Blind Adaptive Interference Rejection Based on Doppler/Delay Diversity Between Desired Signal & Interferences/Clutters

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Abstract
This blind adaptive interference rejection algorithm aims to null multiple smart jammers and background clutters impinging from unknown azimuths and elevations using a sensor array with unknown or uncalibrated array manifold. This algorithm, by exploiting the Doppler diversity and delay diversity amongst the desired signal and interferences, separately estimates the spatial correlation matrix ($R_{I+N}$) encompassing only the interference and noise and the spatial correlation matrix ($R_{S+I+N}$) encompassing the desired signal plus interferences and noise. The eigenvector corresponding to the largest generalized eigenvalue of the matrix pencil pair ($R_{S+I+N}, R_{I+N}$) represents the optimum adaptive beamforming weight vector $w^*$ that will maximize the signal-to-interference-plus-noise ratio (SINR). This alternate ISR (intelligence / surveillance / reconnaissance) technology could lower the cost and may enhance the reliability of existing sensor-array adaptive beamforming technology in pulse radar systems, in military and commercial Global Positioning System (GPS) navigation devices, and in UAV electronic surveillance systems.

1 Introduction
This blind adaptive interference rejection algorithm achieves its objective by exploiting the Doppler diversity and delay diversity of the desired signal vis-a-vis its interferences and background clutter and noise. This present approach is applicable to a sensor array of arbitrary and (possibly) unknown array manifold (i.e., not even the nominal sensor array gain/phase or spatial characteristics are needed) to reject multiple smart jammers and clutters interfering from unknown azimuths and elevations and to maximize the desired signal’s power, possibly in the presence of multipaths. Like other adaptive interference rejection algorithms (such as the customary sidelobe canceler [1-5]), this proposed algorithm estimates the effects of interferences, clutter, and noise in the absence of the desired signal. These undesired effects are then subtracted from the data containing the desired signal along with the interferences, clutter, and noise. Customary interference rejection algorithms, such as the sidelobe canceler, assumes the availability and a priori information of certain spatial angular sector or temporal moment where only the interferences and clutter and noise (but not the desired signal) are present so that their denigrating effects may be isolated, estimated, and subtracted. However, a priori information of such spatial diversity or temporal may be unavailable in many missions scenarios. In contrast, this proposed method requires no such a priori known spatial or temporal diversity of the desired signal vis-a-vis the interferences

1 This research work was partly funded by the Power Projection Systems Department of the Johns Hopkins University Applied Physics Laboratory in Laurel, Maryland, U.S.A.
1 An interference rejection algorithm is herein characterized as “blind” in that it performs with no a priori information concerning the desired signal’s and the interferences’ arrival angles, polarization states, Dopplers and delays and without a priori information of the sensor array’s manifold. The transmitted signal’s waveform is, however, assumed known.

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and clutter and noise. Rather, this approach recognizes that smart jammers or background clutters will not generally jam with pin-point precision only the Doppler/delay bin in which the desired signal lies. Rather, smart jammers and background clutter may reasonably be modeled as spreading their jamming power over into several adjacent delay-Doppler bins.

The present algorithm uses those matched filter bank outputs containing the desired signal as well as the interferences and clutter to estimate the spatial correlation matrix \( \mathbf{R}_{S+I+N} \) carrying the desired signal and interferences and noise. This algorithm then uses those other matched filter bank outputs containing the desired signal as well as the interferences and clutter and noise to estimate the spatial correlation matrix \( \mathbf{R}_{I+N} \) carrying the interferences and clutter and noise. Then the eigenvector corresponding to the largest generalized eigenvalue of the matrix pencil pair \( \{ \mathbf{R}_{S+I+N}, \mathbf{R}_{I+N} \} \) represents the optimum beamforming weight vector that will maximize the signal-interference-noise ratio (SINR).

The above procedure assumes that the I+N group of Doppler/delay bins does not include any Doppler/delay bin with significant energy contents from the desired signal; however, the S+I+N group may contain many bins that have little or no energy contents from the desired signal. In other words, it would be unnecessary to identify the desired signal bins precisely, only that all desired signal bins be among the S+I+N bins. Furthermore, a rough but inexact correspondence between the I+N bins and the actual set of jammed bins would also suffice for the proposed algorithm. This detection assumption is reasonable in many radar and GPS applications because: (a) this detection could be achieved by the Generalized Likelihood Ratio Test (GLRT), (b) a priori knowledge may exist for very rough estimates of the approximate Doppler and range characteristics of the targets or the GPS satellites, (c) in heavily jammed scenarios, jamming effectively affects all Doppler/range bins, (d) in certain missions, the echo reflected off the intended target may have the most direct path and thus the shortest delay relative to hostile smart jammers and background clutters.

The algorithm under development has been inspired by an interference cancelation algorithm that Zoltowski & Ramos [8,9] and Zoltowski, Chen & Ramos [10] have developed for code-division multiple-access (CDMA) wireless communications.

2 Relevance to Various Weapon Programs

This alternate adaptive beamforming approach maximizes the SINR with no a priori information of the nominal or actual array spacing/gain/phase characteristics and no a priori arrival angle information of any of the targets or jammers. These innovative features imply: (a) improved operational versatility by allowing antennas to be arbitrarily placed at unknown locations on the battlefield, or on aircrafts, or aboard ships, (b) improved affordability and operational ease due to great tolerance of imprecisely manufactured antennas and due to circumvention of the need of array calibration, (c) increased reliability and operational integrity due to improved systems robustness against failure of individual antennas, (d) improved responsiveness to complex and unascertained target/jamming scenarios.

This proposed adaptive nulling technique is applicable to most sensor array pulse radar or sonar system, for example, advanced array sensors for the detection and engagement of Theater Ballistic Missiles or the High Speed Anti-radiation Missile (HARM) program.

This alternate technology also possesses the potential to improve jamming immunity, responsiveness and hardware affordability of military/commercial Missile-GPS navigators. Precision-guided missiles, civilian jet planes and civilian intelligent highway systems all rely heavily on GPS. Such military or civilian GPS navigators often suffer from deliberate broadband or narrowband jamming from smart jammers and/or unintentional interference from RF transmitters on-board the same aircraft and nearby aircrafts and ground vehicles or from nearby ground transmission stations. Furthermore, antenna costs often embody a significant share of the total cost of many GPS navigators. The intrinsic flexibility and tolerance on antenna characteristics in the proposed algorithm could imply notable reduction in antenna costs, and thus in navigator unit cost. Relevant defense-related GPS programs include the Tomahawk
Baseline Improvement Program (TBIP) or Tomahawk Block 4 program, the Standoff Land Attack Missie (SLAM), the Joint Direct Attack Munitions (JDAM) program, the F/A-18 Hornet and indeed any other GPS-guided missile or aircraft program. This algorithm is also relevant to civilian GPS programs funded by the U.S. Federal Aviation Administration [6] and the U.S. Department of Transportation.

This alternate technology can also enhance the robustness of unmanned aerial vehicle (UAV) programs, such as the Predator program and the Pioneer program, where a fleet of UAVs are airborne over the battlefield to perform surveillance missions. Each individual UAV operates as a sensor; and the entire fleet comprises a sensor array to detect electromagnetic radiation from the battlefield. The relative spatial position among the UAVs is necessary for such surveillance and this location information is ordinarily provided by GPS. However, under heavy jamming, GPS could be drowned out by interference and the relative spatial relation among the UAVs become unavailable. In effect, the UAV fleet becomes like a sensor array with unknown sensor locations—which is exactly the sensor-array model used in this proposal. Thus, the beamforming technique proposed herein would enable the UAV fleet to perform its electronic surveillance missions even when heavy jamming denies the UAV fleet their GPS position information.

3 Mathematical Data Model

Point emitters from the far-field, having traveled through an homogeneous isotropic medium, impinge upon an array of sensors of arbitrary and (possibly) unknown gain/phase characteristics and location. The $k$th signal pulse train consists of $N$ pulses, $\{s_{k,1}(t), \ldots, s_{k,N}(t)\}$, where $s_{k,n}(t)$ may be non-zero only in the time interval $t \in [(n-1)T_p, nT_p]$, where $T_p$ signifies the pulse repetition period. The $n$th pulse of the $k$th signal pulse train impinging upon the $l$th sensor at time $t$, thus, registers the scalar measurement:

$$p_k s_{k,n}(t) c_l(\theta_k, \phi_k) q_l(\theta_k, \phi_k)$$  \hspace{1cm} (1)

where $P_k$ denotes the $k$th signal’s power, $0 \leq \theta_k < \pi$ symbolizes the signal’s elevation angle measured from the vertical $z$-axis, $0 \leq \phi_k < 2\pi$ symbolizes the azimuth angle, $c_l(\theta_k, \phi_k)$ signifies the $l$th sensor’s scalar response to a unit-power source incident from $(\theta_k, \phi_k)$, $q_l(\theta_k, \phi_k)$, the spatial phase-factor for the $k$th narrowband\(^2\) incident source to the $l$th sensor located at $(x_i, y_i, z_i)$ equals:

$$q_l(\theta_k, \phi_k) \equiv e^{j2\pi \left(\frac{\lambda u_x + u_y \phi_k + u_z \theta_k}{\lambda}\right)}$$  \hspace{1cm} (2)

where $\lambda$ refers to the signals’ carrier wavelength, $u_x \equiv \sin \theta_k \cos \phi_k$ represents the direction cosine along the $x$-axis, $u_y \equiv \sin \theta_k \sin \phi_k$ symbolizes the direction cosine along the $y$-axis, and $u_z \equiv \cos \theta_k$ signifies the direction cosine along the $z$-axis. If the $k$th incident pulse train (which may be either the desired signal or smart jammer or clutter) would be a Doppler-shifted and delayed version of the transmitted waveform $h(t)$:

$$s_{k,n}(t) = h((1 + f_k)(t - \tau_k - nT_p))$$  \hspace{1cm} (3)

where $f_k$ represents the Doppler shift and $\tau_k$ symbolizes the delay.

With a total of $K \leq L$ co-channel signals and complex-valued additive white noise for each sensor’s measurement, the $l$th sensor produces the measurement:

$$z_l(t) = \sum_{k=1}^{K} p_k s_{k,n}(t) c_l(\theta_k, \phi_k) q_l(\theta_k, \phi_k) + n_l(t),$$  \hspace{1cm} (4)

for $l = 1, \ldots, L$; and \((n - 1)T_p \leq t \leq nT_p$.

\(^2\)These incident signals are narrowband in that their bandwidths are very small compared to the inverse of the wavefronts’ transit time across the array. The case involving broadband signals may be reduced to a set of narrowband problems via the use of a comb of narrowband filters.
For the entire $L$-element arbitrarily spaced sensor array there exists a $L \times 1$ vector measurement at each $t$:

$$
z(t) \overset{\text{def}}{=} \begin{bmatrix} z_1(t) \\ \vdots \\ z_L(t) \end{bmatrix} = \sum_{k=1}^{K} P_k s_{k,n}(t) a_1(\theta_k, \phi_k) + n(t) 
$$

where $\otimes$ denotes the Kronecker product and $A$ represents the $L \times K$ matrix:

$$
A \overset{\text{def}}{=} \begin{bmatrix} a(\theta_1, \phi_1), \ldots, a(\theta_K, \phi_K) \end{bmatrix}
$$

(7)

$$
a_{k,n}(t) \overset{\text{def}}{=} \begin{bmatrix} c_1(\theta_k, \phi_k) g_1(\theta_k, \phi_k) \\ \vdots \\ c_L(\theta_k, \phi_k) g_L(\theta_k, \phi_k) \end{bmatrix}
$$

(8)

$$
s_n(t) \overset{\text{def}}{=} \begin{bmatrix} s_{1,n}(t) \\ \vdots \\ s_{K,n}(t) \end{bmatrix} \quad \text{and} \quad n(t) \overset{\text{def}}{=} \begin{bmatrix} n_1(t) \\ \vdots \\ n_L(t) \end{bmatrix}
$$

(9)

where $n_i(t)$ symbolizes the additive white noise at the $l$th sensor.

The matched filter bank, through which the array’s measurement samples are to be passed, consists of $M$ matched filters, each of which embodies a Doppler-shifted version of the transmitted signal waveform $h(t)$. The $m$th matched filter is characterized by its impulse response $h^{(m)}(t) = h((1 + f_m)f_t)$, where $f_m \neq f_n$, for all $m \neq n$. The output of the $m$th matched filter at the $l$th sensor, denoted $y^{(m)}_l(i)$, equals:

$$
y^{(m)}_l(i) = h^{(m)}(i) \ast z_l(i)
$$

(10)

where $\ast$ symbolizes the convolution operator. At the $m$th Doppler and the $i$th delay bin of the entire $M \times I$ matched filter bank (where $I$ equals the integer such that $y^{(m)}_l(i) = 0$, for all $i > I$ and for all $l$ and all $m$), the filter output equals the $L \times 1$ vector:

$$
y^{(m)}(i) = \begin{bmatrix} y^{(m)}_1(i) \\ \vdots \\ y^{(m)}_L(i) \end{bmatrix}
$$

(11)

There thus exist $MI$ number of $L \times 1$ vectors of match filter output.

4 Algorithmic details

The key assumption behind this algorithm is that smart jammers or background clutters will generally be unable to achieve pin-point precision jamming only the one Doppler/delay bin in which a target lies. Rather, smart jammers and background clutters may reasonably be modeled as spreading their jamming power over into several adjacent Doppler/delay bins. If the signal-carrying Doppler/delay bins may be roughly distinguished from those other bins with only interferences and noise, then the signal-interference-noise spatial correlation matrix $R_s + I + N$ and the interference-noise spatial correlation matrix $R_{I + N}$ may each be estimated from matched filter bank outputs corresponding to each of these two classes. It would be unnecessary to identify the signal-carrying bins precisely, only that the desired signal bins be among the $S+I+N$ bins. Furthermore, a rough but inexact correspondence between the $I+I$ bins and the actual set of jammed bins would suffice in this proposed algorithm. This detection assumption is reasonable in many radar and GPS applications because: (a) this detection could be achieved by the Generalized Likelihood Ratio Test.
(GLRT), (b) a priori knowledge may exist for very rough estimates of the approximate Doppler and range characteristics of the targets or the GPS satellites, (c) in heavily jammed scenarios, jamming effectively affects all Doppler/range bins, (d) in certain missions, the echo reflected off the intended target may have the most direct path and thus the shortest delay relative to hostile smart jammers and background clutters.

Assuming a successful classification of the bins into the \( S+I+N \) and \( S+N \) classes, \( \hat{R}_{S+I+N} \) and \( \hat{R}_{I+N} \) may be estimated as follows:

\[
\hat{R}_{S+I+N} \overset{\text{def}}{=} Y_{S+I+N} Y_{S+I+N}^H \tag{12}
\]

\[
\hat{R}_{I+N} \overset{\text{def}}{=} Y_{I+N} Y_{I+N}^H \tag{13}
\]

where the columns of \( Y_{S+I+N} \) comprises all \( y^{(m)}(i) \) classified into the \( S+I+N \) group and the columns of \( Y_{I+N} \) comprises all \( y^{(m)}(i) \) classified into the \( I+N \) group, and \( H \) denotes the Hermitian operation.

The beamformer signal-to-interference-plus-noise (SINR) ratio may be defined in terms of \( R_{S+I+N} \) and \( R_{I+N} \) as

\[
\text{SINR} \overset{\text{def}}{=} \frac{w^H (\hat{R}_{S+I+N} - \hat{R}_{I+N}) w}{w^H \hat{R}_{I+N} w} \tag{14}
\]

\[
= \frac{w^H \hat{R}_{S+I+N} w}{w^H \hat{R}_{I+N} w} - 1 \tag{15}
\]

The SINR would be maximized by the \( L \times 1 \) weight vector

\[
w^* \overset{\text{def}}{=} \arg \max_w \frac{w^H \hat{R}_{S+I+N} w}{w^H \hat{R}_{I+N} w} \tag{16}
\]

which equals the eigenvector corresponding to the largest generalized eigenvalue of the matrix pencil pair \( \{ R_{S+I+N}, R_{I+N} \} \).

Note that unlike the well known sidelobe canceler beamformer, the optimum null beamforming weight vector is determined here with no a priori knowledge of the targets' arrival angles or the jammers' arrival angles. In the sidelobe canceler approach, the main beam is pointed towards the desired targets while the side beams are directed ideally towards only the jammers and noise. Thus, the jammers and noise portion of the main lobe's response may be minimized by subtracting out the side lobe responses. This sidelobe canceler approach has thus implicitly assumed that the targets and the jammers may be spatially distinguishable based on a priori knowledge of their respective angles-of-arrival or that they may be temporarily distinguishable based on the a priori knowledge of their different incident time intervals. In contrast, the present approach presumes no such a priori knowledge.

Furthermore, the proposed algorithm would not require any knowledge of the array manifold, thereby eliminating the need for frequent and expensive array calibration and minimizing the denigrating effects of array miscalibration—major problems in most other sensor-array beamforming methods.

5 Reference


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