5 Conclusion

In this paper, we have proposed a new method for partial-band interference cancellation in frequency-hopping communication systems. The constant power algorithm consists in maintaining a constant power at the output of the array. It is a simple method which does not require any *a priori* information except coarse synchronization of the hop timing of the received signal. It enables to achieve nearly optimum signal to interference and noise ratio.

6 References

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