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Electromagnetic Compatibility Parameters of an Airborne Slot Antenna System

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Abstract*: *For the purpose of building extended surfaces in the long and short wave sections of the wavelength spectrum, it is deemed relevant to know the parameters of electromagnetic compatibility of airborne slot radiator antenna systems operating in the multi-wave receiving mode. One of the efficient research methods is to solve the problem of electromagnetic wave propagation in an infinitely extended slot radiator. The research purpose is to study the electromagnetic compatibility parameters of airborne slot radiator antenna systems operating in the multi-wave receiving mode. The main research methods were the statistical processing of the experimental data of tests in situ and mathematical description of the electromagnetic radar setting, as well as its computer simulation. The main result of the study can be formulated as the development of analytical and software methods for calculating the internal, external and equivalent conductivities of longitudinal and transverse slots on a wide wall of a rectangular waveguide of an antenna system at main frequency harmonics and at frequencies exceeding the main one, with account of slot width. We calculated the radiation characteristics of a slotted-waveguide antenna system at harmonics. The results of the study were used in the development of a universal technique for designing and developing radar systems capable of adequate functioning in the conditions of the electromagnetic setting under consideration. The results of the study will enable a more complete description of the electrodynamic pattern of wave propagation, making it possible to increase their generation and propagation.

Index Terms*: *electromagnetic compatibility, frequency harmonics, rectangular waveguide, slot antenna systems.

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I. INTRODUCTION

Of great importance for radar survey of various locations on the land surface is the quantitative description of phenomena, which are due to its inhomogeneous structure, and accounting for the impact of the conditions of its substrate on the characteristics of received signals [1]. An analysis of electromagnetic wave dispersion characteristics with a perfectly conducting strip shield located on the flat boundary of the conducting dielectric half-space enables studying problems associated with drastic changes in the electrophysical properties of land surface and presence of highly reflective objects on it [2].

II. LITERATURE REVIEW

There are solutions for a perfectly conducting disc and strip that allow evaluating wave dispersion characteristics for various locations [3]. However, the solutions are not applicable for radiometric surveys, as the parameters of thermal radiation of a medium are determined by the electromagnetic wave loss in the medium, i.e. the possibility of radiometric observation of objects depends on the difference between their absorbing capacity. Studies [1–3] considered the problem of diffraction on an infinitely extended strip in empty space. In [4], the problem was reduced to a system of dual integral equations with respect to spectral functions for surface current densities, which could be solved by the method of moments. Such problems for both the impedance strip (Fig. 1) and a slot in the impedance shield can be solved by the eigenfunction method [5–6].

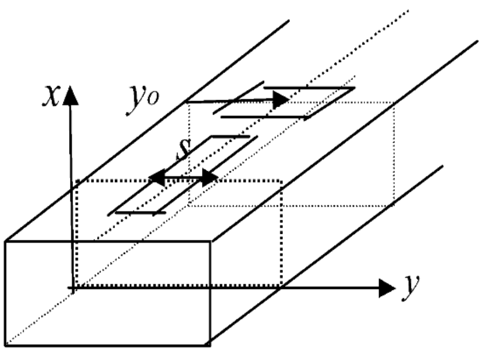

Fig. 1. A slotted-waveguide antenna system

The slotted-waveguide radar system (Fig. 2) is of multi-mode type. Therefore, in the course of radiation and reception, we need the data on the integral characteristics of the radiating systems (for both within the working bandwidth and far beyond it), such as the gain factor (GF) and radiation pattern (RP) in such modes [7].

In non-linear elements, several electromagnetic wave types start propagating at high frequencies. This requires accounting for the peculiar features of electromagnetic waves in solving certain problems of nonlinear radar-location (NRL). Such devices include radar systems of nonlinear slotted radiators. Certain changes in the characteristics of a nonlinear system consisting of longitudinal slotted radiators take place and can be used [8].

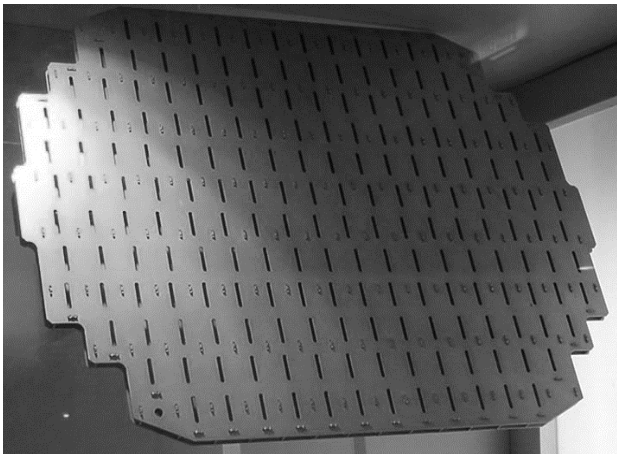

Fig. 2. Slotted radiators in a radar system

Several studies have dealt with theoretical calculation and structural surveys, suggesting to use irregular analytical electrodynamic analysis methods, namely by solving integral and differential equations of higher order through changing the coordinate system, solving a boundary problem by the variable separation method based on expanding a flat electromagnetic wave into series by Mathieu's function, Bessel and Hunkel's functions [10].

The most recent developments were provided by I.Ya. Immoreev, A.G. Loshchilov, [1, 3, 4], in patents and studies by E.V. Semyonov and V.P. Likhachyov [2, 6]. Using the ability of certain objects not only to disperse radio waves striking on them, but also to convert their spectrum enables us to consider a broad list of problems, which are hard or impossible to solve by conventional methods. Submit your manuscript electronically for review.

III. RESEARCH METHODS

The field emitted by such a system is the result of interference of the primary field emitted by a vibrator (vibrators) and secondary (diffracted) field created by the system of currents induced on the surface of such a body by the primary field. Finding the total field becomes easier if we only regard the field in the farther area, i.e. the radiation pattern of the system [11-12]:

The electrodynamic component is proportional to the found component of the total field, parallel to the vibrator axis, and is a function of the angles of incidence of the flat wave. Based on the principal of mutual impact, this radiation pattern is also the radiation pattern of the transmitting vibrator. Its shape depends on the geometrical parameters of the body and coordinates of the vibrator location point.

The infinite impedance strip as well as the infinite slot in the impedance shield are represented as the coordinate surfaces *ξ=0* in an elliptical system of coordinates (Fig. 1). The diffraction field of the randomly oriented slot in the shield in the Cartesian reference system, which does not depend on its orientation, is determined by substituting ratios for the field components in the elliptical system into formulas linking the vector functions in these coordinate systems [13-14].This requires us to express the slot field in an elliptical coordinate system associated with the slot for the case when the electric field vector $\dot{\vec{E}}$ or magnetic field vector $\vec{H}$ of the incident wave is oriented along the slot. The above expressions result from expressing the boundary problem through the variable separation method based on the expansion of a flat wave into series by Mathieu's function [15-16].We introduce the Cartesian coordinate system *XYZ* so the screen plane coincides with the plane *XY*, and the electromagnetic wave incidence plane coincides with the plane *ZY*. The direction of the normal towards the front side of the incident flat wave creates the angle *y* with the axis *OZ*. The straight slot of infinite length and finite width *2h* is rotated at the angle *τ* with respect to the axis *OZ* (Fig. 3) [18-19].

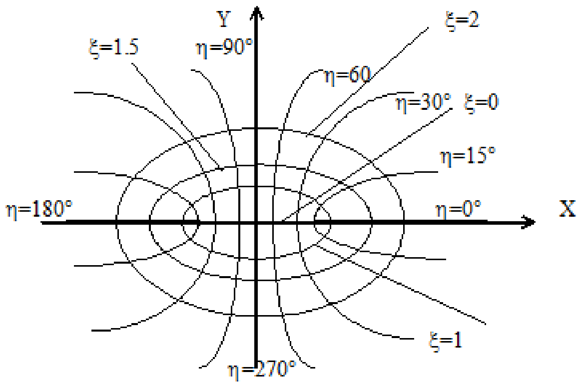

Fig. 3. Radar-location system model

All the following formulas are calculated by authors themselves, if other sources are not mention.

In general terms, the field of the incident flat electromagnetic wave in a free space is formulated as follows:

$$\dot{\vec{E}} = \dot{\vec{E}}_0\left(\vec{x}_0\cos(\alpha_1) + \vec{y}_0\cos(\alpha_2) + \vec{z}_0\cos(\alpha_3)\right)e^{-i\kappa\left(x\cos(\gamma_1)+y\cos(\gamma_2)+z\cos(\gamma_3)\right)}$$

$$\dot{\vec{H}} = \dot{H}_0\left(\sqrt{\frac{\dot{\mu}\mu_0}{\dot{\varepsilon}\varepsilon_0}}\right)^{-1}\left(\vec{x}_0\cos(\beta_1) + \vec{y}_0\cos(\beta_2) + \vec{z}_0\cos(\beta_3)\right)e^{-i\kappa\left(x\cos(\gamma_1)+y\cos(\gamma_2)+z\cos(\gamma_3)\right)}, \quad (1)$$

The total field is the sum of the incident field and the field diffracted by the strip is found as follows:

$$\begin{matrix}\vec{H} = \vec{H}_{inc} + \vec{H}_{dif} \\ \vec{E} = \vec{E}_{inc} + \vec{E}_{dif}\end{matrix}\left\}\begin{cases}\vec{E} = -i\omega\mu_a\vec{A}^e + \dfrac{1}{i\omega\varepsilon_a} grad\ div\ \vec{A}^e - rot\ \vec{A}^m \\ \vec{H} = -i\omega\varepsilon_a\vec{A}^e + \dfrac{1}{i\omega\mu_a} grad\ div\ \vec{A}^m - rot\ \vec{A}^e\end{cases}\right.$$

$$\vec{A}(x,y,z)=\vec{A}(x,z)e^{-ip_0y\frac{1}{a}}=\vec{A}(x,z)e^{-ip_0\frac{y}{a}}$$
$$=\vec{A}(x,z)e^{-ik_0a\sin\theta\cos\varphi_0y\frac{1}{a}}$$
$$=\vec{A}(x,z)e^{-ik_0\sin\theta\cos\varphi_0y}$$
$$p_0=k\sin\theta\cos\varphi_0=k_0a\sin\theta\cos\varphi_0\,, \qquad (2)$$

$\vec{A}\left(\frac{x}{a};\frac{z}{a}\right)=\frac{1}{2\pi}\int_{-\infty}^{\infty}\vec{a}_{\pm}(\varepsilon)e^{i(\varepsilon\frac{x}{a}\pm w\frac{z}{a})}d\varepsilon$ is a solution for complex amplitudes meeting the requirements for radiation on an infinite plane:

$$\vec{a}_{\pm}(\varepsilon)=\frac{1}{2}\frac{z_0}{iw}(\vec{j}^m(\varepsilon)\mp\vec{M}^{-1}[\vec{v}_0\vec{j}^e(\varepsilon)]), \qquad (3)$$

where $(\pm)$ are the spectral densities for amplitudes in the upper and lower half-spaces:

$$w=\sqrt{k^2-\varepsilon^2-{p_0}^2}$$
$$w=\sqrt{k^2-\varepsilon^2-(k\sin\theta\cos\varphi_0)^2}$$
$$\vec{A}(x,y,z)=\int_v\vec{j}^{st}(x',y',z')G(x,y,z,x',y',z')dx'dy'dz'$$

$$G(x,y,z,x',y',z')$$
$$=\frac{1}{(2\pi)^3}\iiint_{-00}^{00}\frac{e^{-i\gamma_1(x-x')-i\gamma_2(y-y')-i\gamma_3(z-z')}}{x_1^2+x_2^2+x_3^2-k^2}dx_1dx_2dx_3$$

$$\gamma=\sqrt{x_1^2+x_2^2-k^2}$$
$$w=\sqrt{k^2-\varepsilon^2-p_0^2}$$
$$-x_2=\sqrt{x_1^2+x_3^2-k^2}=\gamma$$
$$\frac{\partial^2A}{\partial x^2}+\frac{\partial^2A}{\partial y^2}+\frac{\partial^2A}{\partial z^2}+k^2A=\vec{j}^{st}$$
$$A(x,y,z)=X(x)Y(y)Z(z), \qquad (4)$$

The non-homogeneous wave equation is solved through Fourier expansion:

$$\vec{A}(x,y,z)\frac{1}{(2\pi)^3}\iiint_{-00}^{00}g(x_1,x_2,x_3)\,e^{-ix_1x-ix_2y-ix_3z}dx_1dx_2dx_3$$
$$Y(y),Z(z)=X(x)=\frac{1}{\sqrt{2\pi}}\int_{x_1=-00}^{00}g(x_1)e^{-ix_1x}dx_1$$
$$g(x_1,x_2,x_3)=-\frac{1}{(2\pi)^3}\iiint_{-00}^{00}\frac{j^{st}_{(x,y,z)}e^{-ix_1x-ix_2y-ix_3z}}{k^2-x_1'^2-x_2'^2-x_3'^2}dxdydz, \qquad (5)$$

Spectral density depends on the distribution of impressed current across the space. The integral is taken over only those points in the space where there are points.

In this case, the wave equation is solved as follows:

$$\frac{\partial^2\vec{A}}{\partial x^2}+\frac{\partial^2\vec{A}}{\partial y^2}+\frac{\partial^2\vec{A}}{\partial z^2}+k^2\vec{A}=-\vec{j}^{st}$$
$$A(x,y,z)=X(x)Y(y)Z(z), \qquad (6)$$

Then, we find through Fourier function expansion:

$$X(x)=\frac{1}{\sqrt{2\pi}}\int_{x_1=-00}^{00}g(x_1)e^{-ix_1x}dx_1$$
$$Y(y)=\frac{1}{\sqrt{2\pi}}\int_{x_2=-00}^{00}g(x_2)e^{-ix_2y}dx_2$$
$$Z(z)=\frac{1}{\sqrt{2\pi}}\int_{x_3=-00}^{00}g(x_3)e^{-ix_3z}dx_3, \qquad (7)$$

where $g(x_1)$ is spectral density. This expression can be interpreted as an infinite set of flat homogeneous waves in the positive and negative direction of the axis x with the wave velocities $v_{st}=\frac{\omega}{x_1}$, where x_1 acts as the wavelength constant, the amplitude of these waves $g(x_1)$ does not depend on the coordinate x.

Thus, the wave equation can be solved as follows:

$$\frac{\partial^2\vec{A}}{\partial x^2}+\frac{\partial^2\vec{A}}{\partial z^2}-\vec{A}(x,z)e^{-i\frac{p_0y}{a}}\left(\frac{-ip_0}{a}\right)^2+k_0^2\vec{A}(x,z)=-\vec{j}^{st}$$
$$\frac{\partial^2\vec{A}(x,y,z)}{\partial x^2}+\frac{\partial^2\vec{A}(x,y,z)}{\partial z^2}-\vec{A}(x,z)e^{-i\frac{p_0y}{a}}\left(k_0^2-\left(\frac{p_0}{a}\right)^2\right)=-\vec{j}^{st}, \qquad (8)$$

Complex amplitudes satisfying the wave equation and depending on the transverse coordinates x and z, and meeting the requirements of radiation at infinity can be solved through a Fourier integral. To solve the wave equation, we use the method of variable separation, the Fourier method:

$$Z\left(\frac{\partial^2}{\partial x^2}+\frac{\partial^2}{\partial y^2}+k_0^2\right)XY+XY\frac{\partial^2}{\partial z^2}Z=0, \qquad (9)$$

dividing it by XYZ :

$$\frac{1}{XY}\left(Y\frac{\partial^2X}{\partial x^2}+X\frac{\partial^2Y}{\partial y^2}+k_0^2XY\right)+\frac{1}{Z}\frac{\partial^2Z}{\partial z^2}=0$$
$$Y=A(x,z)e^{-i\frac{P_0Y}{a}}$$
$$\frac{1}{XY}\left(Y\frac{\partial^2X}{\partial x^2}+XY\left(k_0^2-\left(\frac{P_0}{a}\right)^2\right)\right)+\frac{1}{Z}\frac{\partial^2Z}{\partial z^2}=0$$
$$\frac{1}{XY}\left(Y\frac{\partial^2X}{\partial x^2}+XY\left(k_0^2-\left(\frac{P_0}{a}\right)^2\right)\right)=-\frac{1}{Z}\frac{\partial^2Z}{\partial z^2}, \qquad (10)$$

Since the left and right sides of the equation are equal and independent, it follows that they are equal to a certain separation constant. Take $\varepsilon^2=G^2$:

$$\begin{cases}\frac{1}{XY}\left(Y\frac{\partial^2X}{\partial x^2}+XY\left(k_0^2-\left(\frac{P_0}{a}\right)^2\right)\right)=\varepsilon^2\\ -\frac{1}{Z}\frac{\partial^2Z}{\partial z^2}=\varepsilon^2\end{cases}$$
$$\begin{cases}Y\frac{\partial^2X}{\partial x^2}+XY\left(k_0^2-\left(\frac{P_0}{a}\right)^2\right)=XY{\varepsilon_0}^2\\ \frac{\partial^2Z}{\partial z^2}=-Z{\varepsilon_0}^2\end{cases}$$
$$\begin{cases}Y\frac{\partial^2X}{\partial x^2}+XY\left(k_0^2-{\varepsilon_0}^2-\left(\frac{P_0}{a}\right)^2\right)=0\\ \frac{\partial^2Z}{\partial z^2}+Z\varepsilon^2=0\end{cases}, \qquad (11)$$

Then, we reduce the first equation by Y and obtain a new system of equations:

$$\begin{cases}\frac{\partial^2X}{\partial x^2}+X\left(k_0^2-{\varepsilon_0}^2-\left(\frac{P_0}{a}\right)^2\right)=0\\ \frac{\partial^2Z}{\partial z^2}+Z{\varepsilon_0}^2=0\end{cases}$$
$$\begin{cases}w_0^2=k_0^2-\varepsilon_0^2-\left(\frac{P_0}{a}\right)^2\\ w=(k_0a)^2-(\varepsilon_0a)^2-(P_0)^2,\\ w=k^2-\varepsilon^2-P_0^2\end{cases} \qquad (12)$$

Consider the following homogeneous equation:

$$\frac{\partial^2 \vec{A}(x,y,z)}{\partial x^2} + \frac{\partial^2 \vec{A}(x,y,z)}{\partial y^2} + \frac{\partial^2 \vec{A}(x,y,z)}{\partial z^2} + k_0^2 \vec{A}(x,y,z) = 0, \qquad (13)$$

Taking into account $\vec{A}(x,y,z) = \vec{A}(x,z)e^{-\frac{ip_0 y}{a}}$, we have (14, 15):

$$\left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z^2}\right)\vec{A}(x,z)e^{-\frac{ip_0 y}{a}} - \vec{A}(x,z)e^{-\frac{ip_0 y}{a}}\left(\frac{p_0}{a}\right)^2 + k_0^2 \vec{A}(x,z)e^{-\frac{ip_0 y}{a}} = 0$$

$$\left(\frac{\partial}{\partial x^2} + \frac{\partial^2}{\partial z^2}\right)\vec{A}(x,z) + \vec{A}(x,z)\frac{1}{a^2}(k^2 - p_0^2) = -\vec{j}^{st}, \qquad (14)$$

where $k = k_0 a$.

$$\vec{A}(x,z) = \vec{X}(x)\vec{Z}(z)$$

$$\vec{Z}(z)\frac{\partial^2 \vec{X}(x)}{\partial x^2} + \vec{X}(x)\frac{\partial^2 \vec{Z}(z)}{\partial z^2} + \vec{X}(x)\vec{Z}(z)\frac{1}{a^2}(k^2 - p_0^2) = -\vec{j}^{st}$$

$$Z\frac{\partial^2 X}{\partial x^2} + X\frac{\partial^2 Z}{\partial z^2} + XZ\frac{1}{a^2}(k^2 - p_0^2) = 0, \qquad (15)$$

Next, separate the variables:

$$\begin{cases} \frac{\partial^2 Z}{\partial z^2} + Z\frac{1}{a^2}(k^2 - \varepsilon^2 - p_0^2) = 0 \\ \frac{\partial^2 X}{\partial x^2} + X\frac{1}{a^2}\varepsilon^2 = 0 \end{cases}, \qquad (16)$$

Using the dimensionless spatial variables $u = \frac{x}{a}$ and $v = \frac{z}{a}$, we obtain the following system of equations:

$$\begin{cases} \frac{\partial^2 Z}{\partial v^2} + Zw^2 = -\vec{j}^{st} \\ \frac{\partial^2 X}{\partial u^2} + X\varepsilon^2 = -\vec{j}^{st} \end{cases}, \qquad (17)$$

where $w = \sqrt{k^2 - \varepsilon^2 - p_0^2}$.

The non-homogeneous wave equations can be solved by Fourier expansion.

The function *X(u)* is defined at the interval $-\infty \le u \le \infty$; therefore, its expansion into a Fourier integral as follows:

$$\vec{X}(u) = \frac{1}{\sqrt{2\pi}}\int_{\varepsilon=-\infty}^{\infty} \vec{g}(\varepsilon)e^{-i\varepsilon u}d\varepsilon, \qquad (18)$$

is similar for the function *Z(v)*:

$$\vec{Z}(v) = \frac{1}{\sqrt{2\pi}}\int_{w=-\infty}^{\infty} \vec{g}(w)e^{-iwu}dw, \qquad (19)$$

Thus, for the vector potential $\vec{A}(x,z)$, we have:

$$(*)\vec{A}(u,v) = \frac{1}{2\pi}\int_{w=-\infty}^{\infty}\int_{\varepsilon=-\infty}^{\infty} \vec{g}(\varepsilon,w)e^{-i\varepsilon u - iwv}\,dw d\varepsilon \;\; (iu)$$

$$k^2 = \varepsilon^2 + p_0^2 + \gamma^2, \qquad (20)$$

where $\gamma = \gamma_0 a$, $-\gamma^2 = w^2$; $w_1 = -i\gamma$; $w_2 = i\gamma$.

Consider the inner integral:

$$L = \int_{w=-\infty}^{\infty} e^{-iwv}\vec{g}(w)\,dw$$

$$\vec{A}(x,z) = \int_s \vec{j}^{st}(x',z')G(x,z;x',z')dx'dz'$$

Substituting the expression for $\vec{A}(u,v)$ into wave equation (12), we obtain (21):

$$\begin{cases} \frac{1}{2\pi a^2}\int_{-\infty}^{\infty}\int_{-\infty}^{\infty}(k^2 - p_0^2 - \varepsilon^2 - w^2)\vec{g}(\varepsilon,w)e^{-i\varepsilon u - iwv}dw d\varepsilon = -\vec{j}^{st} \\ \frac{1}{2\pi}\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} e^{-i\varepsilon u - iwv}dw d\varepsilon = \vec{A}(u,v) \end{cases} \qquad (21)$$

comparing (18) and (21), we find out that the spectral plane of the sought function $\vec{A}(u,v)$ differs from the spectral plane of expansion of the impressed current $-\vec{j}^{st}$ only by a factor of $-\frac{1}{a^2} - (k^2 - p_0^2 - \varepsilon^2 - w^2)$. To define the spectral plane of the sought function $\vec{g}(\varepsilon)$, we multiply (21) by the complex conjugate functions $\frac{1}{2\pi a^2}e^{-i\varepsilon' u - iw' v}$, where ε', w' are fixed values ε, w, and then integrate it over the entire infinite space:

$$-\frac{1}{2\pi}\iint_{-\infty}^{\infty}\vec{j}^{st}(u,v)\,e^{i\varepsilon' u + iw' v}dudv = \iint_{-\infty}^{\infty}(k^2 - p_0^2 - \varepsilon^2 - w^2)g(\varepsilon,w)\delta(\varepsilon' - \varepsilon)\delta(w' - w)d\varepsilon dw$$

$$-\frac{1}{2\pi}\iint_{-\infty}^{\infty}\vec{j}^{st}(u,v)\,e^{i\varepsilon' u + iw' v}dudv = (k^2 - p_0^2 - \varepsilon^2 - w^2)g(\varepsilon',w'), \qquad (22)$$

Then the prime is carried over from wavelength constants ε'и $w'$to spatial coordinates u и v:

$$(k^2 - \varepsilon^2 - p_0^2 - w^2)g(\varepsilon,w) = -\frac{1}{2\pi}\iint_{-\infty}^{\infty}\vec{j}^{st}(u',v')e^{i\varepsilon u' + iwv'}du'dv', \qquad (23)$$

The expression determines spectral density $g(u,v)$ in the expansion (18). it depends on the expansion of impressed current in space, with integral (23) calculated only by those points in space, where there currents:

$$\vec{A}(u,v) = \int_s \vec{j}^{st}(u',v')G(u,v,u',v')du'dv' =$$

$$\int_s \vec{j}^{st}(u',v')\left(\frac{1}{(2\pi)^2}\iint_{-\infty}^{\infty}\frac{e^{i(u-u')\varepsilon - i(v-v')w}}{k^2 - \varepsilon^2 - p_0^2 - w^2}d\varepsilon dw\right)du'dv', \quad (24)$$

S designates the transverse section of the source volume *x* by the plane *y=const*. Then we find the integral *w*:

$$L = -\int_{-\infty}^{\infty}\frac{e^{-i(v-v')w}}{(w - iw')(w + iw')}dw$$

$$w' = \sqrt{k^2 - \varepsilon^2 - p_0^2}$$

$$w' = \sqrt{\varepsilon^2 + p_0^2 - k^2}, \qquad (25)$$

Moving to the plane of the complex variable w and assuming that $k^2 < \varepsilon^2 + p_0^2$, which does not limit the similarity of the results obtained, we need to note that there are two exceptional points (strips) in the expression under the integral sign: one in the upper half-plane $w = iw'$ and the other in the lower half-plane: $w = -iw'$. For $(v - v') < 0$, the integral L can be complemented with a disappearing integral along the circumference of the infinitely large radius in the upper half-plane, where for $w \to i\,\infty$, the expression under the integral sign tends to zero.

In this case:

$$L_{(z-z')<0} = \int_G t(w)dw = \sum_{j=1}^{n} \int_{\gamma_j} f(w)dw + \int_{\gamma_R} f(w)dw$$

$$\int_F f(w)dw = 2\pi i \left(\sum_{j=1}^{n} \underset{w=w_j}{res}\, t(w) + \underset{w=\infty}{res}\, t(w)\right) \quad (26)$$

there is no significant exceptional point at infinity:

$$L_{(v-v')<0} = 2\pi i \underset{w=-iw'}{res} f(w) = \frac{\pi e^{-w'(v-v')}}{w'} \quad (27)$$

Consequently, the Green's function:

$$G(u,v;u',v') = \frac{1}{4\pi}\int_{\infty} \frac{e^{-i(u-u')\varepsilon \pm w'(v-v')}}{w'} d\varepsilon\,, \quad (28)$$

(9) is substituted in (14) with account of (15) the finite functions of densities of the superficial electric and magnetic currents:

$$\begin{cases} \vec{j}^e(u) = \frac{iw}{2\pi z_0}\tilde{M}[\int_{-\infty}^{\infty}[\vec{v}_0\vec{a}_+(\varepsilon)]e^{i\varepsilon u}d\varepsilon - \int_{-\infty}^{\infty}[\vec{v}_0\vec{a}_-(\varepsilon)]e^{i\varepsilon u}d\varepsilon] \\ \vec{j}^m(u) = (-)\frac{iw}{2\pi z_0}\int_{-\infty}^{\infty}[\vec{v}_0[\vec{v}_0\vec{a}_+(\varepsilon)]]e^{i\varepsilon u}d\varepsilon - \frac{iw}{2\pi z_0}\int_{-\infty}^{\infty}[\vec{v}_0[\vec{v}_0\vec{a}_-(\varepsilon)]]e^{i\varepsilon u}d\varepsilon \end{cases}, (29)$$

Consider Fourier transform for finding the spectral density $\vec{a}_{\pm}(\varepsilon)$:

$$\begin{cases} (\pm)\frac{z_0}{iw}\tilde{M}^{-1}\vec{j}^e(u)_{\pm} = \frac{1}{2\pi}\int_{-\infty}^{\infty}[\vec{v}_0\vec{a}_{\pm}(\varepsilon)]e^{i\varepsilon u}d\varepsilon \\ (-)\frac{z_0}{iw}\vec{j}^m(u)_{\pm} = \frac{1}{2\pi}\int_{-\infty}^{\infty}\left[\vec{v}_0[\vec{v}_0\vec{a}_{\pm}(\varepsilon)]\right]e^{i\varepsilon u}d\varepsilon \end{cases}, \quad (30)$$

We take the inverse Fourier transform from this system by multiplying the left side of the equation by the complex conjugate functions $e^{-i\varepsilon' u}$ and integrate the obtained expression over the space with currents, i.e. over the variable and within -1 to 1 (according to the hollow width):

$$\begin{cases} (\pm)\frac{z_0}{iw}\tilde{M}^{-1}\int_{-1}^{1}\vec{j}^e(u)e^{-i\varepsilon' u}du = \int_{-1}^{1}\frac{1}{2\pi}\int_{-\infty}^{\infty}[\vec{v}_0\vec{a}_{\pm}(\varepsilon)]e^{i(\varepsilon-\varepsilon')u}dud\varepsilon \\ (-)\frac{z_0}{iw}\int_{-1}^{1}\vec{j}^m(u)e^{-i\varepsilon' u}du = \int_{-1}^{1}\frac{1}{2\pi}\int_{-\infty}^{\infty}\left[\vec{v}_0[\vec{v}_0\vec{a}_{\pm}(\varepsilon)]\right]e^{i(\varepsilon-\varepsilon')u}dud\varepsilon \end{cases}, (31)$$

Using $\frac{1}{2\pi}\int_{-\infty}^{\infty}e^{i(\varepsilon-\varepsilon')u}dud\varepsilon = \delta(\varepsilon-\varepsilon') = \begin{cases}\infty; with\ \varepsilon' = \varepsilon \\ 0; with\ \varepsilon' \neq \varepsilon\end{cases}$ we find (32):

$$\begin{cases} (\pm)\frac{z_0}{iw}\tilde{M}^{-1}\int_{-1}^{1}\vec{j}^e(u)e^{-i\varepsilon' u}du = \int_{-1}^{1}[\vec{v}_0\vec{a}_{\pm}(\varepsilon)]\delta(\varepsilon-\varepsilon')d\varepsilon \\ (-)\frac{z_0}{iw}\int_{-1}^{1}\vec{j}^m(u)e^{-i\varepsilon' u}du = \int_{-1}^{1}\left[\vec{v}_0[\vec{v}_0\vec{a}_{\pm}(\varepsilon)]\right]\delta(\varepsilon-\varepsilon')d\varepsilon \end{cases}, \quad (32)$$

Consider the vector potential through the spectral density:

$$\vec{A}(u,v) = \frac{1}{2\pi}\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} g(\varepsilon,w)e^{-iu\varepsilon-iwv}dwd\varepsilon\,, \quad (33)$$

Then solve the inner integral:

$$L = \int_{-\infty}^{\infty}\frac{e^{-iwv}dw}{k^2-\varepsilon^2-p_0^2-w^2} = \int_{-\infty}^{\infty}\frac{e^{-iwv}dw}{(-)(w-iw')(w+iw')}\,, \quad (34)$$

Any function $t(z)$ of appropriate behavior in the circle $|z-a|<\rho$ is expanded into a power series, converging in this circle $t(z) = \sum_{n=0}^{\infty} C_n(z-a)^n$, where the coefficient is found by the formulas $C_n = \frac{1}{n!}t^n(a)$ or $C_n = \frac{1}{2\pi i}\int_{|\varepsilon-a|=\rho_1}\frac{t(\varepsilon)d\varepsilon}{(\varepsilon-a)^{n+1}}$, where $\rho_1 < \rho$ is the Tailor's power series expansion near the point $z = a$:

$$\left.\begin{matrix} \sum_{n=0}^{\infty} C_n(z-a)^n \\ \sum_{n=-1}^{-\infty} C_n(z-a)^n = \sum_{n=1}^{\infty}\frac{C_{-n}}{(z-a)^n}\end{matrix}\right\}, \quad (35)$$

Then we multiply the former equation by the vector $\vec{v}_0$ and substitute $\vec{\varepsilon} \to \varepsilon$:

$$\begin{cases} (\mp)\frac{z_0}{2iw}\tilde{M}^{-1}\left[\vec{v}_0\left(\int_{-1}^{1}\vec{j}^e(u)e^{-i\varepsilon u}du\right)\right] = a_{\pm}(\varepsilon) \\ \frac{z_0}{2iw}\int_{-1}^{1}\vec{j}^{M}(u)e^{-i\varepsilon u}du = a_{\pm}(\varepsilon) \end{cases}, \quad (36)$$

Further, by combining the spectral densities for the magnetic and electric currents, we find:

$$a_{\pm}(\varepsilon) = \frac{z_0}{2iw}\left[\vec{j}^{M}(\varepsilon) \mp \tilde{M}^{-1}[\vec{v}_0\vec{j}^e(\varepsilon)]\right], \quad (37)$$

We apply the theorem of mean $a_{\pm}(\varepsilon) = \left(\vec{a}^{M}{}_{\pm}(\varepsilon) + \right.$ *ae±ε12.*

The mean value of $\frac{1}{b-a}\int_a^b f(x)dx$ at the interval [-a;b] is:

$$\vec{A}(u,v) = \frac{1}{2\pi}\int_{-\infty}^{\infty}\vec{a}_{\pm}(\varepsilon)\,e^{i\varepsilon u \pm iwv}d\varepsilon, \quad (38)$$

Spectral density depends on the distribution of impressed current. In its turn, the distribution of impressed current over the strip is associated with the fulfillment of impedance boundary conditions on the strip surface (on the upper and lower sides), i.e. on the coordinate plane *v=0*, for tangential components of the fields $\vec{E}_t$ and $\vec{H}_t$ the proportions are met.

$$\left.\begin{matrix}\vec{E}_t(x,y,{}^{+}0) = zz_0[\vec{v}_0,\vec{H}_t(x,y,{}^{+}0)] \\ \vec{E}_t(x,y,{}^{-}0) = zz_0[\vec{v}_0,\vec{H}_t(x,y,{}^{-}0)]\end{matrix}\right\} when\ |x| < a\ or\ |a| < 1, \quad (39)$$

where H is the capacitative nature of the wave, E is the inductive nature of the wave.

The presence of superficial impedance can be considered as the condition for the existence of a surface wave in the second medium. The flat impedance strip (of dielectric structure) is virtually simulated as a waveguide. Obviously, waveguide-type waves can exist in it. Outside the strip (layer), above and below, there is a surface wave. For the flat dielectric strip waveguide, the critical frequencies $w_{kr}^{(n)} = \frac{n\pi}{d}\frac{c}{\sqrt{\varepsilon_1\mu_1 - G_2\mu_2}} = 2\pi f$.

The condition $w = w_{kr}^{(n)}$ corresponds to the condition $\varphi=\varphi^*$, the Brewster angle, where φ is the wave incidence angle. For $\varphi=\varphi^*$, a simple homogeneous wave propagates along the strip, outside the layer. Consequently, the longitudinal component H_y or E_y is lost at the boundary, meaning there is no dependence on the polarization type, i.e. for $f \le f_{kr}(\varphi \le \varphi^*)$, the waveguide-type wave inside the strip (layer) collapses.

Outside the strip, on the plane *v=0*, the following equations are true:

$$\begin{cases} \frac{1}{4\pi}\int_{-\infty}^{\infty}\{(Z\tilde{M} + \tilde{I})\vec{j}^m(\varepsilon)\}\,l^{i\varepsilon u}d\varepsilon = (-)2z\vec{H}_t^{H} \\ \frac{1}{4\pi}\int_{-\infty}^{\infty}\{(\tilde{M} + Z\tilde{I})\vec{j}^l(\varepsilon)\}\,l^{i\varepsilon u}d\varepsilon = -2[\vec{n}\vec{H}_t^{n}] \end{cases}, \quad (40)$$

For the systems of currents outside the strip $|u|>1$, integral equations (40) are equal to 0:

$$\begin{cases} \frac{1}{4\pi}\int_{-\infty}^{\infty}\vec{j}^m(\varepsilon)l^{i\varepsilon u}du = 0 \\ \frac{1}{4\pi}\int_{-\infty}^{\infty}\vec{j}^l(\varepsilon)l^{i\varepsilon u}du = 0 \end{cases}$$

, (41)

The system of equations (40) and (41) is the initial system for solving by the asymptotic or numerical methods [20-22]. In this case, it is solved by the method of moments. For this purpose, the unknown functions of the density of the currents are expanded into a series by a complete system of functions orthogonal along the strip width.

IV. RESULTS AND DISCUSSION

In the development of the methodology, we used the following initial conditions:

a) the length of the antenna system slot (the length of the working wave λ_0, the slot width S is calculated by the formula [23];

$$L \approx \lambda_0/2 - \lambda_0 \cdot 42.5 \cdot \left(W_c \left(\ln \left(2 \cdot \lambda_0/s \right) - 1 \right) \right)^{-1} \qquad (42)$$

b) the laws of electromagnetic field distribution along the slot:

$\sin\left(\nu k_0 \left(\xi + L/2\right)\right)$ where ν is the frequency harmonic number;

$k_0 = 2\pi/\lambda_0$ where λ_0 is the length of the main wave; coordinates $\xi = z$ for the longitudinal slot of the antenna system radiator, $\xi = y - y_0$ for the transverse slot of the antenna system radiator are the current coordinate (Fig. 1);

c) the law of the electromagnetic field distribution across the antenna system's radiator slot is assumed to be constant $p(\tau) = const$, where $\tau = y$ for the longitudinal slot of the antenna system radiator or $\tau = z$ for the transverse slot of the antenna system radiator;

d) we take into account only the types of waves that propagate in the antenna system and the supply waveguide ($\nu k_0 > \chi_{mn}$, where m, n are the wave type indices; $\chi^2_{mn} = \left(m\pi/a\right)^2 + \left(n\pi/b\right)^2$, where a, b are the waveguide dimensions) [24];

e) the walls of the rectangular waveguide are considered infinitely thin.

Waves of types H_{mn} , E_{mn} propagate in the rectangular waveguide feeding the antenna system [6]. Taking into account the approximations b), c), e), we represent the longitudinal component of the bulk density of the magnetic current along the slot, created by a wave of the H_{mn} type as follows:

$$J^M_{\xi} = U^{(mn)}_{0(\nu)} \sin\left(\nu k_0 \left(\xi + L/2\right)\right) \cdot p(\tau) \cdot \delta(x - x_0)/s , \qquad (43)$$

where $U^{(mn)}_{0(\nu)}$ is the voltage at the slot antinode; $\delta(x - x_0)$ is the -function ($x_0 = b$ is the shift of the slot with respect to the narrow wall of the waveguide). Based on the fact that the lossless radiation power of the waveguide slot $P^{(i)}_{\Sigma} = 0.5 \cdot \left(U^{(mn)}_0\right)^2 \cdot G^{(i)}_{mn}$ [25], we find the internal conduction of the slot at the frequency f_ν of the wave of H_{mn} type by the following formula:

$$G^{(i)}_{mn(\nu)} = 2 \cdot P^{(i)}_{\Sigma} \Big/ \left(U^{(mn)}_{0(\nu)}\right)^2 , \qquad (44)$$

At the same time, $P^{(i)}_{\Sigma}$ is the power carried along the waveguide P_z , which can be found by integrating the Pointing vector over the waveguide cross-section:

$$P^{(i)}_{\Sigma} = P_z = \int_0^a \int_0^b \left(\dot{E}_x \dot{H}^*_y - \dot{E}_y \dot{H}^*_x \right) dx dy , \qquad (45)$$

where $\dot{E}_{x,y}$, $\dot{H}^*_{x,y}$ are the electric and magnetic fields' strengths.

Thus, considering the transverse components $\dot{E}_{x,y}$, $\dot{H}^*_{x,y}$ and given their connection with the longitudinal components $\dot{H}^M_{z(mn)}$, $\dot{E}^M_{z(mn)}$, and the latter's connection with vector potentials $\dot{A}^M_{z(mn)}$, $\dot{A}^M_{y(mn)}$ based on the Fourier method, after integrating (45) over the waveguide cross-section and applying formula (44), we find:

For the longitudinal slot:

$$G^{(i)}_{mn(\nu)} = \frac{\in^2_m \in^2_n}{2\pi\mu_a f_\nu abk^{(\nu)}_{mn}} \frac{(\nu k_0)^2}{\chi^2_{mn}} \left(\frac{\sin \frac{m\pi s}{2a}}{\frac{m\pi s}{2a}} \right)^2 B , \qquad (46)$$

for $\nu = 1, 2, \ldots$, $B = \cos^2 \dfrac{m\pi y_0}{a} \cos^2 \left(\dfrac{\pi}{2} \dfrac{1 + (-1)^\nu}{2} + k^{(\nu)}_{mn} \dfrac{L}{2} \right)$,

for $\nu = 1, 2, \ldots$, $B = \cos^2 \dfrac{m\pi y_0}{a} \cos^2 \left(\dfrac{\pi}{2} \dfrac{1 + (-1)^\nu}{2} + k^{(\nu)}_{mn} \dfrac{L}{2} \right)$,

where $k^{(\nu)}_{mn} = \sqrt{(\nu k_0)^2 - (m\pi/a)^2 - (n\pi/b)^2}$; $\in_t = \begin{cases} 1, & t = 0 \\ 2, & t > 0 \end{cases}$

are Neumann numbers.

For the transverse slot: for convenience, we denote

$$B1 = \sin^2 \left(\frac{1 + (-1)^\nu}{4} \pi + \frac{2m\pi y_0}{a} \right) \cos^2 \left(\frac{1 + (-1)^\nu}{4} \pi + \frac{m\pi L}{2a} \right),$$

$$B2 = \left(\sin \left(\frac{m}{a} \pi \left(y_0 - \frac{L}{2} \right) \right) - \sin \left(\frac{m}{a} \pi \left(y_0 + \frac{L}{2} \right) \right) \cos(\nu k_0 L) \right)^2 ,$$

then, for the H_{mn} types of waves:

$$G^{(i)}_{mn(\nu)} = \frac{2 \in^2_n}{\pi\mu_a f_\nu abk^{(\nu)}_{mn}} \frac{(\nu k_0)^2}{\chi^2_{mn}} \left(\frac{\sin k^{(\nu)}_{mn} \frac{s}{2}}{\nu k_0 - \frac{m\pi}{a}} \right)^2 \left(\frac{m\pi s}{a} \right)^2 B$$

, (47)

for $\nu = 1, 2, \ldots$, $B = 4B1$; for $\nu > 1$, $B = B2$;

then, for the E_{mn} types of waves:

$$G^{(i)}_{mn(\nu)} = \frac{2\in_n^2 \pi\varepsilon_a f_\nu}{ab\left(k_{mn}^{(\nu)}\right)^3} \frac{(\nu k_0)^2}{\chi_{mn}^2} \left(\frac{\sin k_{mn}^{(\nu)} \frac{s}{2}}{\nu k_0 - \frac{m\pi}{a}} \right)^2 \left(\frac{n\pi s}{a}\right)^2 B, \quad (48)$$

for $\nu = 1, 2, \ldots$, $B = 16B1$; for $\nu > 1$, $B = 4B2$;

Next we find the external conductivity of the slot with account of the principle of duality by analogy with the electric vibrator radiation resistance for the longitudinal and transverse position of the slot:

for $\nu = 1, 2, \ldots$,

$$G_\nu^{(\Sigma)} = \frac{2}{\pi W_c} \int_0^{\pi/2} \frac{\cos^2\left(\frac{\pi}{2}\nu \sin\theta + \frac{\pi}{2}\frac{1+(-1)^\nu}{2}\right)}{\cos\theta} d\theta, \quad (49)$$

for $\nu > 1$,

$$G_\nu^{(\Sigma)} = \frac{2}{\pi W_c} \int_0^{\pi/2} \frac{\left[\frac{\cos kL \cos\left(k\frac{L}{2}\sin\theta\right) - \cos k\frac{L}{2}}{\sin\theta \sin kL \sin\left(k\frac{L}{2}\sin\theta\right)} + 1\right]^2}{\cos\theta\left(\sin\theta \sin kL \sin\left(k\frac{L}{2}\sin\theta\right)\right)^{-1}} d\theta, \quad (50)$$

where W_c is the wave resistance of the medium, $k = k_0\nu$.

Using the representation of the slotted-waveguide antenna system as an equivalent long line with equivalent conductivities included in it at intervals, we find normalized equivalent conductivity of the slot:

$$g_{mn(\nu)} = 2G^{(i)}_{mn(\nu)} \Big/ G_\nu^{(\Sigma)}, \quad (51)$$

The radiation pattern (RP) of the antenna system, in accordance with (2), is written as:

$$F_\nu^2(\theta) = \left|\sum_{m=0}^{M}\sum_{n=0}^{M}\sum_{p=1}^{N} F^\nu_{\mu,\omega}(\theta)\sqrt{g_{mn(\nu)}}\left(U^+_{p,mn} + U^-_{p,mn}\right) \times e^{(ik(p-1)d\cos\theta)}\right|^2, (52)$$

for $\nu = 1, 2, \ldots$, $F^\nu_{\mu,\omega}(\theta) = \frac{\cos\left(\frac{\pi}{2}\nu\sin\theta + \frac{\pi}{2}\frac{1+(-1)^\nu}{2}\right)}{\cos\theta}$ is the RP of the single radiator slot; for $\nu > 1$,

$$F^\nu_{\mu,\omega}(\theta) = \frac{\cos kL \cos\left(k\frac{L}{2}\sin\theta\right) - \cos k\frac{L}{2} + \sin\theta \sin kL \sin\left(k\frac{L}{2}\sin\theta\right)}{\cos\theta},$$

where *d* is the distance between the radiators; θ is the angle between the normal and the antenna system; *N* is the number of slots. The amplitude of the radiated electromagnetic field can be found as follows:

$$\left|\tilde{U}_{p,mn}\right| = \sqrt{g_{mn(\nu)}}\left|U^+_{p,mn} + U^-_{p,mn}\right|, \quad (53)$$

where $U^+_{p,mn}$, $U^-_{p,mn}$ are the amplitudes of the strength at the input and output of the p^{th} slot.

The distribution of the equivalent voltage is calculated by formulas (53) from the last N^{th} slot to the first slot of the antenna system:

$$U^-_{N,mn} = G^{(\nu)}_{mn} U^+_{N,mn} \exp\left(-2ik^{(\nu)}_{mn} d_H\right),$$
$$\left|U^+_{N,mn}\right| = \left(P^{(H)}_{mn(\nu)} \Big/ \left(1 - \left|G^{(\nu)}_{mn}\right|^2\right)\right)^{1/2}, \quad (54)$$

where $G^{(\nu)}_{mn}$ is the coefficient of reflection from the antenna system load.

The directivity factor (DF) $D_\nu(\theta_0)$ and the gain factor (GF) $G_\nu(\theta_0)$ of the antenna system towards θ_0 can be found by the following formulas:

$$D_\nu(\theta_0) = 2\left|F_\nu(\theta_0)\right|^2 \Big/ \int_{-\pi/2}^{\pi/2} \left|F_\nu(\theta)\right|^2 \cos\theta d\theta, \quad (55)$$

$$G_\nu(\theta_0) = \eta_\nu D_\nu(\theta_0), \quad (56)$$

where θ_0 is the angle, at which the radiation pattern takes the maximum value;

$$\eta_\nu = 1 - \left(\sum_{m,n=0}^{M}\left|U^-_{1,mn}\right|^2 + \sum_{m,n=0}^{M}\left|P^{(H)}_{mn(\nu)}\right|^2\right) \Big/ \sum_{m,n=0}^{M}\left|U^+_{1,mn}\right|^2, \quad (57)$$

is the efficiency factor; $P^{(H)}_{mn(\nu)}$ is the power absorbed in the load of the antenna system.

Fig. 4 and 5 introduce the radiation pattern of a linear slotted-waveguide antenna system with longitudinal slots.

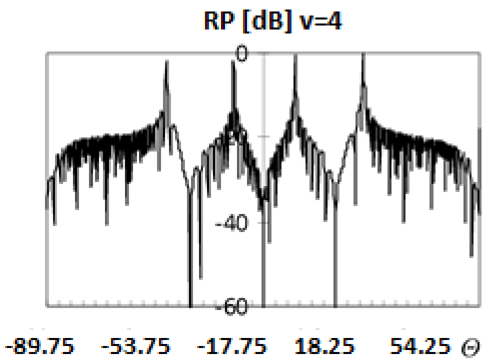

Fig. 4. Radiation pattern of the antenna system at the 4th harmonic

We investigated the parameters of electromagnetic compatibility of the antenna system with the original values: the number of slots of the antenna system *N=118*, the distance between the slots of the antenna system $d=0.575\lambda_0$, the size of the rectangular supply waveguide $a \times b=d=0.78\lambda_0 \times 0.35\lambda_0$, the amplitude distribution of the electromagnetic field in the aperture of the antenna system $U_p=1-0.95cos2(2\pi(p-1)/(N-1))$.

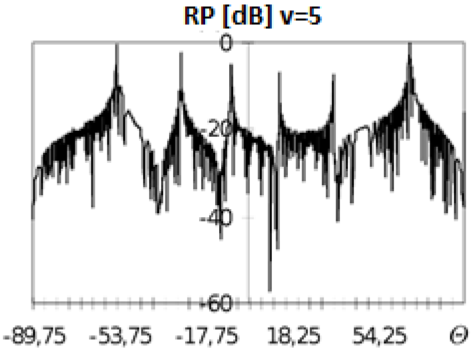

Fig. 5. Radiation pattern of the antenna system at the 5th harmonic

In the radiation pattern (RP) of such an airborne antenna system in the multi-mode mode, there are side highs for an even number of harmonics, and side highs for an odd number of harmonics. Compared to the fundamental frequency, the directivity factor (DF) of the 2nd frequency harmonic is smaller by 5.85 dB; of the 3rd frequency harmonic is smaller by 9.21 dB; of the 4th frequency harmonic is smaller by 11.55 dB; of the 5th frequency harmonic is smaller by 13.23 dB. The gain factor (GF) at the frequency harmonics decreases: 2nd by 7.8 dB, 3rd by 11.7, 4th by 19.3 dB, 5th by 26.8 dB.

V. CONCLUSION

The main result of the study can be formulated as the development of analytical and software methods for analyzing arrays of nonlinear radiators in the multi-wave reception mode, as well as a program methodology for calculating the characteristics of electromagnetic compatibility at frequencies exceeding the fundamental frequency with account of the heterogeneity of electromagnetic fields in such conditions [26].

As part of the main research line, we achieved the following particular objectives:

1) developed a mathematical model of electromagnetic wave propagation in a nonlinear radiator;

2) as a result of the analysis, obtained formulas for calculating dependencies that to a satisfactory extent agree with experimental data;

3) developed a numerical algorithm for studying electrodynamic effects.

In the course of the research, we developed a corresponding algorithm for asymptotic or numerical study of the electromagnetic compatibility parameters of an airborne antenna system in the multi-mode mode, and obtained numerical results using software simulation, which can be used in nonlinear radar location and electromagnetic compatibility tasks when using airborne radio systems of this type. The developed methodology for studying the characteristics of electromagnetic waves can be used for designing and developing radar systems capable of adequate functioning in the conditions of the electromagnetic setting under consideration.

The results of the study can be taken as a basis for developing means of electromagnetic compatibility of radar facilities. Radar system efficiency improvement methods can be developed based on the recommendations proposed in this article.

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2004-2017 – Head of the Department of Higher Mathematics and Information Technologies of the Federal State Budgetary Educational Institution of Higher Education "Chuvash state University named after I.N. Ulyanov"
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The main field of Pichugin's research is the study of the prospects for the development of microstrip antenna technology for identification problems. Among Pichugin's scientific papers, there are 3 articles in the Scopus and Web of Science journals, and the monograph "Antenna grid of slot-hole emitters at the main frequency and in the multi-wave reception mode".

SOLDATOV Anton Aleksandrovich, Phd Candidate of Technical Sciences. Soldatov is the author of more than 30 publications and 1 textbook. The main field of Soldatov's scientific research is the study of the prospects for the development of information-measuring and control systems for the tasks of electric power industry. On this issue Soldatov published 12 scientific papers, including 4 articles in journals included in the HAC (Higher Attestation Commission) list.

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Since 2009, Paravina is the Deputy Dean of the Alatyr branch of the FSBEI of HE "Chuvash state University named after I.N. Ulyanov" on scientific work.
Field of scientific interests Paravina connected with the history of the formation and development of gymnasium education in Russia from the second half of the XIX century to the present, the problems of civil and patriotic education of youth, the historiography of science, as well as issues of the socio-economic development of territories.

DUBROVINA Olga Alexandrovna, Phd Candidate of Economic Sciences, Associate Professor.
Dubrovina is the author of more than 40 publications, including 5 articles in leading peer-reviewed journals, 6 teaching and methodical works.
In 2009 she defended her thesis on the specialty Economics and Management of the National Economy (Regional Economy) on the topic "Formation of the mechanism of interstructural interconnections of industrial enterprises of the region".
Among the scientific interests of Dubrovina, there are the directions for the development of interstructural interconnections of industrial enterprises in the region, including the enterprises of radio engineering.
Since 2017 she is the Chairman of the Methodological Commission of the Alatyr branch of FSBEI of HE "Chuvash state University named after I.N. Ulyanov".