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A NEW-PARTIALLY ADAPTIVE SPACE-TIME FILTERING TECHNIQUE FOR INTERFERENCE SUPPRESSION IN PHASED ARRAY RADAR SYSTEMS

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Abstract__This paper presents a new-partial adaptive space-time filtering technique for clutter and/or jammer suppression in phased array radar systems. In this work, two proposed filter configurations are considered. These filters utilize a rectangular phased array antenna (two-dimension) with "$N_X N_Y$" sensors as an adaptive space-time signal-processing unit. The first configuration is referred to as fully adaptive space-time filter "FASTF", and the second configuration is referred to as partially adaptive space-time filter "PASTF". A computer code has been developed on MATLAB-R12 to simulate the operation processes of both filters as well as the signal and interference environment. The objective is to analyze, investigate and evaluate the performance of the new-presented partial adaptive filter versus the full adaptive one in case of search and tracking radar systems. Results of simulation indicate that the new-partially adaptive filter has a good performance (output signal to interference plus noise ratio or improvement factor) as close as to the full adaptive filter for the same interference conditions. In addition, partially adaptive filter is less complex than the first filter's configuration. Also, a tremendous reduction in the overall processing time has been achieved using the partially adaptive filter's configuration.

Keyword: Radar and Communications

I. Introduction

With the rapid progress in computer technology, particularly, the processor speed, real-time adaptive signal processors became widely used in radar and communication applications [1-2]. These processors can dynamically enhance the desired signal reception and suppress the undesired one through an adaptive algorithm. This type of processing is based on the difference in the space and/or time characteristics of both desired and undesired signals. There are two common types of adaptive signal processing. By using a single complex variable weight at each array element, a deep null can be placed in the direction of the interference (undesired signals). This type of processing is referred to as narrow band space processing or beam forming [3]. However, using a set of complex variable weights (adaptive FIR filter) with each array sensor, a maximum enhancement of the desired signal can be achieved. This type of processing is referred to as adaptive space-time processing or broadband space processing [4-9]. These weights are used to adjust the phase and amplitude of the intercepted signal according to its desired direction and the interference environment. In practice, cost, complexity, and processing time are considered to be the main important parameters to determine the performance of

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an adaptive space-time signal processor for a specific application. Therefore, a compromise between these parameters and the required system performance should be done. To elevate the conflicting problem between these parameters, a partial adaptivity approach is used [3]. In an adaptive filter and/or system, concept of the partial adaptivity can be implemented on the hardware level and/or the software level. The hardware level approach is to physically reduce the total number of adaptive channels of the filter (or system). In software approach, for a specific number of adaptive channels, the rank of the interference space-time covariance matrix (total number of adjustable weights) has to be reduced. Using these levels, a reduction in the over all processing time of an adaptive filter and/or system can be achieved. This includes matrix estimation and computation of the optimum adaptive weights. However, the only restriction to use these levels is that the partial adaptive filter (or system) should achieve a performance relatively close to that achieved by the full adaptive filter (or system). Of course, as we all know, the meaning of the word relatively close is mainly dependent on a specific radar (or communication) application.

Configuration and concept of the proposed filters are discussed in details and presented in section (II). Section (III) presents a mathematical formulation of noise, clutter, jammer and target echo for both full and partial adaptive filters. Simulation results of the proposed filters are investigated and presented in section (V). This includes different clutter and jammer types. Performances of both FASTF and PASTF in search and tracking radar systems are also evaluated and presented in this section. Finally, the paper is concluded in section (VI).

II. The Proposed Full and Partial Adaptive Space-Time Filters: (configuration and concept):

In this section, configuration and concept of the two proposed filters are discussed in details. The first filter configuration is referred to as full adaptive space-time filter "FASTF", and it is presented in Fig.1-a. This filter is composed of three main units. This includes phased array antenna unit, down frequency conversion and amplification unit and adaptive signal processing unit. Such filter is a direct generalization of the one dimension (linear array) FASTF reported in [8-9]. The phased array antenna unit is a rectangular array having "$N_x x N_y$" elements (two-dimension). Each array element (sensor) is connected to the down frequency conversion and amplification unit through RF amplifier, mixer and local oscillator. The IF output from each array element is connected to the adaptive signal processing unit through an adaptive channel. This channel has IF multiplier, analog low pass filter, A/D converter, and adaptive non-recursive filter. This filter has "N_p" complex variable weights to control the phase and amplitude of the desired signal through an adaptive processor. The outputs from each filter's channel are added together to form the output of the FASTF. The second configuration is presented in Fig.1-b, and it is referred to as partial adaptive space-time filter "PASTF". In this 2-D filter's configuration, complexity of the adaptive signal-processing unit including cost and processing time has been reduced using both levels of the partial adaptivity approach (hardware and software). First, a physically reduction in the total number of adjustable weights per each channel has been performed. An integrator unit is used to compensate for this hardware reduction at the final filter's output. Consequently, a single adjustable weight per each channel is used instead of a FIR filter as compared to the signal-processing unit of the FASTF. Second, these channels are organized in different adaptive groups called layers to perform the second level of the partial adaptivity (software level). Each adaptive layer has a separate high-speed processor. The processor of each layer estimates the layer matrix (of a reduced rank), during a specified learning period, and then it computes the optimum weights of the layer per each intercepted echo pulse. A summarization of the main differences between the two proposed filter configurations is presented in Table 1. This includes rank of the space-time covariance matrix, observation vector size, and the number of adjustable weights, adaptive processors, and integrator units.

III. Problem Formulation

Consider a radar system has a rectangular phased array antenna having "N_xN_Y" elements as shown in Fig.1. The cartesian geometry of such antenna configuration is illustrated in Fig.2. Assume that this antenna intercepts an echo from a point target located on the free space. The target location is characterized by its range $\mathbf{R_T}$, and its azimuth and elevation angles (**β&ε**) with respect to the boresight of the array antenna. The power of the intercepted signal is related to its range at certain angular location (**β&ε**). These parameters are considered in our mathematical formulation. The complex envelope of the intercepted data is arranged in a vector form notation, and it is used to estimate the space-time covariance matrix of the interference environment. The direct matrix inversion algorithm (DMI) is used compute the optimum weight vector of the full and partial adaptive space-time filters [5-7]. Thus, the filter has a maximum signal to interference plus noise ratio at its output using this optimum weight vector (Wiener's filter theorem). In the following subsections, a mathematical formulation of interference and target echo signals for both filters is discussed in details.

A. Full Adaptive S-T Filter (FASTF):

Referring to Fig.1-a, the signal intercepted by the two-dimensional phased array antenna is arranged in a vector form notation as:

$$\underline{X}(t) = \left[\underline{X}_1(t)\underline{X}_2(t)....\ \underline{X}_m(t)......\ \underline{X}_{N_x}(t)\right]^T_{(N_xN_yx1)}, \quad , m=1,2,.,N_x \qquad (1)$$

Where,

$$\underline{X}_m(t) = \left[\underline{X}_{m1}(t)\underline{X}_{m2}(t)...\underline{X}_{mn}(t).....\underline{X}_{mNy}(t)\right]^T_{(N_yx1)}, \quad , n=1,2,.,N_y \qquad (2)$$

The continuous time signal X_{mn} (t) at the output of m^{th} and n^{th} array element is sampled at each pulse repetition period (T_r) to feed an adaptive non-recursive filter, and it is expressed as

$$\underline{X}_{mn} = \left[X_{mn}(0), X_{mn}(T_r),.., X_{mn}(k),...X_{mn}((N_p-1)T_r)\right]^T_{(N_px1)} \quad , k=0,1,..,(N_p\text{-}1) \qquad (3)$$

Thus, complex notation of the vector $\underline{X}$(t) of equation (1) is re-expressed as

$$\underline{X} = \left[\underline{X}_1, \underline{X}_2 \cdots \underline{X}_m \cdots \underline{X}_{N_x}\right]^T_{(N_xN_yN_px1)} \quad , 1\le m\le N_x \qquad (4)$$

Where,

$$\underline{X}_m = \left[\underline{X}_{m1}\cdot\underline{X}_{m2},.....\ \underline{X}_{mn},....\ \underline{X}_{mN_y}\right]^T_{(N_yN_px1)} \quad , 1\le n\le N_y \qquad (5)$$

$$\underline{X}_{mn} = \left[\underline{X}_{mn}(1), \underline{X}_{mn}(2),...., \underline{X}_{mn}(k)\right]^T_{(N_px1)} \quad , 1\le k\le N_p \qquad (6)$$

The observation vector of Equation (4) is re-expressed a real notation as:

$$\underline{X} = \left[\underline{XI} : \underline{XQ}\right]_{(2N_xN_yN_px1)} \qquad (7)$$

Where, XI and XQ are the inphase and quadrature samples of the all channels of the filter respectively, and they are expressed in vector notation as

$$\underline{XI} = \left[\underline{XI}_1, \underline{XI}_2, \ldots, \underline{XI}_m, \ldots, \underline{XI}_{N_x}\right]^T_{(N_x N_y N_p x1)}, \quad 1 \le m \le N_x \qquad (8\text{-}1)$$

$$\underline{XQ} = \left[\underline{XQ}_1\, \underline{XQ}_2 \cdots \underline{XQ}_m \cdots \underline{XQ}_{N_x}\right]^T_{(N_x N_y N_p x1)}, \quad 1 \le m \le N_x \qquad (8\text{-}2)$$

Where,

$$\underline{XI}_m = \left[\underline{XI}_{m1}\, \underline{XI}_{m2} \cdots \underline{XI}_{mn} \cdots \underline{XI}_{mN_y}\right]^T_{(N_y N_p x1)}, \quad 1 \le n \le N_y \qquad (9\text{-}1)$$

$$\underline{XQ}_m = \left[\underline{XQ}_{m1}\, \underline{XQ}_{m2} \cdots \underline{XQ}_{mn} \cdots \underline{XQ}_{mN_y}\right]^T_{(N_y N_p x1)}, \quad 1 \le n \le N_y \qquad (9\text{-}2)$$

Where, $\underline{XI}_{mn}$ and $\underline{XQ}_{mn}$ are (N_px1) column vectors represent the samples (or snapshots) of the inphase and quadrature of the mth and nth array element channel of the filter respectively, and they are given by

$$\underline{XI}_{mn} = \left[\underline{XI}^{(1)}_{mn}\, \underline{XI}^{(2)}_{mn} \cdots \underline{XI}^{(k)}_{mn} \cdots \underline{XI}^{(N_p)}_{mn}\right]_{(N_p x1)}, \quad k=1,2,..,N_p \qquad (10\text{-}1)$$

$$\underline{XQ}_{mn} = \left[\underline{XQ}^{(1)}_{mn}\, \underline{XQ}^{(2)}_{mn} \cdots \underline{XQ}^{(k)}_{mn} \cdots \underline{XQ}^{(N_p)}_{mn}\right]_{(N_p x1)}, \quad k=1,2,..,N_p \qquad (10\text{-}2)$$

The mth and nth array element value of the observation vector $\underline{X}$ in equation (4) due to ℓ^{th} source is $X^{\ell}_{mn}(k)$. This value is expressed in terms of the source complex envelope and it is given by

$$\begin{aligned} x^{\ell}_{mn}(k) &= A_\ell(k) e^{j\pi(m-1)\sin\varepsilon_\ell Cos\beta_\ell}\, e^{j\pi(n-1)Sin\varepsilon_\ell Sin\beta_\ell} \\ &= A_\ell(k).e^{jB_\ell(m,n)} \end{aligned} \qquad (11)$$

Where, $A_\ell(k)$ denotes the complex envelope amplitude and the phase of the a source located at certain angular location, and it is expressed as

$$B_\ell(m,n) = X_\ell(m) + Y_\ell(n) \qquad (12\text{-}1)$$

Where,

$$X_\ell(m) = \pi(m-1)\sin\varepsilon_\ell \cos\beta_\ell \qquad (12\text{-}2)$$

$$Y_\ell(n) = \pi(n-1)\sin\varepsilon_\ell \sin\beta_\ell \qquad (12\text{-}3)$$

A1. Target Echo Representation:

In our analysis, complex envelope representation is used to model the target echo as given in equation (4) through equation (12). The mth and nth inphase "$\underline{X}^{mn}_{SI}$" and quadrature "$\underline{X}^{mn}_{SQ}$" samples of the target echo are given by

$$\underline{X}^{mn}_{SI}(k) \stackrel{\Delta}{=} s(k)\cos(\psi_{mn}(k)) \qquad (13\text{-}1)$$

, and

$$\underline{X}^{mn}_{SQ}(k) \stackrel{\Delta}{=} s(k)\sin(\psi_{mn}(k)) \qquad (13\text{-}2)$$

respectively, where,

$$\psi_{mn}(k) = 2\pi(k-1)K_S + B_S(m,n) \qquad (14\text{-}1)$$

denotes the total signal phase, and $K_S = (F_{ds}/F_r)$ denotes the normalized target doppler frequency. The additional phase between the different array channels is given by

$$B_S(m,n) = X_S(m) + Y_S(n), \tag{14-2}$$

Where,

$$X_S(m) = \pi(m-1)\sin(\varepsilon - \varepsilon_S)\cos(\beta - \beta_S), \tag{14-3}$$

$$Y_S(n) = \pi(n-1)\sin(\varepsilon - \varepsilon_S)\sin(\beta - \beta_S), \tag{14-4}$$

Where, β_s & ε_s denote the desired angular location of the target echo with respect to the boresight of the array antenna.

A2. Clutter Return Representation:

Similar to the target echo given by equation (13), the clutter return due to "ℓ" scatterer sources (ℓ =1,2,…,N_C) can be represented by its complex envelope, and it can be expressed as:

$$X_{CI}^{mn}(k) = \sum_{\ell=1}^{N_c} C_\ell(k)\cos\psi_{mn}(\ell,k) \tag{15-1}$$

$$X_{CQ}^{mn}(k) = \sum_{\ell=1}^{N_c} C_\ell(k)\sin\psi_{mn}(\ell,k) \tag{15-2}$$

Where,

$$\psi_{mn}(\ell,k) = \psi_C(\ell,k) + B_C(\ell,m,n) \tag{16-1}$$

is the total clutter phase variation, and $\psi_C(\ell,k) = 2\pi F_C(\ell,k)$ denotes the random phase of the clutter source due to its random motion (weather and chaff). The clutter doppler frequency "F_c" equals to zero in case of fixed clutter source (ground), while in case of weather or chaff, it has a random value. Similarly, as given in equation (14-2) through equation (14-4), the additional phase between the different elements of array antenna for the clutter return is expressed as

$$B_C(\ell,m,n) = X_C(\ell,m) + Y_C(\ell,n) \tag{16-2}$$

Where,

$$X_C(\ell,m) = \pi(m-1)\sin(\varepsilon_\ell - \varepsilon_S)\cos(\beta_\ell - \beta_S) \tag{16-3}$$

$$Y_C(\ell,n) = \pi(n-1)\sin(\varepsilon_\ell - \varepsilon_S)\sin(\beta_\ell - \beta_S) \tag{16-4}$$

A3. Jammer Signal Representation:

Similar to the target echo given by equation (13), the jamming signal due to "ℓ" jammer sources (ℓ =1,2,…,N_J) can be represented by its complex envelope, and it can be expressed as:

$$X_{JI}^{mn}(k) = \sum_{\ell=1}^{N_j} J_\ell(k)\cos(\psi_{mn}(\ell,k)) \tag{17-1}$$

$$X_{JQ}^{mn}(k) = \sum_{\ell=1}^{N_j} J_\ell(k)\sin(\psi_{mn}(\ell,k)) \tag{17-2}$$

Where,

$$\psi_{mn}(\ell,k) = \psi_J(\ell,k) + B_J(\ell,m,n) \tag{18-1}$$

is the total jammer phase variation, and $\psi_J(\ell,k)$ denotes the random phase of the jammer sources. Similar to equation (14-2), the additional phase between the antenna array elements for the jamming signal is expressed as

$$B_J^{mn}(\ell,m,n) = X_J(\ell,m) + Y_J(\ell,n) \tag{18-2}$$

Where,

$$X_J(\ell,m) = \pi(m-1)\sin(\varepsilon_\ell - \varepsilon_S)\cos(\beta_\ell - \beta_S) \tag{18-3}$$

$$Y_J(\ell,n) = \pi(n-1)\sin(\varepsilon_\ell - \varepsilon_S)\sin(\beta_\ell - \beta_S) \tag{18-4}$$

The jammer phase $\psi_J(\ell,k)$ in equation (18-1) is expressed as

$$\psi_J(\ell,k) = 2\pi(k-1)K_J + \phi(\ell,k) \tag{18-5}$$

Where, $K_J = (F_{dj}/F_r)$ denotes the jammer normalized doppler frequency, and the 2nd term in equation (18-5) represents the random phase fluctuation of the jammer source.

B. Partial Adaptive S-T Filter (PASTF):

Referring to Fig.1-b, the signal intercepted by the two-dimensional phased array antenna is organized in "N_L" groups, each one is referred to as layer. Thus, the intercepted data of a layer number "n" (n=1,2,…,N_L= N_y) is arranged in a complex vector notation as:

$$\underline{X}_n(t) = \left[x_n^1(t), x_n^2(t), \ldots, x_n^m(t), \ldots, x_n^{N_E}(t)\right]^T_{(N_E \times 1)} \quad \begin{matrix} 1 \le m \le N_E \\ 1 \le n \le N_L \end{matrix} \tag{19}$$

Where, the element vector $x_n^m(t)$ is expressed as

$$x_n^m(t) = A_n^m(t).e^{jB_\ell(t,m,n)} \tag{20}$$

Where, $B_\ell(t,m,n)$ denotes the phase shift between the antenna array elements, and it s given by equation (12). The vector $\underline{X}_n(t)$ is sampled each (T_r) to form the observation vector of the sampled data of each layer. The signal-processing unit of the PASTF has N_L layers, each one has its own adaptive high-speed processor. A separate and independent processing is performed in parallel for each sampled observation vector of each layer using its processor. This processing includes three processes. First, the sampled observation vector of the layer is used to estimate its space-time covariance matrix during a specified learning period. Second, the optimum weight vector of each layer is computed and updated per each intercepted echo pulse using the corresponding layer matrix. Third, the outputs from each layer's channel are summed to form the layer output. Finally, the outputs from each layer are added together to form the final filter's output.

B1. Target Echo Representation:

Again, the complex envelope representation is used to model the target echo. The real vector notation of equation (19) for the echo $\underline{X}_S^n(t)$ is re-arranged as

$$\underline{X}_S^n(t) = \left[\underline{X}_{SI}^n(t) : \underline{X}_{SQ}^n(t)\right]^T_{(2N_E \times 1)} \tag{21}$$

Where, $\underline{X}^n_{SI}(t)$ and $\underline{X}^n_{SQ}(t)$ are the inphase and quadrature samples of the layer number "n" respectively, and they are arranged in vector notation as

$$\underline{X}_{SI}^{n}(t) = \left[x_{SI}^{1n}(t), x_{SI}^{2n}(t),......, x_{SI}^{mn}(t),...., x_{SI}^{N_E n}(t)\right]_{(N_E x1)}^{T} \quad (22\text{-}1)$$

$$\underline{X}_{SQ}^{n}(t) = \left[x_{SQ}^{1n}(t), x_{SQ}^{2n}(t),......, x_{SQ}^{mn}(t),...., x_{SQ}^{N_E n}(t)\right]_{(N_E x1)}^{T} \quad (22\text{-}2)$$

These vectors sampled at each "T_r", then, the samples $x_{SI}^{mn}(k)$ and $x_{SQ}^{mn}(k)$ are given by equation (13). The total signal phase, the normalized target doppler frequency, and the additional phase between the different array channels are given by equation (14).

B2. Clutter Return Representation:

The clutter return due to "ℓ" scatterer sources (ℓ=1,2,...,N_C) can be represented by its complex envelope. Similar to the signal vector represented by equation (21) and equation (22), the clutter complex envelope can be expressed in vector notation. The inphase and quadrature samples $x_{CI}^{mn}(k)$ and $x_{SQ}^{mn}(k)$ of the clutter vector are given by equation (15). The total clutter phase variation is given by equation (16).

B3. The Jammer Signal Representation:

Similar to the target echo vector represented by equation (21) and equation (22), the jammer complex envelope can be expressed in vector form. The inphase and quadrature samples $x_{JI}^{mn}(k)$ and $x_{JQ}^{mn}(k)$ of the jammer vector are given by equation (17). The total jammer phase variation is given by equation (18).

IV. Simulation Assumption

The interference space-time covariance matrix of the full adaptive space-time filter can be estimated using the DMI algorithm due to its rapid convergence [5-6]. The estimated matrix is expressed as

$$\hat{M}_{ST} = 1/k_1 \sum_{i=1}^{k_1} \{X_i . X_i^T \quad (23\text{-}1)$$

Where, K_1 is number of observation vectors needed to estimate the interference space-time covariance matrix, and T denotes the transpose operator. This number of observation vectors is related to the matrix rank to be estimated, i.e., $K_1=N_K*(2N_PN_XN_Y)$, where "N_K" should be greater than two to minimize the estimation error. The observation vector "X" represents the intercepted data assuming the signal is absent, i.e., $\underline{X}=\underline{X}_C$(clutter)+$\underline{X}_J$(jammer)+$\underline{X}_S$(signal), and $\underline{X}_S=\underline{0}$. The FASTF has a specified learning period equal to $K_1(T_r-1)$ during which the estimated matrix is updated to keep the filer performance above certain required threshold. The observation vector $\underline{X}$ has a size equal to ($2N_xN_yN_p$ x 1) and the matrix size is ($2N_xN_yN_p$ x $2N_xN_yN_p$). This matrix has a huge size e.g., if N_x=10, N_y=10, & N_p=5, the matrix size is (1000 x 1000). Therefore, an especial purpose processor having a high-speed is required to perform the matrix estimation as well as matrix inversion. On the other hand, the overall processing time of the filter is limited in practice by the applications i.e., on line and/or off line processing. In fact, for either search or tracking radar systems, on line processing is required to perform the target extraction information. To elevate this problem, the partial adaptivity concept should be used to reduce complexity, processing time, and cost of the adaptive signal-processing module. In our proposed PASTF presented in Fig.1-b, the partial adaptively has been performed as follows. First, the space-time covariance matrix of the FASTF depicted in Fig.1-a is divided into "N_L" layer sub-matrices, each of which has a size equal to (N_ExN_E). Second, the adaptive FIR filter in each sensor's channel of Fig.1-a is replaced by a single complex weight, i.e., each adaptive channel in each layer has a single adjustable complex weight. Finally, an integrator unit at the filter output is employed to perform the integration process of "N_P" pulses. Obvious, this leads to a reduction in the overall filter's complexity as well as the total processing time. This time includes the matrix estimation, updating the weights and the integration process. The estimated interference space-time covariance matrix of each layer can be expressed as

$$\hat{M}_{ST}(n) = 1/k_1 \sum_{i=1}^{k_1} \{X_i . X_i^T \qquad \text{for all n} \quad ,1 \le n \le N_L \qquad (23\text{-}2)$$

Where, the observation vector $\underline{X}_j$ has a size ($2N_E$ X 1), and the matrix size is ($2N_E$x$2N_E$) for each adaptive layer. All sub-matrices are estimated at the same time, i.e., the time needed to estimate a single sub-matrix is the same, as that required for estimation of all sub-matrices. The reduction of time is due to having "N_L" processors. Using the estimated covariance matrix of each layer, the complex weights of this layer is updated per each intercepted echo pulse. The noise sample is assumed to be white gaussian noise having a normal distributed. A clutter source is represented by a scatterer, which has a gaussian distribution having certain statistical parameters (μ&σ). Three clutter types are assumed in our simulation including ground, weather, and chaff. The doppler frequency of a moving clutter source is assumed a random variable. In case of weather clutter, F_c varies randomly between [0, .01F_r], while for chaff clutter, "F_c" has a random value varies from zero to 0.1F_r. Also, two parameters, phase and amplitude are used to model the jammer source. The jammer phase "Φ"is assumed to be a random variable uniformly distributed (0,2π). A gaussian distribution is used to model the amplitude fluctuations of the jammer. Finally, the output signal to interference plus noise ratio is estimated as:

$$\hat{SINR}_O = \underline{X}_S \hat{M}_{ST}^{-1} \underline{X}_S \qquad (24\text{-}1)$$

Where, X_s is the signal observation vector.

The estimated $SINR_O$ given by equation (24-1) can be computed in terms of detection index of the filter's output as

$$\hat{SINR}_O = D^2 = [E\{Y_1\} - E\{Y_0\}]^2 . Var\{Y_0\}^{-1} \qquad (24\text{-}2)$$

Where, Y_1 denotes the filter's output when the signal is present, Y_0 denotes the filter's output when the signal is absent, E { } denotes the mathematical expectation operator, and Var { } denotes the variance operator. The output of the filter is $Y_i=W^T X_i$ (i=0 or 1). The variance and expectation operators in equation (24-2) are computed by averaging the filter output for "n" independent observation vectors. In our simulation, the outputs of both filters are averaged for n=150 input observation vectors.

V. RESULTS

Performances of both FASTF and PASTF are evaluated and investigated for different clutter and jammer types. This includes ground clutter, chaff clutter, weather clutter, combined clutter, stand-off jammer (SOJ), escort jammer (EJ), and self-screen jammer (SSJ) sources. Table 2 summarizes the assumed data for our simulation. Results of simulation are presented in Fig. 3 through Fig. 6. As it is clear from Fig. 3, as the total number of array elements increases, the output $SCNR_o$ increases (N_XN_Y=4,16&64). Also, the PASTF almost achieves the same performance as the FASTF except in case of weather clutter (compare Fig. 3c & Fig. 3f). Same behavior has been obtained in case of combined clutter for the FASTF where, the effect of the weather clutter has been denominated on the filter's performance as shown in Fig. 4a. A gain, the PASTF has a poor performance in case of combined clutter as shown in Fig. 4b. This is due to the division of the total space-time covariance matrix of FASTF into "N_L" sub-matrices, each one has a reduced rank of $2N_E$x$2N_E$. Consequently, this leads to a relative phase variation between these sub-matrices of the different layers of the PASTF. This phase variation denominates in case of weather clutter, which has a low random phase fluctuation [0, π/50]. Thus, the PASTF should have an optimum number of adaptive layers for the constant product "N_EN_L" (the total number of array elements) to compensate the effect of the phase variation between the sub-matrices of the different layers. Results of optimization in case of combined clutter are shown in Fig. 5a. As it is clear from the figure, for total number of 64-elements, maximum $SCNR_o$ has been achieved with two layers each one has thirty-two elements. However, to achieve good angle resolution (elevation), the number of adaptive layers should be greater than two layers (N_E>2). Therefore, a compromise between the required angle resolution and the maximum $SCNR_o$ should be done to optimize the filter design. Fig. 5b shows the performance comparison between the two filters, where, the PASTF has four adaptive layers each one of sixteen elements. Table 3 illustrates the performance difference between FASTF and PASTF. In this case, a 10-dB difference between the achieved $SCNR_o$ (averaged over the normalized doppler frequency) of the two filters has been observed.

Fig. 6 illustrates the performance comparison between the two filters in case of different jammer types. As it is clear from the figure, both filters have been achieved the same output $SCNR_o$. Also, maximum $SCNR_o$ has been achieved in case of SOJ as compared to EJ and SSJ for both filters. This is due to the fact that SOJ is a side-lobe jammer source while the others are main-lobe jammers. In contrast to the weather clutter case, the random phase fluctuation of the jammer source [0, 2π] dominates over the relative phase variation between the sub-matrices of the different layers. Therefore, for any combination N_E and N_L of constant product, the output $SCNR_o$ of the PASTF is almost the same (within 3-dB variation).

A practical case study of a combined interference (clutter and jammer) in case of search "S" and tracking "T" radar system has been assumed as summarized in Table 4. The other simulation parameters are the same as given in Table 2. Results of simulation are presented in Fig. 7 and Fig. 8. On average, almost, the same performance has been achieved in case of search radar for both filters as shown in Fig. 7. On the other hand, in case of tracking radar, PASTF has a poor performance under SSJ plus combined clutter condition. This can be overcome by increasing the filter size, i.e., N_E and/or N_L. In this case, the number of array elements per each layer is increased while the number of layer is kept constant, and vices versa.

VI. CONCLUSIONS

A new-partially adaptive space-time filter has been presented, analyzed, and investigated versus a full adaptive one. A computer code has been developed on MATLAB-R12 to model the operation processes of both filters. This also includes modeling of the target echo and its associated interference signals. The obtained results indicate that the partial adaptive filter has almost achieved the same performance as the full adaptive filter under the same interference conditions. In addition, the new- presented partially adaptive filter offers a good solution to the main problem challenges the designers of adaptive phased array radar systems. This includes a compromising between the hardware complexity, processing time, cost and the required system performance. In fact, performance of partially adaptive filter ,or system, versus full adaptive filter is a highly dependent function of cost, complexity, and processing time. Thus, performance optimization is mainly application-related problem. Finally, our partially adaptive filter provides an excellent solution to the problem of either detection or tracking of the radar echo immersed in an interference background.

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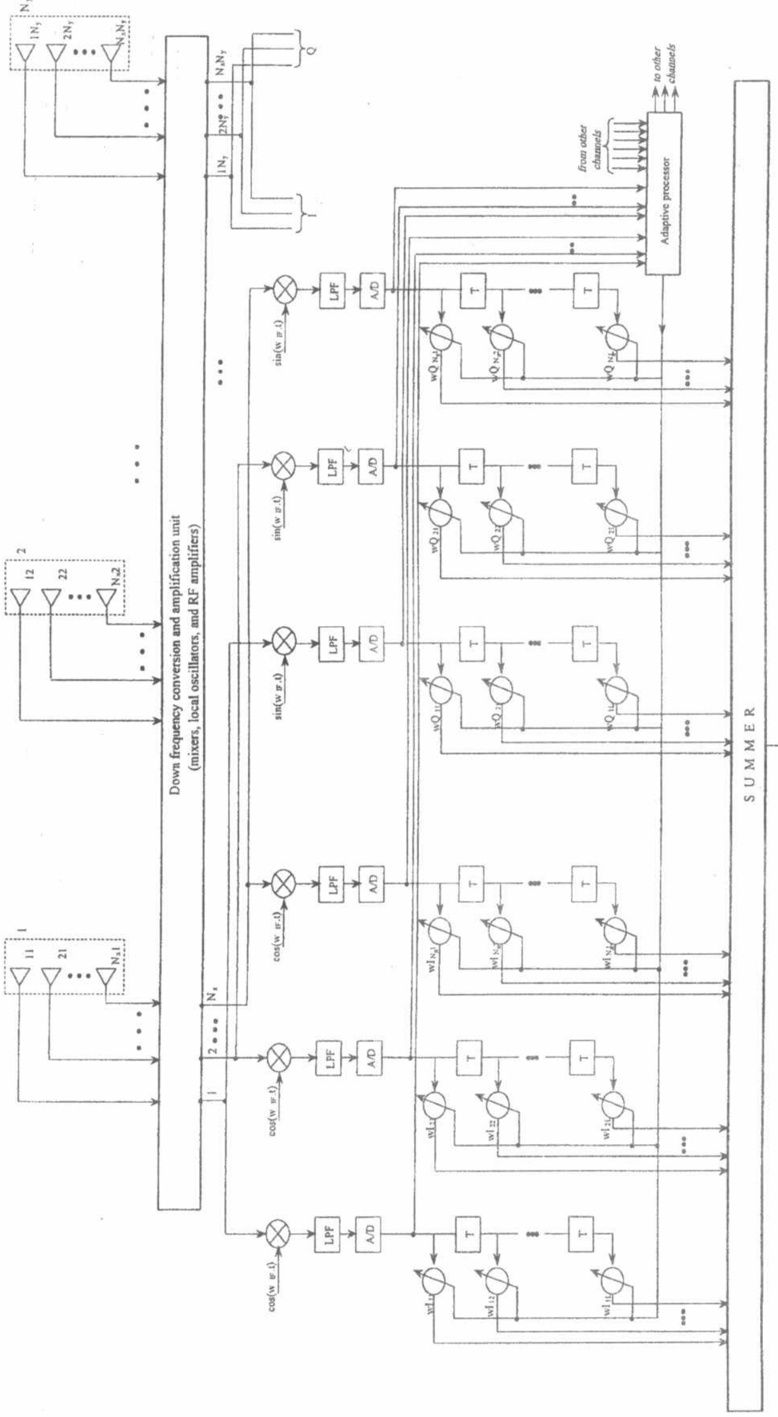


Fig. 1-a A full adaptive space-time filter (FASTF) using planner array configuration

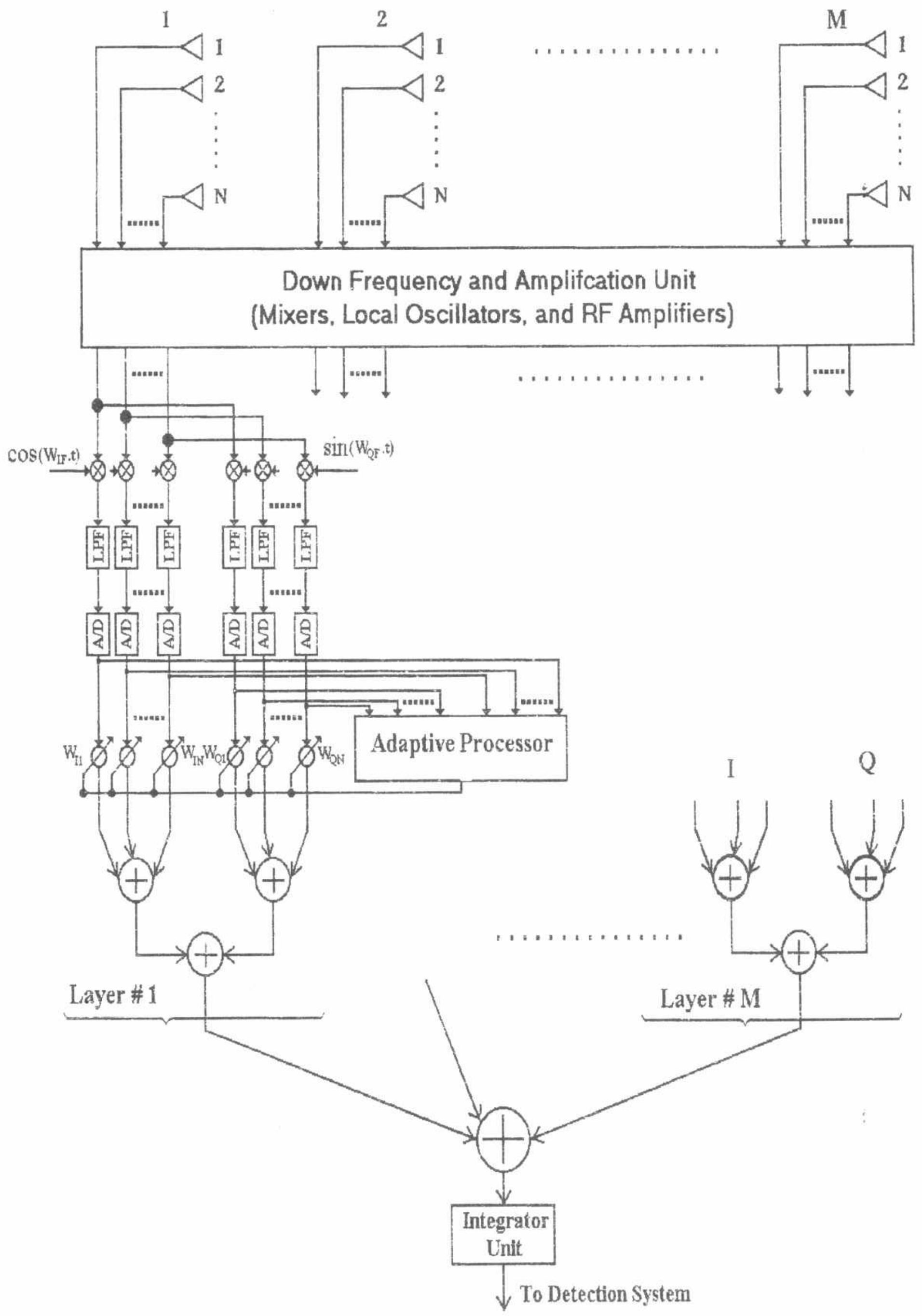


Fig. 1-b A new partial adaptive space-time filter using planner array configuration

Table (1): Summary of the main differences between the two Space-Time filters.

	FASTF	PASTF
FIR	With length N_P	Single complex weight
IU	-	For integration of N_P pulses
OBV	($2N_x * N_y * N_p$ x 1)	($2N_E * N_L$ x 1)
AP	One adaptive processor	N_L adaptive processors
Matrix size	Single matrix of size ($2N_x * N_y * N_p$) x ($2N_x * N_y * N_p$)	N_L matrices each of size ($2N_E$) x ($2N_E$)

Where,

OBV	Observation vector size.
FIR	Finite Impulse Response Filter of size N_P.
I U	Integrator unit.
A P	Adaptive processor.
N_p	Number of snapshot.
N_L	Number adaptive of layer (N_L=N_y).
N_E	Number of elements per one channel (N_E =N_x).

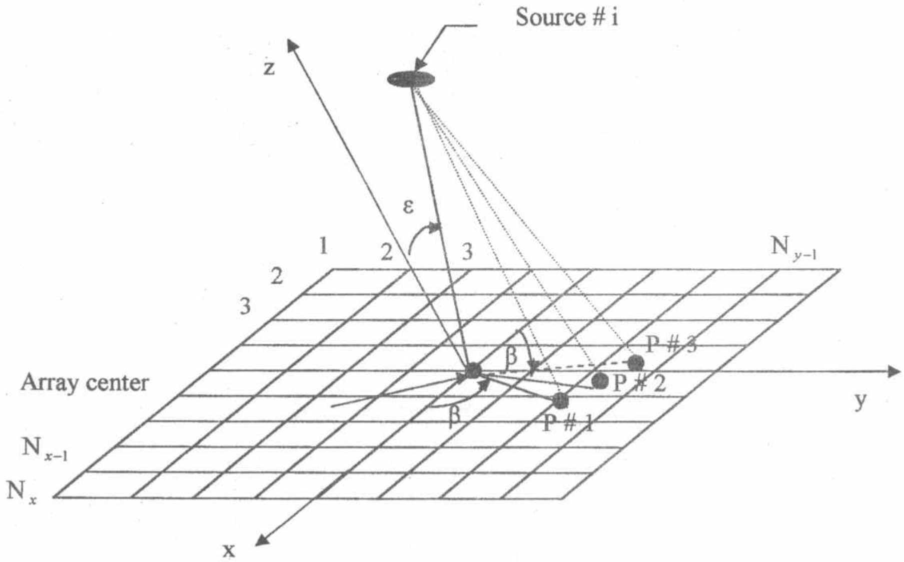


Fig. 2 Element geometry of a two – dimensional array.

Table 2 Simulation data for performance evaluation of the two proposed filters.

Simulation Parameters	Clutter type			Jammer		
	Ground	Weather	Chaff	SOJ	EJ	SSJ
N_c / N_j	1	1	1	1	1	1
N_p	3	3	3	3	3	3
SNR_i (dB)	10	10	10	10	10	10
μ^2 (dB)	40	15	10	-	-	-
σ^2(dB)	20	25	30	30	20	20
ε (deg.)	35	30	15	30	35	36
β (deg.)	80	85	88	80	80	89
F_{dc} / F_{dj}	0.0	Rand (0 - .01F_r)	Rand (0 - .1F_r)	0.0	0.6F_r	0.6F_r

Table 3 Performance comparison between the two filters.

Simulation parameters	FASTF	PASTF
N_XN_Y	64	64
Number of adaptive weights (NOAW)	384 (real)	128 (real)
N_P	3	3
$N_x = N_E$	8	16
$N_y = N_L$	8	4
Matrix size	(384 x 384)	(32 x 32)
$SINR_o$ in dB	32	22

Table 4 Data of simulation for 2D–AST filters for search and tracking radars.

Simulation Parameters	**Clutter type**			Jammer type		
	Ground	Weather	Chaff	SOJ	EJ	SSJ
ε (deg.)	(10, 38, 70) (S) (10, 55, 65) (T)	(25, 38, 55) (S) (30, 38, 50) (T)	(20, 37, 50) (S) (35, 37, 50) (T)	15	35	38
β (deg.)	(40, 85, 120) (S) (20, 45, 115) (T)	(80,89,100) (S) (60,85,110) (T)	(60,85,110) (S) (60,80,100) (T)	30	80	88
N_c / N_j	3	3	3	1	1	1

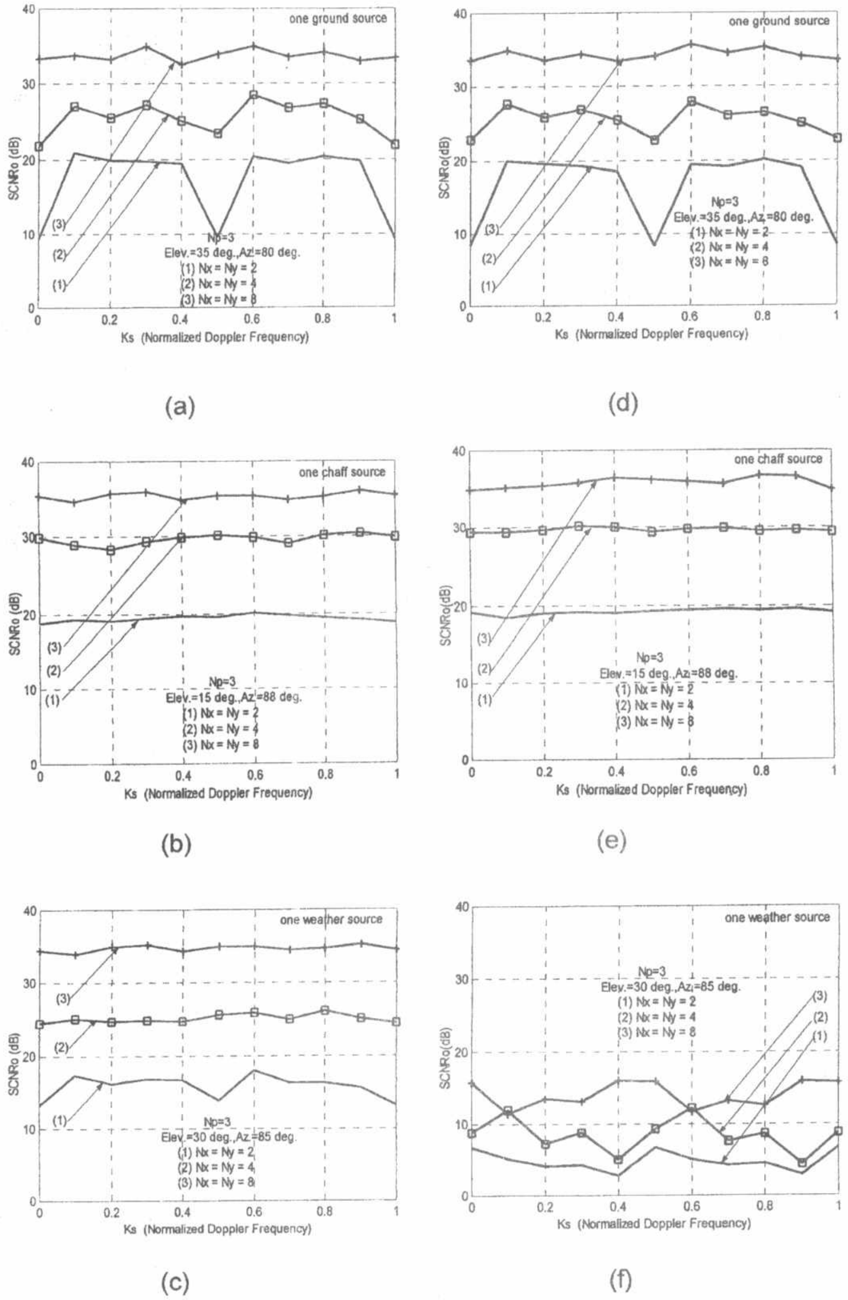
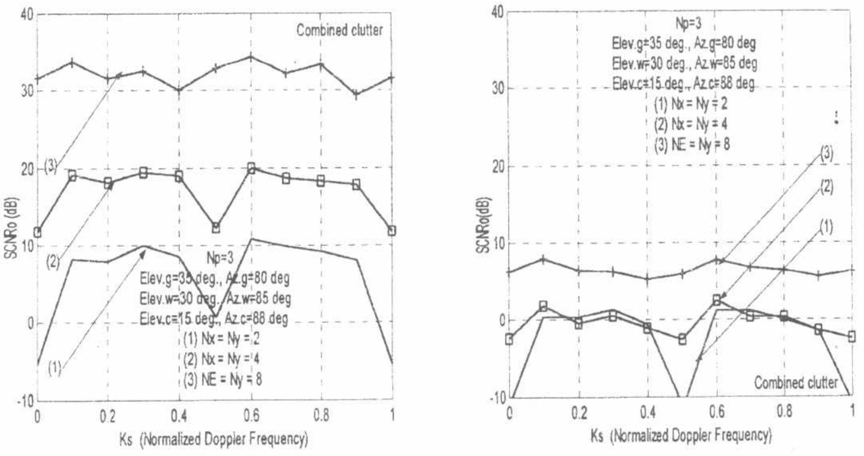
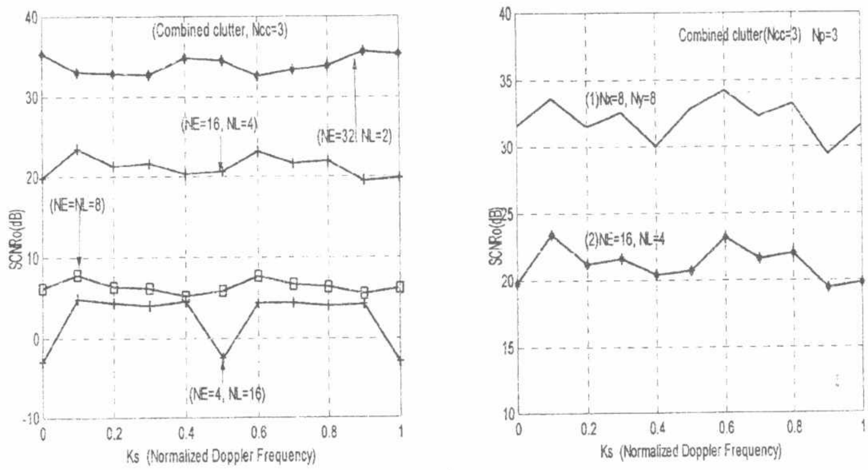


Fig.3 $SCNR_o$ achieved by the FASTF and PASTF versus K_S. (a), (b) & (c) for FASTF and (d), (e) & (f) for PASTF.



(a) (b)

Fig. 4 $SCNR_o$ achieved by FASTF & PASTF versus K_S for combined clutter consists of three clutter sources (one G_c + one W_c source + one Ch. source). 5(a) FASTF & (b) PASTF.



(a) (b)

Fig. 5 (a) $SCNR_o$ achieved by the PASTF versus K_S, for different element number, N_E & different layer number, N_L.
(b) SCNRo achieved by the two proposed filters ((1) FASTF & (2) PASTF) versus K_S for combined clutter consists of three clutter sources (one G_c + one W_c source + one Ch. source).

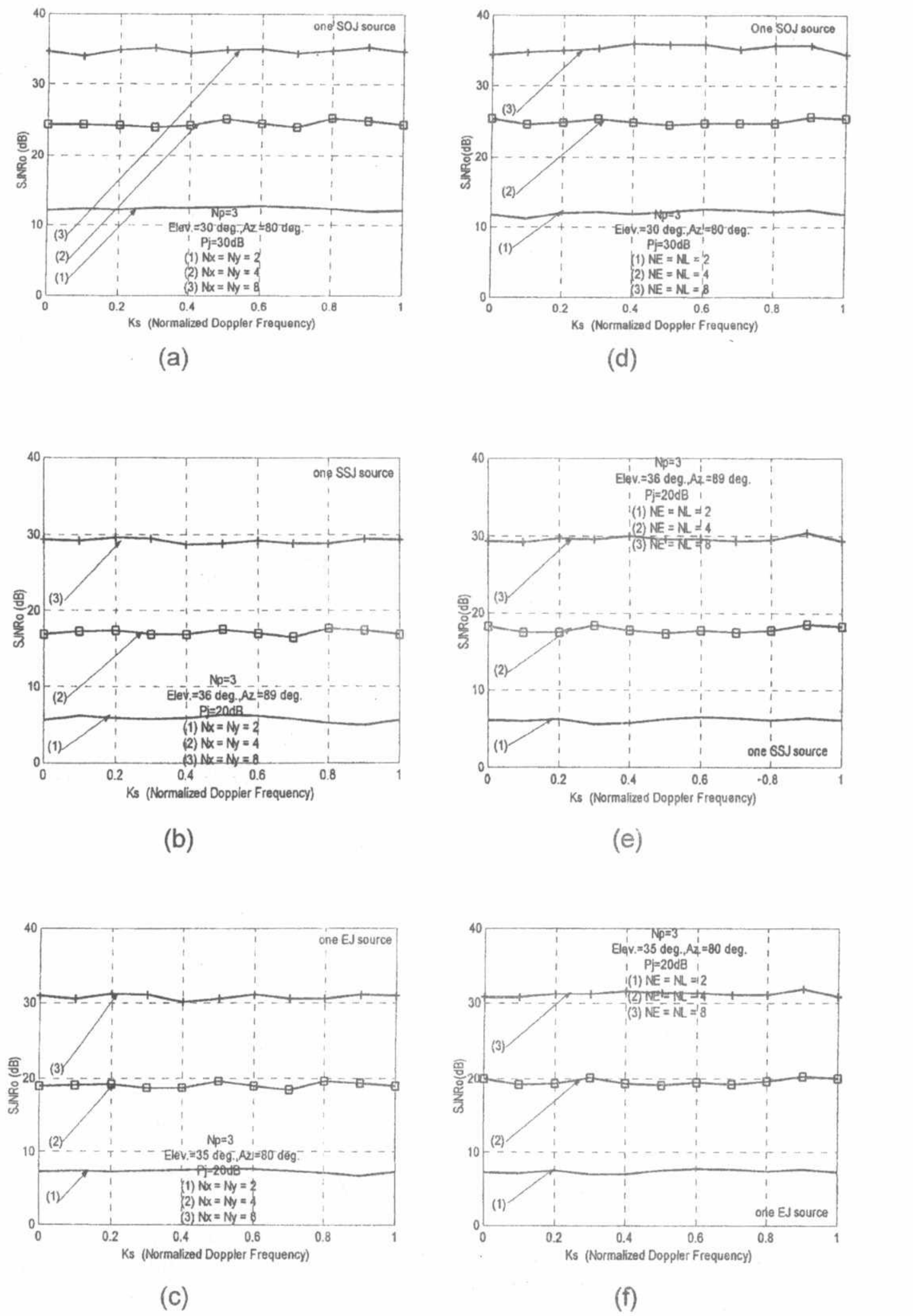
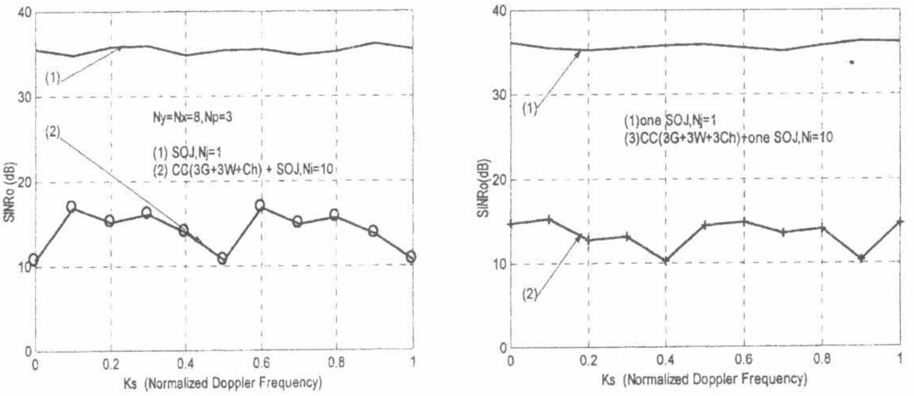
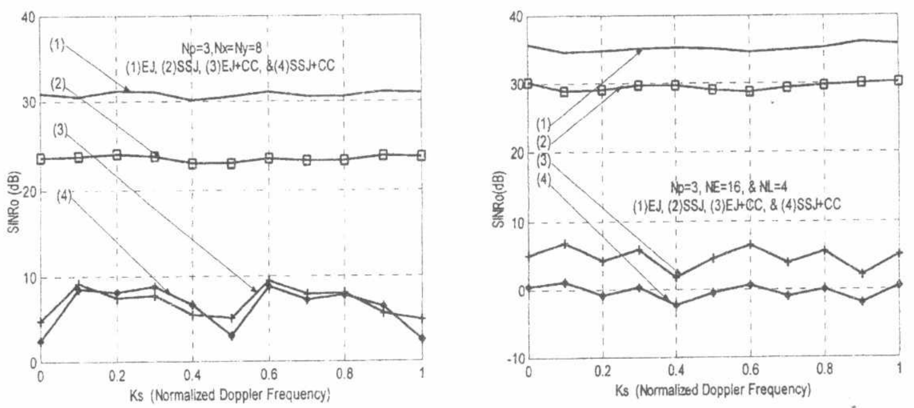


Fig. 6 $SJNR_o$ achieved by the FASTF and PASTF versus K_S. (a), (b) & (c) for FASTF and (d), (e) & (f) for PASTF.



(a) (b)

Fig.7 SINR0 versus Ks in case of search radar system. (1) One SOJ. (2) Combined interference (SOJ + 3 ground clutter sources + 3 weather clutter sources + 3 chaff clutter sources). (a) FASTF. (b) PASTF.



(a) (b)

Fig.8 SINR0 versus Ks in case of tracking radar system. (1) One EJ. (2) One SSJ. (3) Combined interference (EJ + 3 ground clutter sources + 3 weather clutter sources + 3 chaff clutter sources). (4) Combined interference (SSJ + 3 ground clutter sources + 3 weather clutter sources + 3 chaff clutter sources). (a) FASTF. (b) PASTF.