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ANALYZING PLANAR ANTENNA ARRAYS FOR PERSPECTIVE AUTOMATED ELECTRONIC WARFARE STATIONS OPERATING AGAINST ON-BOARD RADAR SYSTEMS OF AIR ASSAULT MEANS

We claim that regarding electronic warfare systems there exists the problem of simultaneous providing large bandwidth, required directivity, quality of spatial signal filtering in interference background and resolution in angular coordinates. It is shown that solving this problem in conditions when computational resources are constrained is possible on the basis of exploiting antenna arrays in the form of Mills cross with Vivaldi elements and digital signal processing, at the same time providing respectively small weight and overall dimensions of antenna system. We develop the field pattern model of planar antenna array in the form of Mills cross that allow substantiating main technical requirements for both receiving and transmitting antenna systems of perspective electronic warfare systems operating against on-board radars of air assault means. We develop the simulation model of spatial signal filtering under interference influence for planar antenna array in the form of Mills cross with electronic control of antenna pattern and digital signal processing, allowing us determining characteristics of signal processing under interference conditions.

Keywords: *planar antenna array, antenna array in the form of Mills cross, antenna array with digital signal processing, antenna pattern, directivity, bandwidth coefficient, Vivaldi element, radar system, electronic warfare against air assault means.*

INTRODUCTION

Warding off armed aggression of russian federation and also the experience of local wars and armed conflicts of last decades earnestly testify to decisive role of air assault means (AAM) in reaching objectives of military operations and armed conflict as a whole. Main types of both tactical aviation planes and cruise missiles of armed forces of russian federation exploit modern radar on-board weapon control systems, surveillance, low height flight safety insurance and navigation systems.

Actuality of developing perspective automated jamming station for electronic warfare (EW) against on-board radar systems is due to the necessity of providing air defence of corps and objects in military operations of Armed Forces of Ukraine, warding off both tactical aviation and cruise missiles strikes and also decreasing the efficiency of enemy air surveillance by means of jamming on-board radar systems of tactical aviation and cruise missiles.

As domestic and foreign analogues of perspective automated EW station operating against on-board radar systems of AAM we consider such jamming stations: SPN-40 (USSR, 1973), SPN-30 (USSR, 1975), SPN-2 (USSR, 1988), SPN-4 (USSR, 1989), Krasuha-2, Krasuha-4 (RF, 2012), Koral (Turkey, Aselsan, 2013), the appearance of some of them is shown in the Fig. 1. Aperture antennas (horn, reflector parabolic) are used as both receiving and transmitting antennas of automated jamming stations (AJS) SPN-40, SPN-30, Krasuha-2, Krasuha-4, whereas AJS SPN-2, SPN-4, Koral exploit antenna arrays.

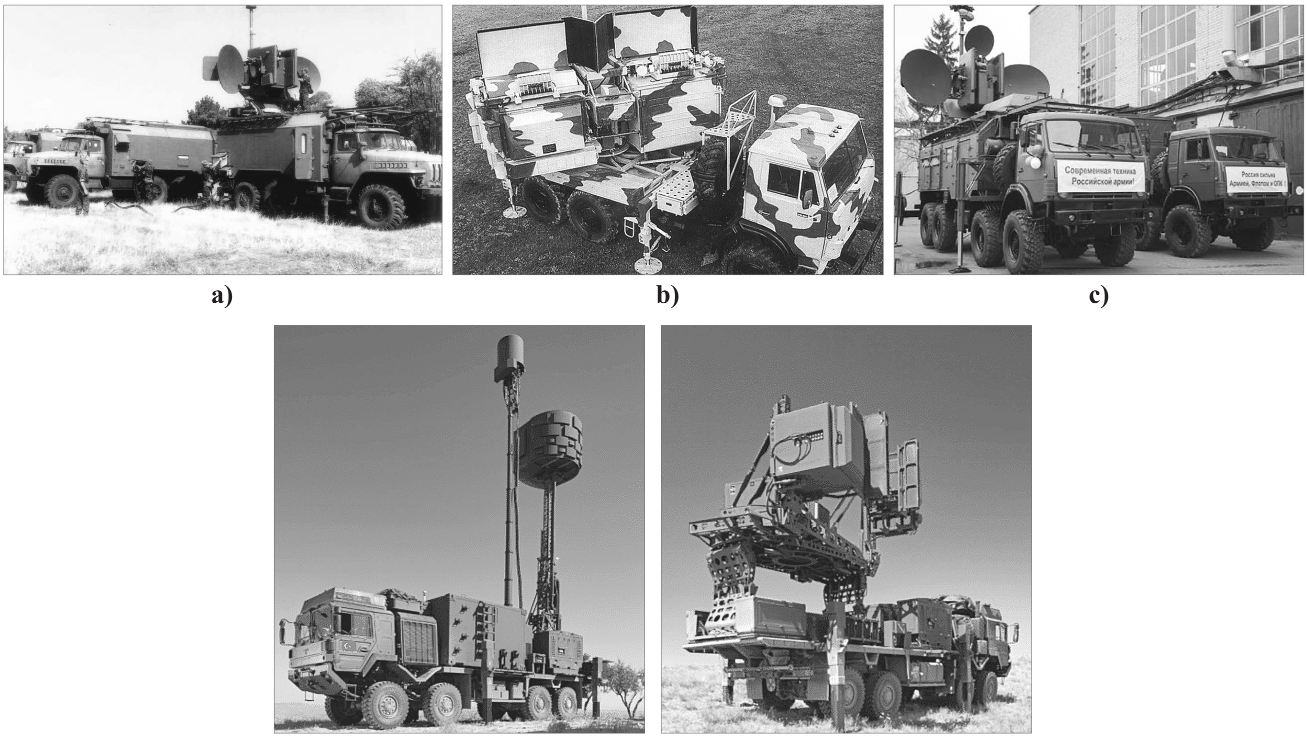
There are a lot of requirements that are raised to antennas of modern electronic systems; among them three requirements have decisive importance. The first one is directivity, i.e. distribution of electromagnetic power in space (or response on arrived electromagnetic field when receiving the signals) according to some dependence. In some cases it is desired to provide omni-directional property of antenna, in others it is necessary to concentrate radiation or to receive the signals within rather narrow angle sector, i.e. main lobe of a field pattern of antenna system. The second and the third requirements mean that either transmitting or receiving the signals must provide maximal antenna efficiency, i.e. antenna has possess first, low voltage standing wave ratio (VSWR) and second, antenna radiation efficiency. As for EW systems, unlike other electronic systems, providing large respective bandwidth (BW) of frequency band (BW coefficient, i.e. the ratio maximal frequency of frequency band to minimal one) has an important significance, so that BW coefficient takes values from 3:1 and greater.

Known EW stations are usually supplied with log-periodic, spiral, biconical, discone and horn antennas providing BW coefficients that take values 10:1, 4:1, 4:1, 3:1 and 2:1, respectively. However, simultaneous providing contemporary requirements with respect to both wide frequency band, large directivity and respectively small weight and overall dimensions of antenna system is considered to be rather problematic.

The possible way of this problem solution is exploiting antenna arrays (including antenna arrays with digital signal processing) based on other technological principles.

BIBLIOGRAPHY ANALYSIS

Wideband antenna arrays based on Vivaldi elements are considered in works [1–7]. The approach to constructing wideband antennas developed on the basis of exploiting controlled coupled elements and also antenna arrays based on current sheet are considered in works [8–10]. Wideband antenna arrays with tightly coupled elements are considered in works [9–11]. The works [12–14, 16] investigate the



F i g . 1. AJS operating against on-board radar systems: a) SPN-30 (USSR); b) SPN-2 (USSR); c) Krasuha-4 (RF); d) Koral (Turkey, Aselsan)

problems of developing planar antenna arrays in the form of Mills cross, that allows sufficiently decreasing overall number of antenna elements with simultaneous preserving the required resolution in angular coordinates.

The objective of the article is first, developing the field pattern model of planar antenna array in the form of Mills cross on whose basis it is necessary to substantiate main technical requirements with respect to both receiving and transmitting antenna systems for perspective automated EW stations operating against on-board radar systems of AAM. **The second objective of the article** is developing the simulation model of spatial signal filtering under interference conditions for planar antenna array in the form of Mills cross with electronic control of antenna pattern and digital signal processing that allow determining an efficiency of signal processing in the presence of jam influence.

RESEARCH RESULTS

Vivaldi elements for wideband antenna arrays

Modern constructions of antenna arrays based on Vivaldi elements can provide VSWR lesser than 2 for relative frequency bandwidth (BW coefficient) 10:1 and even greater [1–7]. Almost ideal forms of antenna array pattern are provided at scanning angles (with respect to aperture normal) $\pm 50\ldots60°$. Antenna arrays based on Vivaldi elements designed and produced commercially do not require expensive materials to provide the aforementioned characteristics. Furthermore, antenna arrays based on Vivaldi elements contain in their structure the matching devices (transformers) that makes them compatible with widely used transmitting lines, e.g. microstrip lines.

Another name of antenna arrays based on Vivaldi elements is tapered slot antenna (TSA). Vivaldi name takes its origin from exponentially-slotted antenna that was first proposed in the work [4]. The development of TSA was being accomplished rather slowly up 1990-s, when numerical analysis methods realized by computers allow exactly describe characteristics of antenna arrays with a large number of elements. Just after that it was understood that successful design of wideband antenna arrays with electronic scanning is mainly based on modelling and inter-elements couple control.

Antenna arrays with electronic scanning and large aperture intended for simultaneous operating either in some wide frequency band or in some distant rather small frequency intervals must meet rough requirements with respect to main antenna characteristics that can be provided by antenna arrays based on Vivaldi elements. Arrays with single or dual polarization can be developed on the basis of TSA and operate within established characteristics in far-field zone.

To provide wide scanning sector without extra diffraction maximums of antenna array pattern in the form of relatively large side lobes, inter-element distance has not $0.5\lambda_{HF}$, where λ_{HF} is the wavelength at the highest frequency (HF) of the frequency band [1]. It is also known that diffraction theory recommends decreasing inter-element distance as far as possible to improve the possibilities of antenna pattern main lobe scanning.

The weight of TSA is mainly determined by the weight of dielectric material by which is filled the space between the elements. Maximum length of a typical element of TSA that operates in three octaves and more is equal to $8\times0.45\lambda_{HF}$. Hence one can calculate volume and weight of antenna array.

Fig. 2 illustrates the parameters defining TSA. All shown geometrical parameters affect the antenna efficiency but some common considerations can be used for improving antenna properties: 1) inter-element distances *a, b* along two coordinates *x, y* have to be $0.45\lambda_{HF}$; 2) usually the longest elements operate better at the lowest frequencies providing the wider frequency band; 3) thickness of dielectric substrate which is 0.1 of element width provides good results; 4) exponential parameter R_a and resonator diameter D_{sl} sufficiently affect the antenna properties.

Planar antenna arrays in the form of Mills cross

Planar antenna arrays in the form of Mills cross is a union of the elements of two linear antenna arrays arranged in two mutually orthogonal axes Ox, Oy [14]. The resulting process $y(t)$ in the output of antenna array in the form of Mills cross (Fig. 3a) is determined by the following expression:

$$y(t)=\sum_{n=1}^{N}\overline{w}_n^{x}a_n^{x}(t)+\sum_{m=1}^{M}\overline{w}_m^{y}a_m^{y}(t), \qquad (1)$$

where $a_i^x(t)$, $a_k^y(t)$ are the received signals in the inputs of *i*-th / *k*-th receiving element of linear antenna array (orthogonal component of planar array), whose elements are allocated along the axes Ox, Oy respectively; w_i^x, w_k^y are weight coefficients in *i*-th / *k*-th processing channel of antenna array, whose elements are allocated along the axes Ox, Oy respectively; $\overline{w}_i=\mathrm{Re}(w_i)-j\,\mathrm{Im}(w_i)$; *N, M* are the numbers of antenna array along the axes Ox, Oy respectively.

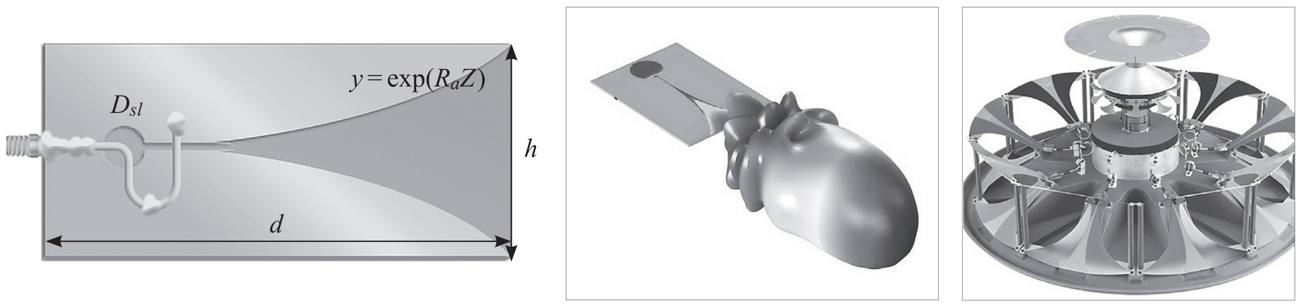
A variant of joint allocation of 8…12 and 12…18 GHz band antenna arrays on the load-bearing slab for perspective AJS operating against on-board radar systems of AAM is shown in Fig. 3b.

Field pattern (FP) $F(\varphi,\theta)$ (or array factor) of planar antenna array is determined by the following equations [15; (2.45), (2.46)]:

$$F(\varphi,\theta)=F_0(\varphi,\theta)\sum_{m=1}^{M}\sum_{n=1}^{N}w_{mn}\cdot\exp\left[j2\pi\left\{\begin{aligned}&(n-1)\frac{d_x}{\lambda}\\&(u-u_s)+\\&(m-1)\frac{d_y}{\lambda}\\&(v-v_s)\end{aligned}\right\}\right], \qquad (2)$$

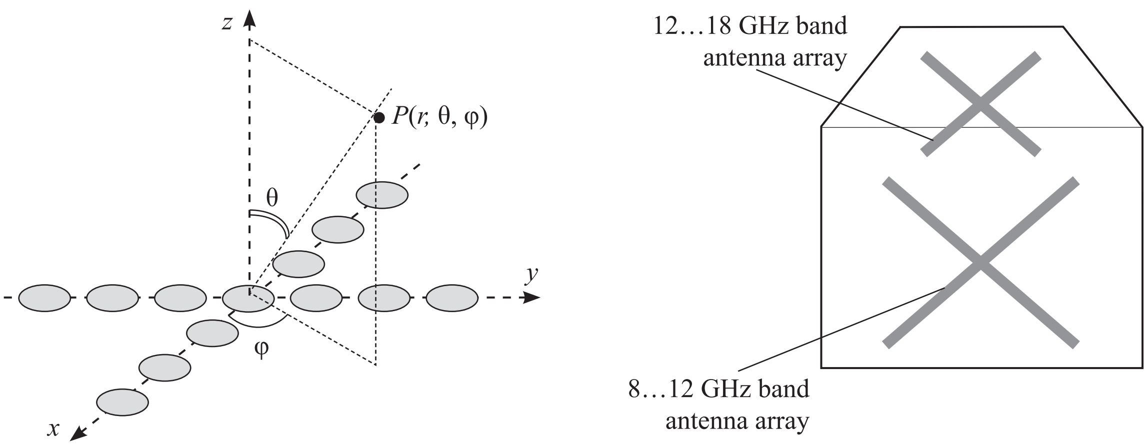
$$\begin{cases}u=\sin\theta\cos\varphi;\\ u_s=\sin\theta_s\cos\varphi_s,\end{cases}\quad \begin{cases}v=\sin\theta\sin\varphi;\\ v_s=\sin\theta_s\sin\varphi_s,\end{cases} \qquad (3)$$

where *N, M* are numbers of the antenna array elements along the coordinate axes Ox, Oy respectively; d_x, d_y are inter-element distances along the coordinate axes Ox, Oy respectively; λ is a working wavelength; $F_0(\varphi,\theta)$ is a FP of a single Vivaldi element; φ_s, θ_s are FP main lobe angular coordinates controlled electronically; w_{mn} are complex



a) b) c)

F i g . 2. Parameters of TSA: a) constructive parameters of the element; b) Vivaldi element field pattern; c) TSA of electronic support measures station TCI 903



a) b)

F i g . 3. Antenna arrays in the form of Mills cross for perspective AJS operating against on-board radar systems: a) common configuration within 3-D coordinate system; b) joint allocation of 8…12 and 12…18 GHz band antenna arrays on the load-bearing slab weight coefficients used when realizing digital signal processing in antenna array.

Thus, from the expressions (1), (2), we obtain FP $F(\varphi,\theta)$ of antenna array in the form of Mills cross:

$$F(\varphi,\theta) = F_x(\varphi,\theta) + F_y(\varphi,\theta); \tag{4}$$

$$F_x(\varphi,\theta) = F_0(\varphi,\theta)\cdot\sum_{n=0}^{N-1} w_n \cdot \exp\left[\begin{array}{c} j2\pi\left(n-\frac{N-1}{2}\right) \\ \frac{d_x}{\lambda}(u-u_s) \end{array}\right]; \tag{5}$$

$$F_y(\varphi,\theta) = F_0(\varphi,\theta)\cdot\sum_{m=0}^{M-1} w_m \cdot \exp\left[\begin{array}{c} j2\pi\left(m-\frac{M-1}{2}\right) \\ \frac{d_y}{\lambda}(v-v_s) \end{array}\right], \tag{6}$$

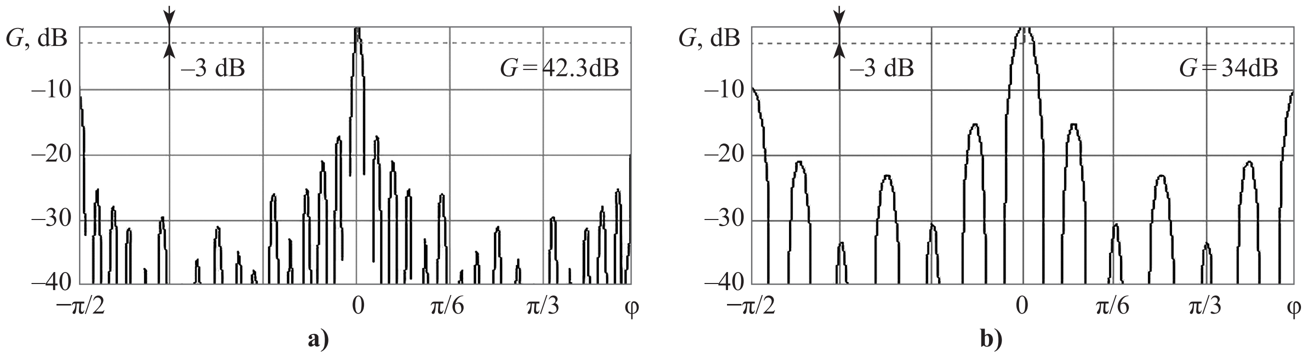
where the variables that are present in formulas (5), (6) have the same matter they have in the expressions (2), (3).

Fig. 4a, b illustrate the intersections of normalized FP of antenna arrays in the form of Mills cross with the numbers of elements that are equal to $2\cdot 65-1$ and $2\cdot 25-1$ respectively. These dependences were obtained on the basis of relationships (4)–(6) for azimuth coordinate plane. The width of FP $F_0(\varphi,\theta)$ of a single array element takes the value 120° for both azimuth and elevation planes measured at the level –3 dB with respect to maximum. The described variants of antenna arrays provide antenna gains 42.3 dB and 34 dB respectively. Inter-element distance takes the value $0.5\lambda_{HF}$, where λ_{HF} is a wavelength at the highest frequency *HF* of frequency band (*HF* = 18 GHz and 12 GHz for the subbands 12…18 and 8…12 GHz respectively), so that main lobe width of FP takes the values 1.6° and 6° respectively measured at the level –3 dB with respect to maximum. It should be noted that at the lowest frequency *LF* of frequency band (*LF* = 12 GHz and 8 GHz for subbands 12…18 and 8…12 GHz respectively) one can notice widening main lobe FP up to the values 3° and 7.6° respectively at –3 dB with respect to maximum.

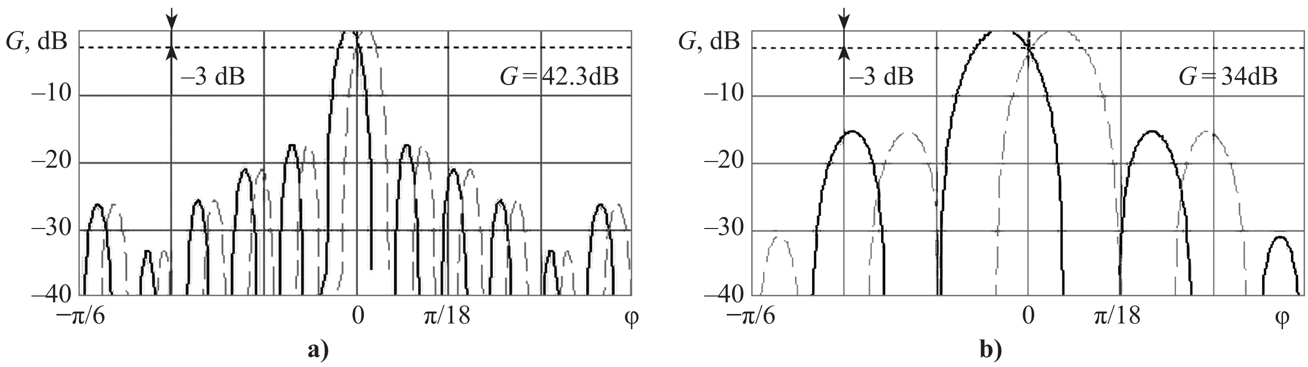
One of the main function of automatic target tracking subsystem of AJS is forming an error signal by exploiting the method of monopulse direction finding radiation emitter, i. e. on-board radar system of AAM. Fig. 5a, b illustrate the intersections of normalized FP of receiving antenna arrays in the form of Mills cross with the numbers of elements that are equal to $2\cdot 65-1$ and $2\cdot 25-1$ respectively while creating monopulse direction finding channels. These dependences were obtained on the basis of relationships (4)–(6) for azimuth coordinate plane. Partial FPs (conditionally left and right) are shown by the solid and dashed lines respectively, so that direction finding characteristic width takes the values 1.6° and 6° respectively at –3 dB with respect to maximum.

Electronically controlled main lobe of antenna array in the form of Mills cross

Here as both receiving and transmitting arrays we suppose exploiting antenna arrays with digital signal processing. In the case of mechanical method of FP main lobe scanning, the key parameters (antenna gain and FP width) provided by the given number of the elements ($2\cdot 65-1$ and $2\cdot 25-1$ respectively) remain quite satisfactory within the whole frequency band of AJS (8…18 GHz) (Fig. 4, 5). In the case



F i g . 4. Intersections of normalized FP of antenna arrays in the form of Mills cross: a) $2\cdot 65-1$ elements antenna array; b) $2\cdot 25-1$ elements antenna array



F i g . 5. Intersections of normalized FP of receiving antenna arrays in the form of Mills cross while creating monopulse direction finding channels: a) $2\cdot 65-1$ elements antenna array; b) $2\cdot 25-1$ elements antenna array of electronic method of FP main lobe scanning, situation of providing these key parameters become worse. This fact is due to widening antenna array FP width while using frequencies that are close to LF of frequency subband (LF = 12 GHz and 8 GHz for subbands 12…18 and 8…12 GHz respectively).

Fig. 6 illustrate the intersections of normalized FP of antenna arrays in the form of Mills cross with the numbers of elements that are equal to $2 \cdot 65 - 1$ with electronic scanning the space in two coordinate planes. The positions of FP main lobe in both azimuth and elevation planes are determined by coordinates $\varphi_s = -\pi/3$; $\theta_s = \pi/4$ respectively and shown by the dashed line. The initial position of antenna array FP corresponding to coordinates $\varphi_s = 0$; $\theta_s = 0$ is shown by the solid line. In the Fig. 6a one can see that electronic scanning is accompanied by appearing parasitic mirror FP lobe which position is determined by coordinates $\varphi\_ = \pi/3$; $\theta\_ = \pi/4$, that negatively affect operating of both receiving and transmitting antenna arrays. In order to eliminate such a negative phenomenon it is necessary exploiting some measures, namely, creating «zero» FP at the direction of parasitic mirror FP lobe with $\varphi\_ = -\varphi_s$ coordinate.

Fig. 7 illustrate the intersections of normalized FP of antenna arrays in the form of Mills cross with the numbers of elements that are equal to $2 \cdot 35 - 1$ with electronic scanning the space in two coordinate planes for the conditions that are similar to ones shown in Fig. 6.

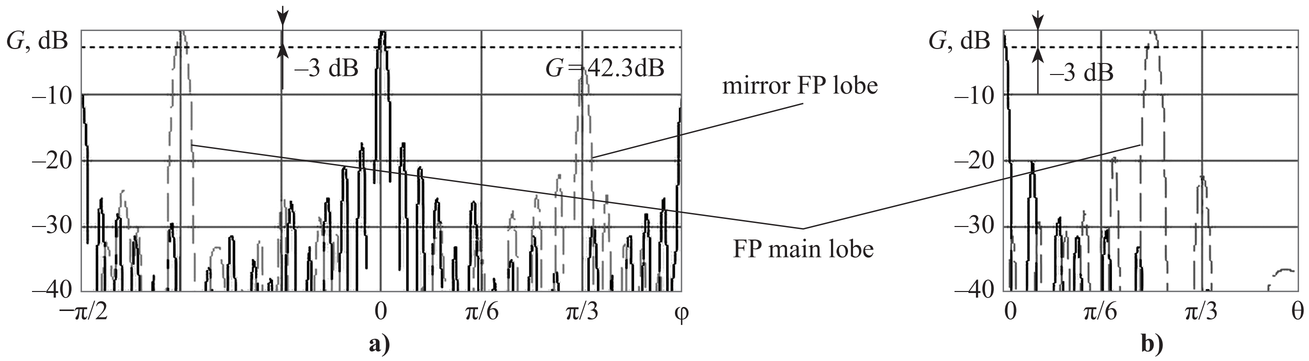
It should be noted that situations concerning FP main lobe electronic scanning shown in Figs. 6, 7 correspond to such conditions: FP width $F_0(\varphi,\theta)$ (see formulas (5), (6)) of a single Vivaldi element takes the value 120° in both azimuth and elevation plane measured at the level –3 dB with respect to maximum. The described variants of antenna arrays provide antenna gain 42.3 dB and 36.9 dB respectively. Inter-element distance takes the value $0.5\lambda_{HF}$, where λ_{HF} is a wavelength at the highest frequency HF of frequency subband (HF = 18 GHz and 12 GHz for the subbands 12…18 and 8…12 GHz respectively), so that FP main lobe width takes the values 1.6° and 4° respectively measured at the level –3 dB with respect to maximum. It should be noted that at the lowest frequency LF of a subband (LF = 12 GHz and 8 GHz for the subbands 12…18 and 8…12 GHz respectively) one can notice widening main lobe FP up to the values 5.1° and 9.8° respectively at the level –3 dB with respect to maximum.

Eliminating parasitic mirror FP lobe in the case when main lobe position is determined by coordinates $\varphi_s = \pi/3$; $\theta_s = \pi/4$, shown in Figs. 6a and 7a for antenna arrays with $2 \cdot 65 - 1$ and $2 \cdot 35 - 1$ elements, by creating «zero» FP at the direction of parasitic mirror FP lobe with coordinate $-\varphi_s$ is realized on the basis of the known relationship [15; (2.45), (2.46)]:

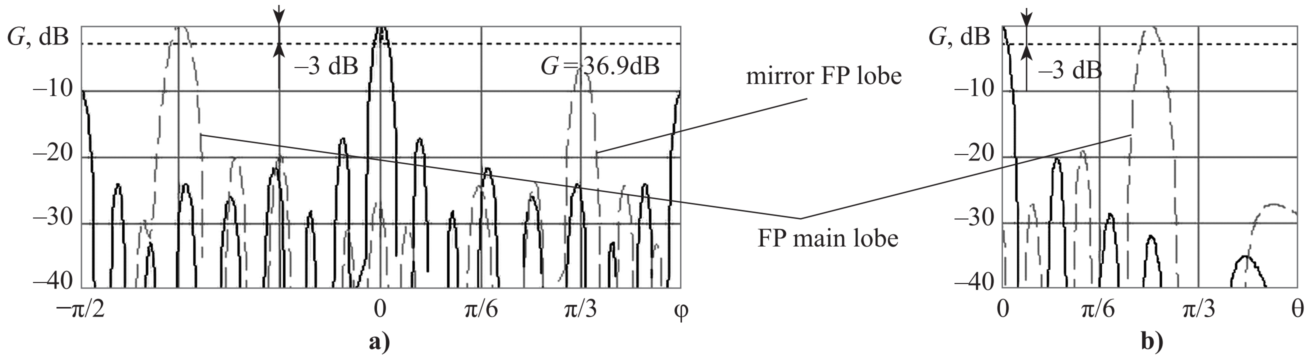
$$F_c(\varphi,\theta) = F(\varphi,\theta)\big|_{\varphi_s} - \gamma \cdot F(\varphi - \varphi\_,\theta)\big|_{\varphi_s = 0}, \qquad (7)$$

where $F_c(\varphi,\theta)$ is corrected FP of antenna array; $F(\varphi,\theta)\big|_{\varphi_s=0}$ is initial FP (without steering angles: $\varphi_s = 0$; $\theta_s = 0$); $F(\varphi,\theta)\big|_{\varphi_s}$ is FP as $\varphi_s \neq 0$ without compensation of receiving in the direction of parasitic mirror FP lobe (see Fig. 6a and 7a); γ is the level of parasitic mirror FP lobe; $\varphi\_$ is the direction of parasitic mirror FP lobe: $\varphi\_ = -\varphi_s$.

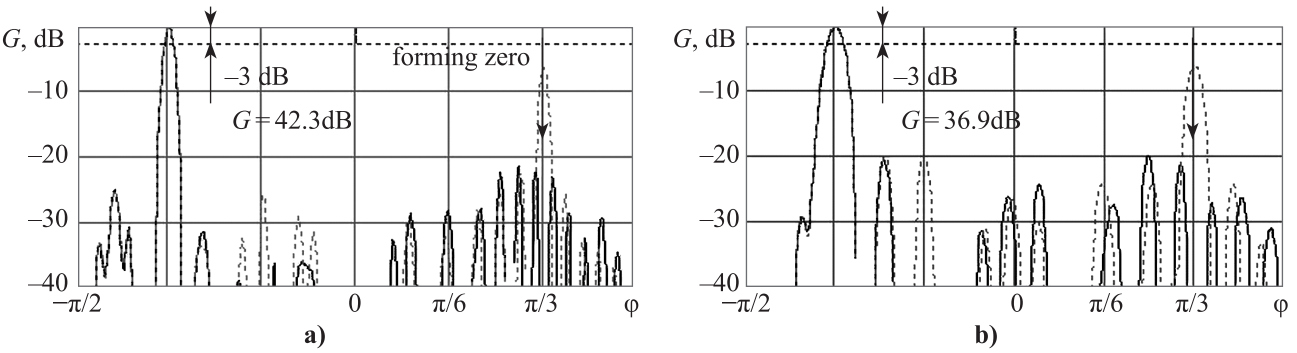
Fig. 8a,b illustrate the resulting corrected FPs $F_c(\varphi,\theta)$ (see formula (7)) for antenna arrays in the form of Mills cross with electronic scanning in azimuth plane (FP main lobe $\varphi_s = -\pi/3$) and suppressing parasitic mirror FP lobe



F i g . 6. Intersections of normalized FP of antenna arrays in the form of Mills cross ($2 \cdot 65 - 1$ elements) with electronic scanning: a) in azimuth plane (FP main lobe $\varphi_s = -\pi/3$); b) in elevation plane (FP main lobe $\theta_s = \pi/4$)



F i g . 7. Intersections of normalized FP of antenna arrays in the form of Mills cross ($2 \cdot 35 - 1$ elements) with electronic scanning: a) in azimuth plane (FP main lobe $\varphi_s = -\pi/3$); b) in elevation plane (FP main lobe $\theta_s = \pi/4$)



F i g . 8. Intersections of normalized FP of antenna arrays in the form of Mills cross with electronic scanning in azimuth plane (FP main lobe $\varphi_s=-\pi/3$) and suppressing parasitic mirror FP lobe ($\varphi_-=\pi/3$): a) array with $2\cdot 65-1$ elements; b) array with $2\cdot 35-1$ elements

($\varphi_-=\pi/3$) with number of elements that are equal to $2\cdot 65-1$ and $2\cdot 35-1$ correspondingly. The corrected FPs $F_c(\varphi,\theta)$ are shown by solid line, whereas FPs $F(\varphi,\theta)|_{\varphi_s}$ without compensating reception at the direction to parasitic mirror FP lobe are shown by dotted line.

Spatial signal filtering under interference conditions in wideband antenna array in the form of Mills cross

Consider the problem of spatial filtering a narrowband useful signal that is solved by a wideband antenna array in the form of Mills cross (Fig. 3a) consisting of $2N-1$ elements (where N is an odd number). We investigate an antenna array with a ratio $f_{\max}/f_{\min}$ maximum to minimum frequencies of processed signals that belong to the interval $f_{\max}/f_{\min}\in[1.1, 2.0]$.

Let $a_i^{x,y}(t)$ be an additive mixture (see formula (1)) observed in the input of i-th element of antenna array orthogonal component (x or y) consisting of useful signal $s(t)$, L complex narrowband interference signals $u_l(t)$, each of them is coming from l-th point emitter, and complex quasi-white Gaussian noise $n(t)$ with zero mean:

$$a_i^{x,y}(t)=s_i(t)+\sum_{l=1}^{L}u_{l,i}(t)+n_i(t)=$$
$$=s(t,\omega_s)\exp\left(-j2\pi\frac{f_s}{f_0}\frac{d}{\lambda_0}\left(i-\frac{N+1}{2}\right)\sin\theta_s\begin{Bmatrix}\cos\varphi_s\\ \sin\varphi_s\end{Bmatrix}\right)+$$
$$+\sum_{l=1}^{L}u_l(t,\omega_l)\exp\left(-j2\pi\frac{f_l}{f_0}\frac{d}{\lambda_0}\left(i-\frac{N+1}{2}\right)\sin\theta_l\begin{Bmatrix}\cos\varphi_l\\ \sin\varphi_l\end{Bmatrix}\right)+$$
$$+n_i(t)\qquad(8)$$

where $t=t_m=t_0+m\cdot\Delta t;\ m=0,1,...,M-1;\ \Delta t$ is a sampling interval of the signals; $s_i(t)$ is the useful signal $s(t)$ observed in i-th channel of antenna array; $\omega_s=2\pi f_s=\text{const}$, f_s is a carrier frequency of the useful signal $s(t)$; (θ_s, φ_s) are the directions of arrival of the signal $s(t)$; (θ_l, φ_l) are the directions of arrival of the interference signal $u_{l,i}(t)$; $\omega_l=2\pi f_l=\text{const}$, f_l is a carrier frequency of the interference signal $u_{l,i}(t)$ received in the i-th channel of antenna array; the frequencies $\{f_l\}$ of the signals $\{u_l(t)\}$ and the carrier frequency f_s of the useful signal $s(t)$ are different and distributed in some interval: $f_l, f_s\in[f_{\min}, f_{\max}]$, $f_0\le f_{\min}$, $f_{\max}/f_{\min}\in[1.1, 2.0]$; $n_i(t)=n_i^c(t)+j\cdot n_i^s(t)$ is a noise observed in the i-th channel of antenna array, $j=\sqrt{-1}$; $\mathbf{M}\{u_l(t)u_k(t)\}=0$, $l\ne k$; $\mathbf{M}\{n_i(t)n_j(t)\}=0$, $i\ne j$; $\mathbf{M}\{*\}$ is the symbol of mathematical expectation; $\mathbf{M}\{(n_i^c(t))^2\}=\mathbf{M}\{(n_i^s(t))^2\}=D_n/2$, D_n is a variance of quasi-white Gaussian noise $n_i(t)$ with zero mean; d is inter-element distance of antenna array, $d=0.5\lambda_0$, where λ_0 is the wavelength corresponding to the minimal frequency of the processed signals $f_0\le f_{\min}$.

The resulting process $y(t)$ in the output of signal processing unit of antenna array in the form of Mills cross consisting of $2N-1$ elements (see Fig. 3a) is defined by weighted sum of delayed signal copies $a_i(t)$ in the receiving channels [15; (10.17)], [16; (61.2)]:

$$z(t)=\sum_{i=1}^{N}\sum_{k=0}^{K-1}[\overline{w}_{i,k}^{x}\cdot a_i^x(t-\Delta_k)+\overline{w}_{i,k}^{y}\cdot a_i^y(t-\Delta_k)],\qquad(9)$$

where $t=t_m=t_0+m\cdot\Delta t;\ m=0,1,...,M-1;\ \Delta t$ is a sampling interval of the signals; $a_i^{x,y}(t)$ is the received signal in the input of i-th receiving element of antenna array orthogonal component (see formula (8)); $w_{i,k}$ is weight coefficient in the i-th processing channel in the output of k-th delay element; $\overline{w}_{i,k}=\text{Re}(w_{i,k})-j\,\text{Im}(w_{i,k})$; Δ_k is a value of signal delay in the output of k-th delay element, $\Delta_{k=0}=0$, $\Delta_k<\Delta_{k+1}$; K is a number of delay elements in each channel.

According to spatial filtering algorithm (9), on the basis of observations $a_i^{x,y}(t)$ in the input of i-th antenna array element (8) it is necessary to form $M\times 2N$ input signals matrix

$$\mathbf{a}=\|a_{m,i}\|=\begin{Vmatrix}a_1^x(t_0) & \dots & a_N^x(t_0) & a_1^y(t_0) & \dots & a_N^y(t_0)\\ a_1^x(t_1) & \dots & a_N^x(t_1) & a_1^y(t_1) & \dots & a_N^y(t_1)\\ \vdots & \ddots & \vdots & \vdots & \ddots & \vdots\\ a_1^x(t_{M-1}) & \dots & a_N^x(t_{M-1}) & a_1^y(t_{M-1}) & \dots & a_N^y(t_{M-1})\end{Vmatrix},$$

including signal samples $a_i^{x,y}(t)$ and their delayed on Δ_k copies, so that the resulting matrix involving these matrices takes the following form:

$$\mathbf{A}=\|A_{m,l}\|=\text{augment}(\mathbf{A}_0,\mathbf{A}_1,...,\mathbf{A}_{K-1});\qquad(10)$$

$$\mathbf{A}_k=\|a_{m,i}\|;\ \mathbf{A}_0=\mathbf{a};\ a_{m,i}=a_i(t_m-\Delta_k),\qquad(10a)$$

where augment$(\mathbf{A},\mathbf{B},\mathbf{C},\dots)$ is the function that returns an array formed by placing matrices $\mathbf{A},\mathbf{B},\mathbf{C},\dots$ left to right; $t=t_m=t_0+m\cdot\Delta t$; $m=0,1,...,M-1$; $i=1,2,\dots,2N$; $k=0,1,\dots,K-1$; $l=1,2,\dots,2N\cdot K$; $\Delta_{k=0}=0$, $\Delta_k<\Delta_{k+1}$.

Thus, matrix A has a size $M\times(2N\cdot K)$, where M is a number of temporal samples of the signal $a_i(t)$ used in processing; $2N-1$ is a number of elements of antenna array in the form of Mills cross; K is a number of delayed elements in each channel. Then the estimator $\hat{\mathbf{R}}_a$ of correlation matrix $\mathbf{R}_a$ that is formed on the basis of observations (8) has the size $(2N\cdot K)\times(2N\cdot K)$:

$$\hat{\mathbf{R}}_a = \left\| R^a_{r,c} \right\| = \frac{1}{M}\left\| \left(\mathbf{A}^{<r>}\right)^T \overline{\mathbf{A}^{<c>}} \right\|, \qquad (11)$$

where $r = 1,\ldots,2N\cdot K$; $c = 1,\ldots,2N\cdot K$; $\mathbf{A}^{<c>}$ is a vector formed on the basis of c-th column of the matrix $\mathbf{A}$ (10); $\overline{\mathbf{A}^{<c>}}$ is complex conjugate to $\mathbf{A}^{<c>}$; M is a number of temporal samples of the signals $a_i^{x,y}(t)$ used in signal processing.

Antenna array operating on the basis of minimum variance criterion provides optimal solution for vector of weight coefficients $\mathbf{w}$ in the following form [15; (3.117)], [17; (61.18)]:

$$\mathbf{w} = [w_1,\ldots,w_{2N\cdot K}]^T = \frac{\hat{\mathbf{R}}_a^{-1}\mathbf{E}(\theta_s,\ f_s)}{\mathbf{E}(\theta_s,\ f_s)^T\hat{\mathbf{R}}_a^{-1}\mathbf{E}(\theta_s,\ f_s)}; \qquad (12)$$

$$\begin{aligned}\mathbf{E}(\theta_s,\varphi_s,\ f_s) = \mathrm{stack}[\ &\mathbf{e}(\theta_s,\varphi_s,\ f_s),\\ &\mathbf{e}(\theta_s,\varphi_s,\ f_s)\cdot\exp(-j2\pi f_s\cdot\Delta_1),\ldots\ ; \qquad (13a)\\ &\ldots,\ \mathbf{e}(\theta_s,\varphi_s,\ f_s)\cdot\exp(-j2\pi f_s\cdot\Delta_{K-1})]\end{aligned}$$

$$\mathbf{e}(\theta_s,\varphi_s,\ f_s) = \mathrm{stack}(\ \mathbf{e}_x(\theta_s,\varphi_s,\ f_s),\ \mathbf{e}_y(\theta_s,\varphi_s,\ f_s)); \qquad (13b)$$

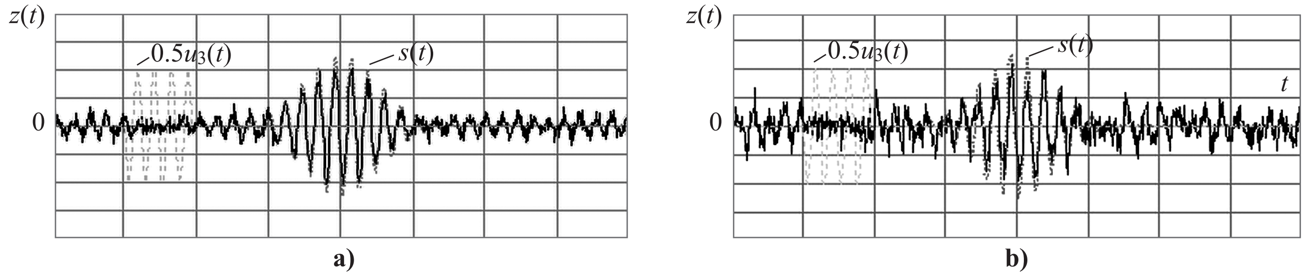
$$\mathbf{e}_{x,y}(\theta_s,\varphi_s,\ f_s) = \left[\exp\left(\begin{matrix}-j2\pi\dfrac{f_s}{f_0}\dfrac{d}{\lambda_0}\left(i-\dfrac{N+1}{2}\right)\\ \sin\theta_s\begin{Bmatrix}\cos\varphi_s\\ \sin\varphi_s\end{Bmatrix}\end{matrix}\right)\right], \qquad (13c)$$

where $\hat{\mathbf{R}}_a$ is the estimator of correlation matrix $\mathbf{R}_a$ (11) of the observations; $\mathbf{e}(\theta_s,\varphi_s,\ f_s)$ is vector of complex spatial harmonics (steering vector) as a function of angles of signal arrival (θ_s,φ_s) and carrier frequency f_s of the signal $s(t)$; $\mathbf{e}_x(\theta_s,\varphi_s,f_s)$, $\mathbf{e}_y(\theta_s,\varphi_s,f_s)$ are the vectors of complex spatial harmonics for orthogonal components of planar antenna array in the form of Mills cross; $\mathbf{E}(\theta_s,\varphi_s,f_s)$ is a modified vector of complex spatial harmonics (steering vector); $\mathrm{stack}(\mathbf{A},\mathbf{B},\mathbf{C},\ldots)$ is the function that returns an array formed by placing matrices $\mathbf{A},\mathbf{B},\mathbf{C},\ldots$ top to bottom.

The considered algorithm (9)…(13) is illustrated by an example that is realized by simulation modelling this algorithm.

Example. Assume that in the inputs of wideband planar antenna array in the form of Mills cross consisting of $2\cdot 35-1$ elements we observe an additive mixture (8) consisting of the useful signal $s(t)$ coming from the direction $(\theta_s, \varphi_s) = (\pi/4, -\pi/3)$, interference signals $u_1(t)$, $u_2(t)$, $u_3(t)$, $u_4(t)$, $u_5(t)$ arriving from the directions $(\theta_1,\varphi_1) = (2\pi/9, -3\pi/8)$; $(\theta_2,\varphi_2) = (4\pi/13, \pi/10)$; $(\theta_3,\varphi_3) = (\pi/3, 3\pi/14)$; $(\theta_4,\varphi_4) = (3\pi/7, 7\pi/18)$; $(\theta_5,\varphi_5) = (2\pi/5, -3\pi/14)$ correspondingly and also uncorrelated quasi-white Gaussian noise $n(t)$, that affects in the receiving channels of antenna array. Interference signals $u_1(t)$ and $u_2(t)$ are amplitude-modulated and frequency-modulated jams respectively; interference signal $u_3(t)$ is a radio-frequency pulse with rectangular envelope and uncorrelated interference signals $u_4(t)$, $u_5(t)$ are quasi-white Gaussian noises. Signal to k-th interference ratios in a single i-th receiving channel of antenna array take the values that are equal to 0.25; 0.25; 1; 0.05; 0.05 respectively. Signal to noise ratio takes the value 30 dB. Signal to resulting interference ratio took the value – 10 dB. The frequencies of received interference signals as well as the frequency of useful signal are distributed in some interval: $[f_{\min}, f_{\max}]$, $f_0 \le f_{\min}$, $f_{\max}/f_{\min} \approx 1.5$; inter-element distance of antenna array is equal to $d = 0.5\lambda_0$, where λ_0 is a wavelength corresponding to minimal frequency of the processed signals $f_0 \le f_{\min}$. The signal delay values Δ_k in the receiving channels (9) are chosen to be equal to $\Delta_k = 0; 1\Delta t; 3\Delta t; 4\Delta t; 5\Delta t$, $\Delta t = 1/(2 f_{\max})$. The number of samples M of signals $a_i^{x,y}(t)$ used in processing is chosen to be equal to $M = 1024$.

Fig. 9a,b illustrate estimates $z(t) = \hat{s}(t)$ of useful signal $s(t)$ that are formed according to spatial filtering algorithm (9)…(13) based of correlation matrix estimator (11), so that the results are shown for the cases of both mechanic and electronic scanning FP main lobe respectively. The difference in signal processing within the framework of general algorithm (9)…(13) lies in the fact that while controlling mechanically the antenna array is considered to be adjusted for the direction of arrival of useful signal so that the vectors of complex spatial harmonics $\mathbf{e}_x(\theta_s,\varphi_s,\ f_s)$, $\mathbf{e}_y(\theta_s,\varphi_s,\ f_s)$ figuring in the formula (13c) are unitary: $\mathbf{e}_x(\theta_s,\varphi_s,\ f_s) = [1,1,\ldots,1]^T$, $\mathbf{e}_y(\theta_s,\varphi_s,\ f_s) = [1,1,\ldots,1]^T$. Coefficients of interference suppression for the first and the second variants of signal processing took the values $K_z = 40\ldots45$ dB and $35\ldots39.4$ dB respectively. Here coefficient of interference suppression is defined as the ratio $K_z = 10\lg(D_J / D[z(t) - s(t)])$, where D_J is a variance of resulting interference, $D_J = \sum_{l=1}^{L} D_l$, D_l is a variance of l-th interference; $D[z(t)-s(t)]$ is a variance of a noise component of the process $z(t)$ in the output of signal processing unit. Both figures illustrate the position of pulse interference $u_3(t)$ that is shown twice smaller in amplitude. As one can see



F i g . 9. Estimates $z(t) = \hat{s}(t)$ of useful signal $s(t)$ that are formed according to spatial filtering algorithm (9)…(13) in planar antenna array in the form of Mills cross consisting of $2\cdot 35-1$ elements for the cases:
a) mechanic scanning FP main lobe; b) electronic scanning FP main lobe

from the figures, amid the same interference environment, antenna array that is adjusted for the useful signal receiving (steering vectors from (13c) are unitary) provides better spatial filtering (~5 dB) than array with electronic scanning the space.

CONCLUSIONS

1. Solving the problem of substantiating contemporary requirements for AJS operating against on-board radar systems of AAM concerning simultaneous providing wide frequency band, large directivity and small both weight and size parameters is possible by exploiting wideband planar antenna arrays built on the basis of new technological principles, namely, basing on exploiting new wideband elements for antenna arrays and also digital signal processing in both receiving and transmitting antenna arrays.

2. Main types of wideband planar antenna arrays built on new technological principles that can be exploited when developing perspective EW stations, are antenna arrays based on Vivaldi elements.

3. Providing wide frequency band, high quality of spatial signal filtering amid interference influence, high directivity and required resolution over angular coordinates while minimizing a number of antenna array elements and hence minimizing the required calculation recourses, is possible on the basis of antenna arrays in the form of Mills cross with Vivaldi elements and digital signal processing.

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АНАЛІЗ МОЖЛИВОСТЕЙ ВИКОРИСТАННЯ ПЛАНАРНИХ АНТЕННИХ РЕШІТОК У ПЕРСПЕКТИВНИХ ЗРАЗКАХ ТЕХНІКИ РАДІОЕЛЕКТРОННОЇ БОРОТЬБИ З БОРТОВИМИ РАДІОЛОКАЦІЙНИМИ СИСТЕМАМИ ЗАСОБІВ ПОВІТРЯНОГО НАПАДУ

Стверджується, що для систем радіоелектронної боротьби характерна проблема одночасного забезпечення значного діапазону робочих частот, потрібних коефіцієнту направленої дії, якості просторової фільтрації сигналів на фоні перешкод і розрізнювальної здатності по кутових координатах. Показується, що вирішення цієї проблеми при обмежених обчислювальних ресурсах можливо шляхом застосування антенних решіток з цифровою обробкою сигналів у виді хреста Міллса на елементах Вівальді, при цьому забезпечуються відносно невеликі масо-габаритні показники антенної системи. Розроблено модель діаграми направленості планарної антенної решітки у вигляді хреста Міллса, на основі якої обґрунтовано основні технічні характеристики до приймальних і передавальних антенних систем перспективних засобів радіоелектронної боротьби з бортовими радіолокаційними системами засобів повітряного нападу. Розроблено модель просторової фільтрації сигналів на фоні перешкод для планарної антенної решітки у вигляді хреста Міллса з електронним способом управління головним променем діаграми направленості та цифровою обробкою сигналів, на основі якої визначено характеристики обробки корисних сигналів в умовах впливу перешкод.

Ключові слова: *планарна антенна решітка, антенна решітка у вигляді хреста Міллса, антенна решітка з цифровою обробкою сигналів, діаграма направленості, коефіцієнт направленої дії, коефіцієнт широкосмуговості, елемент Вівальді, радіолокаційна*

система, радіоелектронна боротьба з засобами повітряного нападу.

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