THE DEVELOPMENT OF ANTENNA THEORY AND TECHNIQUES IN LVIV POLYTECHNIC NATIONAL UNIVERSITY

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Abstract

The results of the theory and techniques of modulated impedance and dielectric antennas, antenna arrays with modulated design parameters, mathematical models, test specimens, electrical characteristics and technical features, as well as progress in the development of antennas for the problems of remote monitoring facilities.

Keywords: Theory, technique, modulated impedance antenna, mathematical model, test specimens, electrical characteristics, objects monitoring, microwave antennas.

1. INTRODUCTION

The development of ultra-high frequency technology (UHF) at Lviv Polytechnic Institute begins in 1959 at research laboratories RL-2, from which eventually appear Research Laboratory RL-16 and then in 1986 was created the Research department of radio systems. Research area covered a wide range of problems of development the antenna theory and techniques, including complex antenna systems, including phased antenna arrays. Significant contribution to the development of the antenna theory and techniques, as well as training of specialists highly qualified to antenna problems made Professor A. F. Chaplin, who worked at the Lviv Polytechnic Institute from 1979 to 1991 years.

Currently, development of the theory of antennas is conducted at areas such as:

1. Development of mathematical models, methods of analysis and constructive synthesis of radiating and waveguide structures with N-fold periodicity on the basis of a new class of chain branching fractions;
2. Generalization of analysis techniques and development of mathematical models of antenna systems for the problems of Earth remote sensing;
3. Development of mathematical models and methods for wave propagation in complex heterogeneous mediums, particularly in radiophysical models of the human body for medical and ecological issues and problems of unauthorized access to facilities.

Technology of antennas covers a wide range of utilization for telecommunications systems and remote monitoring of facilities and medical needs, the novelty of which is patented in Ukraine.

2. THE DEVELOPMENT OF ANTENNA THEORY

2.1. PROBLEM FORMULATION AND ITS SOLUTION

Excitation problems of broad class of modulated impedance and dielectric structures that underlie the construction of modulated surface wave antennas (Fig. 1) can be reduced to the analysis of integral equations of the form [1,2]:

$$g(y)I(y) - \int_{-\infty}^{\infty} G(y-y')I(y')d(y') = H(y), \quad (1)$$

where $-\infty \le y \le \infty$; $g(y)$ - surface admittance distribution; $I(y)$ - magnetic currents distribution; $H(y)$ - foreign currents distribution.

Let functions $g(y), I(y), G(y-y'), H(y)$ permit the conversions of Fourier transformation of the type moreover: $\tilde{f}(\chi) = \frac{1}{2\pi}\int_{-\infty}^{\infty} f(y)e^{i\chi y}dy,$

moreover: $g(y) = g_0 + \sum_{i=1}^{N} g_{M_i} \sum_{n_i=-\infty}^{\infty} e^{-(y-n_i d_i)^2/2a^2}$,

where

$d_N = \prod_{i=1}^{N} p_i d_0$; p_i is a sequence of integers;

$g_0, g_{M_i} \in C; d_i, a \in R; n_i \in N.$

Then similarly as it is in [3] it is possible to show, that for any integer N a required functions $\tilde{I}_N(\chi)$ is described by such recurrence formula:

$$\tilde{I}_N(\chi) \cong \tilde{I}_{N-1}(\chi) - \frac{W_N \sum\limits_{n_N=-\infty}^{\infty} \dfrac{\tilde{I}_{N-1}(\chi - n_N T_N)}{\exp(\alpha n_N^2)}}{\prod\limits_{m=0}^{N} G_{m,\Delta}(\chi)}; \quad (2)$$

where

$$G_{N,\Delta}(\chi) = 1 + W_N \sum_{n_N=-\infty}^{N=\infty} \frac{\exp(\alpha n_N^2)}{\prod\limits_{m=1}^{N} G_{m-1,\Delta}(\chi - n_N T_N)};$$

$W_N = g_{M_N} \Delta \sqrt{2\pi}$, $\Delta = a / d_1$, $T_N = 2\pi / d_N$,

$\alpha = 2\pi^2 \Delta^2$, $\tilde{I}_0(\chi) = \tilde{H}(\chi) / G_{0,\Delta}(\chi)$,

$G_{0,\Delta}(\chi) = \tilde{G}(\chi) + g_0$.

The function $\tilde{I}_N(\chi)$ describes the generalized directional characteristic of periodically inhomogeneous impedance and dielectric structures. Recurrent formula (2) generalizes a wide class of mathematical models of modulated surface wave antennas. These mathematical models in the unfolded form create a new class of chain branching fractions arising in problems of diffraction of waves on the structures of N-fold periodicity.

2.2. Mathematical models of antenna arrays with multiple periodicity

The system of Kirchhoff equations in single-mode case for rectilinear equidistant antenna array (AA), based on identical type of emitters, which are loaded on various resistances, has the form [4]

$$Z_0 Z_L(n) \dot{I}_n + \sum_{m=-N}^{N} Z_{n-m} \dot{I}_m = \dot{U}_n, \quad n = 0, \pm 1, ..., N, \quad (3)$$

or in matrix form:

$$\left\{ \left[Z \right] + E Z_0 \mid Z_L \mid \right\} \mid \dot{I} \mid = \mid U \mid, \quad (4)$$

where $\left[Z \right]$ — Toplitz matrix of mutual resistances; E — unitary matrix; Z_0 — own resistance of emitter; $\mid \dot{I} \mid$ — column vector of unknown amplitudes of harmonic current; $\mid U \mid$ — known column vector of extraneous voltages; $\mid Z_L \mid$ — load impedance distribution. Lattice multiplier $f_P(\chi)$, which is connected with the complex amplitude of currents $\dot{I}_n$ with following relations:

$$f_P(\chi) = \sum_{n=-N}^{N} I_n e^{ind\chi}, \quad (5)$$

where d — distance between neighbouring emitters; $2N+1$ — quantity of emitters in AA.

In the paper [5] was constructed a mathematical model that describes the electromagnetic characteristics of AA for a wide class of lattice load impedance (LI) modulation laws. Such laws are the result of the imposition of one single periodic sequence with multiple periods, i.e. when the system (3) or (4):

$$Z_L \mid (n) = 1 + \sum_{i=1}^{M} Z_{L_i}(n), \quad (6)$$

Where $Z_{L_i} = \begin{cases} Z_{L_i}, n = mL_M \\ 0, n \neq mL_M \end{cases}$; $\quad L_M = \prod\limits_{i=1}^{M} l_i$;

$m = 0, \pm 1, ...$ M — integer; $l_i > 1$, integers.

In this case, the lattice factor $f_{PM}(\chi)$ is calculated by formulas that are built using the following recurrent relation:

$$f_{PM}(\chi) = f_{P(M-1)}(\chi) - \frac{\dfrac{Z_{LM}}{L_M} \sum\limits_{m_M=0}^{L_M-1} f_{P(M-1)}(\chi - m_M \dfrac{T}{L_M})}{\prod\limits_{m=0}^{M} D_m(\chi)},$$

where (7)

$$D_M(\chi) = 1 + \frac{Z_{LM}}{L_M} \sum_{m_M=0}^{L_M-1} \frac{1}{\prod\limits_{k=1}^{M} D_{k-1}(\chi - m_M \dfrac{T}{L_M})}$$

To construct solutions of the problem for an arbitrary integer M in the recurrent formula (7) as a source to its constituents substitute the following expressions:

$$f_{P0} = \tilde{U}(\chi) / D_0(\chi); \quad D_0(\chi) = Z_0 + \tilde{Z}(\chi); \quad (8)$$

Sign ~ (tilde) in (6) used to denote the discrete Fourier transform.

$$\tilde{U}(\chi) = \sum_{n=-N}^{N} U(n) e^{in\frac{2\pi}{T}\chi}; \quad \tilde{Z}(\chi) = \sum_{n=-L}^{L} Z(n) e^{in\frac{2\pi}{T}\chi},$$

where $L = 2N$; $T = 2\pi / d$.

Thus, construct a mathematical model for the modulation of AA LI periodic sequence of one pulse functions. Further, on the basis of the decision construct a mathematical model using the recurrent formulas (7) for the case of periodic modulation LI AA sequence of impulse functions with multiple intervals. This process is used more for more complex modulation laws LI antenna array.

3. Antennas for telecommunication systems

This section presents the research laboratories of antennas and ultra-high frequency devices for telecommunications systems, communication systems, radio internet and medical needs. Fig. 1. represents current models of modulated surface wave antennas.

a) b)

Fig. 1. Modulated impedance surface waves antennas centimeter (a) [6] and millimeter-wave bands (b).

3.1 Technical specifications surface waves antennas

Modulated dielectric inhomogeneous pivot antenna (a) [6] and disk-on-rod antenna (b) have a discoid directional characteristic (Fig. 2.) and provide the transmission of messages with randomly distributed correspondents.

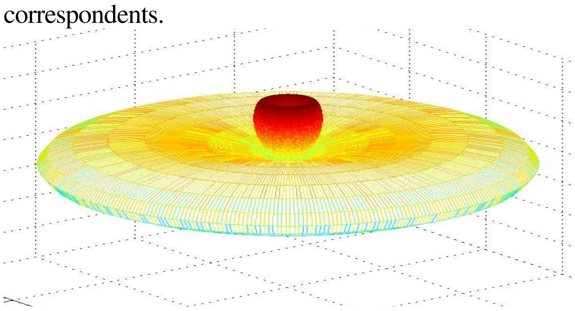

Fig. 2. Directional characteristic of modulated surface wave antennas Fig. 1. (a, b).

Electrical length of antenna is 10λ, where λ — wave length.

Directional characteristic width at half power is $6{,}5^0$.

3.2. Application and technical specifications of complex antenna system

Ring antenna array (Fig. 3a) is designed for study of the construction of facilities microwave hyperthermia for medical purposes. Operating frequency — 2.4 GHz. The maximum diameter of the working area of microwave hyperthermia is 300 mm [7].

Pseudo-parabolic antenna (Fig. 3b) and the antenna based on slot transmission line (Fig.3c) are intended for use in radiointernet at a frequency of 2.4 GHz. Gain Antenna (Fig. 3b) is 16 db, and antenna (Fig. 3c) 12 db respectively.

Reflecting antenna array (Fig. 3d) with dimensions of 300mm * 300mm designed to work in a range 9,75 ÷ 10,5 GHz.

Gain Antenna ≈ 28 db.

Side lobe level of ≤ 14.5 db.

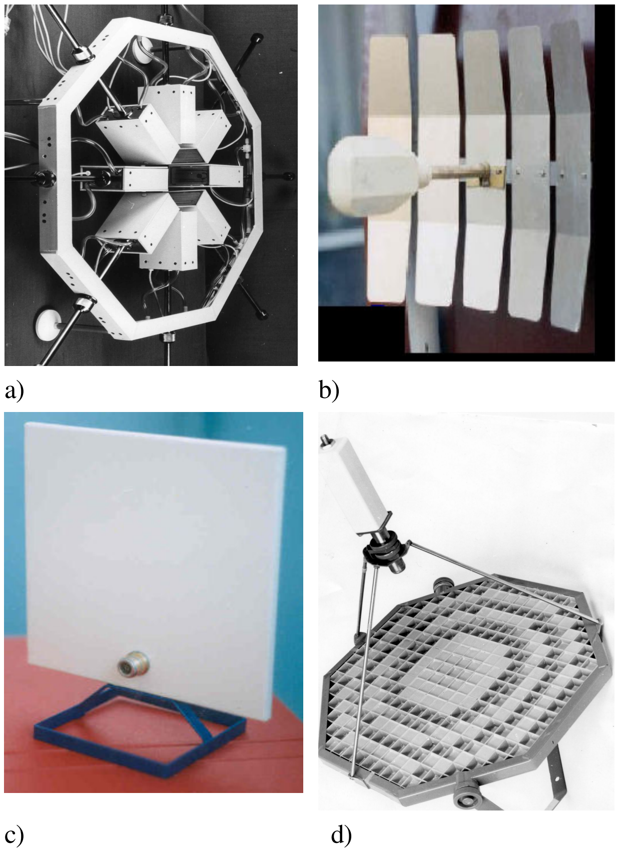

Fig. 3. Complex antenna systems. a) ring antenna array for microwave hyperthermia [7]; b, c) an antenna for radiointernet; d) reflecting antenna arrays for communication systems in the 3 cm band waves.

4. Antennas for systems of monitoring objects and scenes

Extensive application of monitoring systems in geology, geodesy, agriculture, military purposes and emergencies calls for the development of methodology of their construction, creation of new methods of real-time data processing, increasing information demands, decreasing energy and mass-dimensional characteristics.

Most monitoring systems are radiotechnical systems of active and passive location. Their qualitative characteristics, to a great extent, are determined by the parameters of antenna-feeder section which is a complex system with distributed feeder lines and high-frequency devices for forming and controlling position of radiation pattern. Thus, the demands set to antennas of monitoring systems are big and often, in a way, contradictory. Such antennas are to ensure the following:

- high discrimination;
- high coefficient of aperture surface utility;
- low level of side lobes;
- possibility to control spatial position of the main lobe of radiation pattern;
- ability to receive signals with arbitrary polarizations;
- optimal economic and mass-dimensional characteristics while providing preset electric parameters and characteristics.

In monitoring systems at frequencies higher than 1 GHz there are used one- and two- reflector antennas, lens antennas, horn antennas and phased arrays on the basis of wave-aperture, dielectric-rod and microband elements. At lower frequencies there are used phased arrays on the basis of vibratory elements, helical antennas.

Within the frames of developing antennas for monitoring systems there is presented an antenna with orthogonal field polarizations and a polarization-adaptive array.

4.1 An antenna with orthogonal field polarizations

Fig. 4 presents an antenna structure that, in a single design solution, is intended for ensuring reception of information signals with arbitrary polarization structure. The antenna is built on two sections 1 and 1' of rectangular waveguide with two open walls.

The sections are powered by two types of exciters:

- an exciter of electric type in the form of a transformer on a strip line 2 connected to metal wavelike sections in points F and F' and a symmetrization barrel 3 presenting the model as an electric radiator;
- an exciter of a magnetic type in a waveguide section aperture in the form of a vibrator with a symmetrizing device 4 connected in points M and M' to side walls of sections 1 and 1' that are excited cophasingly and this presents the model as a magnetic radiator.

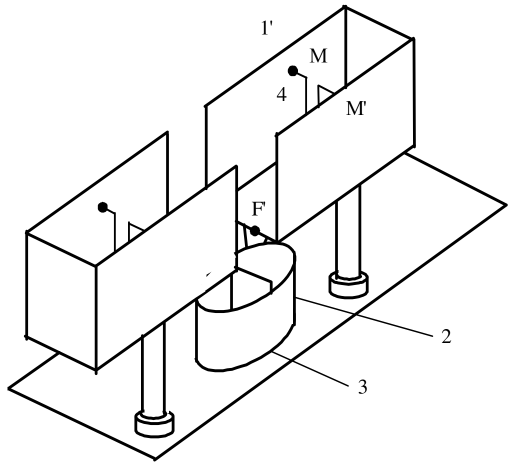

Fig. 4. Antenna structure with orthogonal field polarizations.

Electrodynamic analysis of the antenna is performed on a radiator model presented in fig. 5.

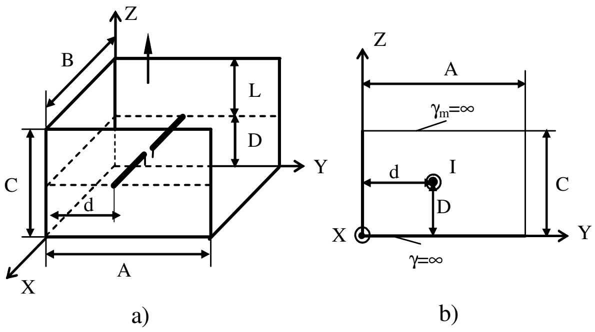

Fig. 5. Model of antenna radiator (a) and its coordinate system (b).

Allocation of exciters and structure configuration allow to consider that the field in the structure is transverse-electric with TE wave. The resultant field of such a radiator in the observation point is determined by the superposition of fields radiated by apertures AB and BC.

$$\dot{E}_x = E_x^A + E_x^D + E_x^L \,, \tag{9}$$

where E_x^A is intensity of the field electric component created by aperture AB

$$E_x^A = \frac{S_A \cdot e^{-j\beta_0 r}}{2r \cdot \lambda_0}\left(E_{\max.}\right)_A \left(1+\cos\Theta\right)\times \sum_{k=1}^{P} \sin\left(\frac{\pi}{2}\left(1-\frac{2k-1}{2P}\right)\right)\times \exp\left(-j\frac{\pi}{2q}\cdot\frac{2k-1}{2P}\sin\Theta\right), \tag{10}$$

where $S_A = \dfrac{B \cdot A}{P}$ is the area of Huygens elements and P is their quantity;

E_x^D is intensity of the field electric component created by aperture BD:

$$E_x^D = \frac{S_D \cdot e^{-j\beta_0 r}}{2r \cdot \lambda_0}\left(E_{\max.}\right)_D \times \sin\Theta \times \\ \times \sum_{i=1}^{Q_D} \sin\left(\beta_l Z\right) \times \exp\left[-j\beta_0 \left(C - \frac{D}{2Q_D}\left(2i - 1\right)\right) \cdot \cos\Theta\right], \quad (11)$$

where $S_D = \dfrac{B \cdot D}{Q_D}$ is the area of aperture radiator on aperture BD;

Q_D is the number of elementary aperture radiators on aperture BD. Electric intensity created by aperture BL is

$$E_x^L = \frac{S_L \cdot e^{-j\beta_0 r}}{2r \cdot \lambda_0} \cdot \left(E_{\max.}\right)_L \cdot \sin\Theta_x \times \sum_{n=1}^{Q_L} e^{-j\beta_l\left(L - Z\right)} \times \\ \times \left(1 + p \cdot e^{-j2\beta_l Z}\right) \times \exp\left[-j\beta_0 \frac{L}{2Q_L}\left(2n - 1\right) \cdot \cos\Theta\right] \quad (12)$$

where $S_L = \dfrac{B \cdot L}{Q_L}$ is the area of elementary aperture radiator on aperture BL;

Q_L is the number of aperture radiators. In expressions (10) – (12) the origin of coordinates of field point (r, Θ) in plane $x = \dfrac{B}{2}$ has been chosen in point $y = A, x = C$. Angle Θ has been defined between axis *Z* and distance direction *r*.

The presented expressions (9) – (12) allow to define the waveguide section field as a magnetic radiator. The resultant field of two sections of (antenna) waveguide is determined by superposition of their fields.

Radiation patterns of electric and magnetic elements of excitation in meridian plane are presented in fig. 6 and fig. 7, where dependences 1 are valid $0{,}85f_0$, dependences 2 – for f_0 - an average antenna frequency, dependences 3 – for $1{,}15f_0$.

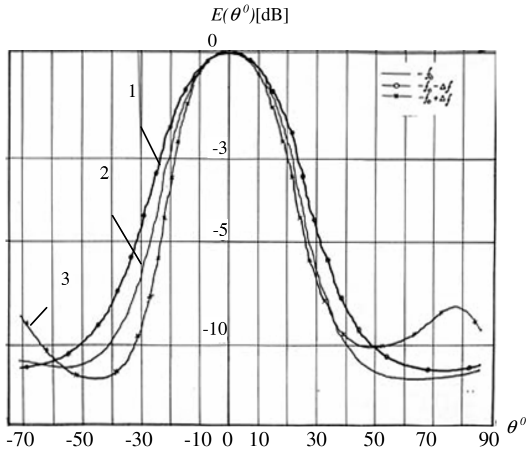

Fig. 6. Radiation pattern of an electric radiator in a meridian plane.

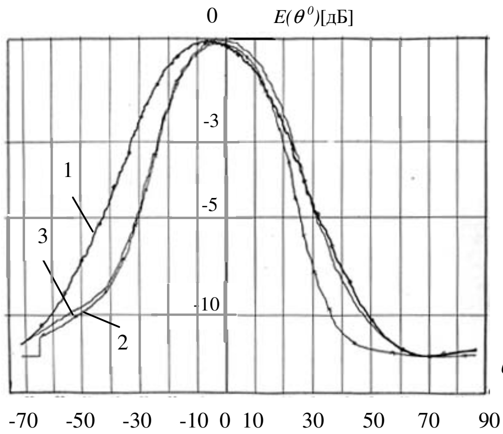

Fig. 7. Radiation patterns of a magnetic radiator in a meridian plane.

4.2. Polarization-Adaptive Array

There often arises the task of signal spatial filtration because of the noise caused by reflections from the earth and water surfaces as well as from the local objects.

Solution to this scientific and technical task can be found by constructing polarization-adaptive arrays [8-10], containing an adaptive processor, a device providing complex weighing of a signal, a commutator ensuring the necessary polarization of the radiator, and an algorithm to maximize signal/interference ratio plus noise.

To exclude polarization losses it is advisable for the antenna element to be able to receive signals with arbitrary polarization (vertical, horizontal, circular left and circular right), for example, as shown above. But at this, the construction of antenna elements becomes more complicated and interconnection between antenna elements increases.

For this reason, there were developed two versions of experimental arrays samples consisting of two types of radiators. The first type of radiators (V) is intended for receiving radio signals with linear (vertical) field polarization and has one coaxial input with 50 ohm resistance. The second type of radiators (VH) [11] is intended for receiving radio signals with orthogonal polarizations and has two coaxial inputs with 50 ohm resistance.

Radiator V of the first version is constructed in the form of two cophased half-wave vibrators of strip type being placed above a solid screen and ensuring reception of waves with vertical field polarization. The use of two combined vibrators is caused by the necessity to ensure the required beamwidth in the vector plane of field electric component for one array radiator.

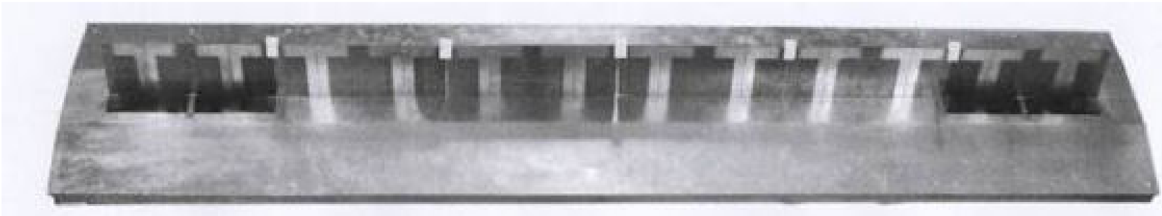

Fig. 8. An experimental sample of an array fragment with microstrip radiators.

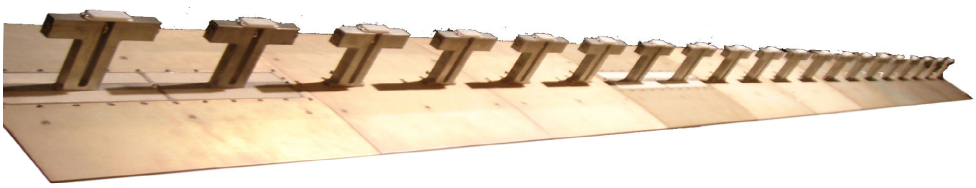

Fig. 9. An experimental sample of an array fragment with radiators of rigid construction.

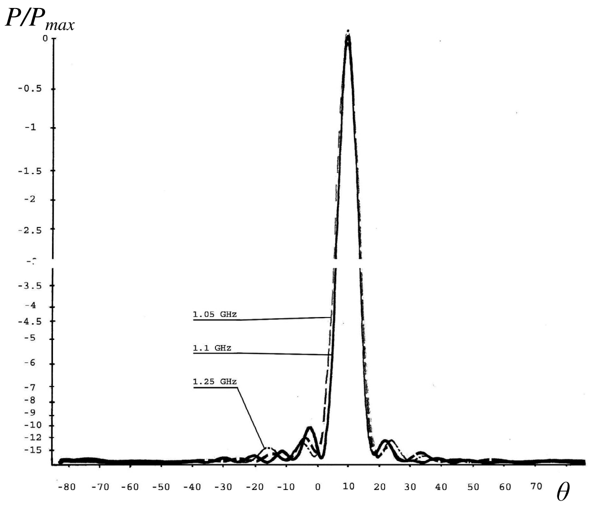

Fig. 10. Radiation pattern in a vertical plane in the channels of the adaptive spatial filter.

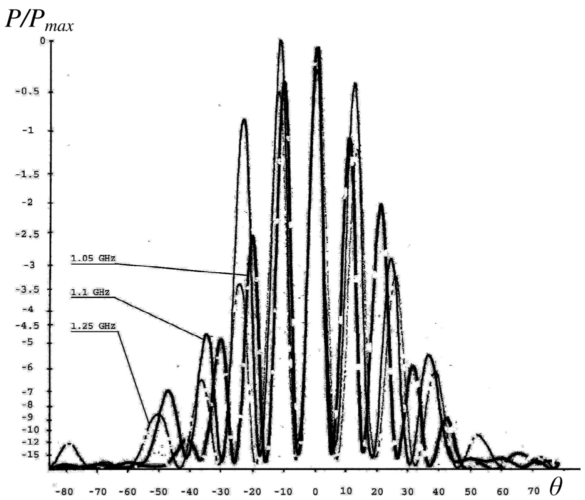

Fig. 11. Radiation pattern in a vertical plane in interferometer channels.

Radiator VH combines two radiators: radiator V to receive waves with vertical polarization and waveguide radiator H to receive waves with horizontal polarization. Radiator waveguide chamber H has an enlarged broad wall to ensure the necessary beamwidth in a vector plane of field magnetic component. To equalize field distribution in the radiator aperture H there was performed co-phased excitation of the waveguide chamber in two points using probes. The experimental sample of an array fragment to receive radio signals with orthogonal field polarization of the first version is shown in fig. 8.

An experimental sample of an array fragment performed by second technological version [12] differs from the first version in that antenna elements in the form of vibrators (fig. 9) are made of aluminum plate material and placed coaxially on a pattern-forming screen.

The first version of prodicing array elements is more technologically advantageous for batch production, but from the point of view of meeting the demands of stability and repeatability of spatial position of phase centers along reception channels the second version of array elements construction is more advantageous.

In experimental samples of arrays there were received the following parameters and characteristics: SWR for radiators of both polarizations in a set fre

quency range is not more than 1.5. The beamwidth of an array along azimuth for vertical polarization in adaptive spatial filter channels is: in a horizontal plane $(120±20)^o$; in a vertical plane – $(7±2)^o$ (fig. 10). An amplification factor, with the account of losses in coaxial cables of the array feeding system, is 15.2 db. The beamwidth of an array fragment in a horizontal plane in interferometer channels for vertical polarization is similar to that in channels of an adaptive spatial filter, and that of horizontal polarization is $(130±20)^o$ in a vertical plane for both types of polarization the radiation pattern consists of nine difractional maximums with the lobe width of about 6^o and the general width of $(37±5)^o$ (fig. 11).

The presented array has been implemented in a passive small-base hyperbolic system for location of air objects that was developed at Lviv Research Radio Engineering Institute.

4.3. THE INTEGRATED PARABOLIC ANTENNA OF THE OBJECTS MONITORING SYSTEMS.

Creation of high-performance on-board monitoring systems for probing the surface of the Earth and the objects from height demands application of small-sized antennas and microwave frequency devices of signal processing.

Besides, it is expedient to use the same antenna with a single electrodynamic axis and spatial coordinates for both active and passive location.

The paraboloidal reflector antenna of the objects monitoring systems consists of a parabolic reflector itself, a number of feed horns of passive location channels, and a central feed of an active location channel. Radiometric channels operate at a wavelength of $\lambda \approx 8$ mm, and a radar channel works at a wavelength of $\lambda \approx 3$ mm.

To construct such monitoring systems antennas it is necessary to have an exact method for calculating radiation patterns for various feeds parameters and location of feeds. The paper considers parabolic antenna aperture excitation by the feed square aperture in a single Cartesian coordinate system. The feed aperture is assumed to be a square with the sizes $A_P \times B_P$ ($A_P > B_P$) (fig. 12). Distances from the centre of the feed aperture to a point of the phase centre in E-plane – R_E and similarly in H-plane – R_H are supposed to be known. Points on the feed aperture are considered in Cartesian coordinate system (*x, y, z*). The radiation field of the reflector circular aperture is written in spherical coordinate system (*r, φ, θ*).

In the method of aperture radiation analysis [13], [14] the feed aperture area is divided into elements of $dxdy$ size, radiation of which is expressed as a field of equivalent Huygens elements. A resultant radiation field of the feed in a given point on the reflector surface is written by surface integral of the feed aperture area. Such an algorithm allows to take account of both feed defocus, and nonconcurrency of beams, which can have a significant effect in a millimetric wave band.

The method of aperture radiation analysis has also been used to determine a radiation field of parabolic antenna aperture. However, it becomes necessary to calculate the phase of radiation field of the beam reflected by parabolic antenna aperture. Thus, the general formula for calculating nonnormalized radiation pattern of parabolic antenna looks like this:

$$f(\theta) = \int_0^{2\cdot\pi}\int_0^{R_0}\int_{\frac{-B_P}{2}}^{\frac{B_P}{2}}\int_{\frac{-A_P}{2}}^{\frac{A_P}{2}} \cos(\vartheta)\cdot\frac{1+\cos(\psi_P)}{\rho^* + \Delta l}\times$$
$$\times e^{-j\cdot\beta\cdot(\rho^*+\Delta l)}\cdot\cos\left(\frac{\pi\cdot y_{20}}{A_P}\right)\cdot e^{-j\cdot\frac{\pi}{\lambda_P}\cdot\frac{(y_{20})^2}{R_H}}\, e^{-j\cdot\frac{\pi}{\lambda_P}\cdot\frac{(x_{20})^2}{R_E}}\times$$
$$\times e^{+j\cdot\beta\cdot R'\cdot\sin(\theta)\cdot\cos(\phi-\Delta)}\cdot R'\,dy_{20}\,dx_{20}\,dR'\,d\phi \quad (13)$$

Field distribution in the feed aperture is described in Cartesian coordinate system (x_{20}, y_{20}) (fig. 12) by the function of amplitude distribution $\cos\left(\frac{\pi\cdot y_{20}}{A_P}\right)$ and by the functions of phase distribution: $e^{-j\cdot\frac{\pi}{\lambda_P}\cdot\frac{(y_{20})^2}{R_H}}$;

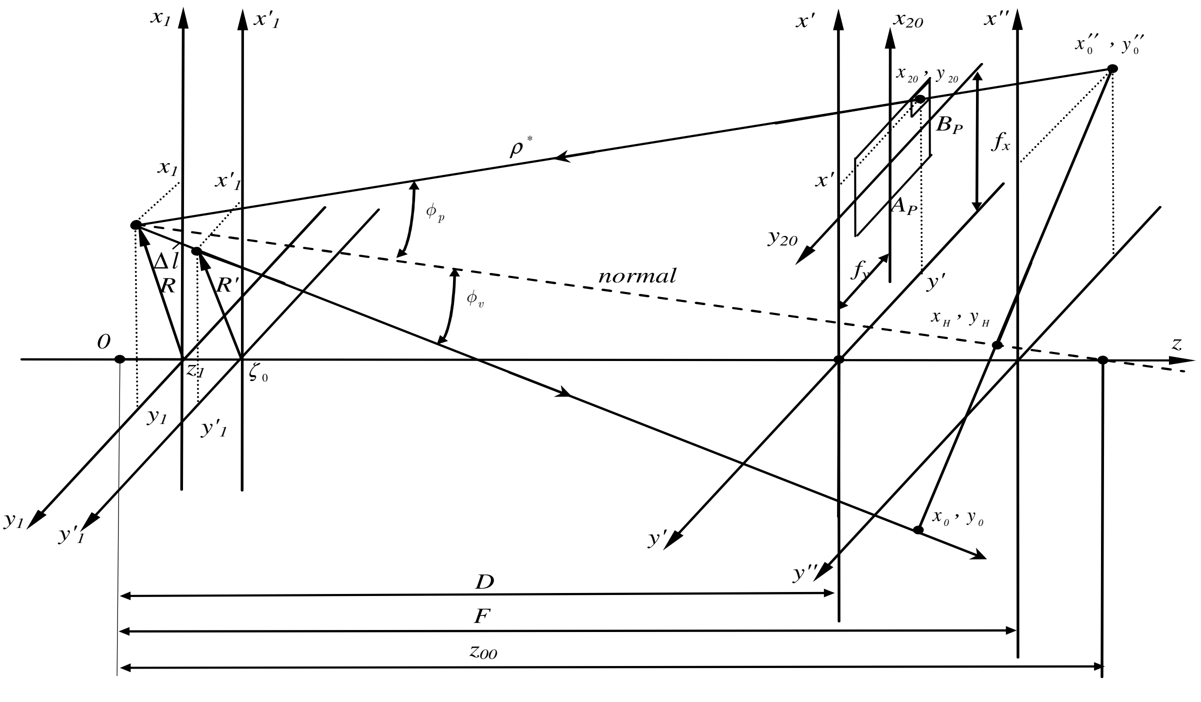

Fig. 12. Cartesian coordinate system for analysis of the parabolic reflector antenna with aperture feed defocus.

$e^{-j\cdot\frac{\pi}{\lambda_P}\cdot\frac{(x_{20})^2}{R_E}}$, where λ_P is the field wavelength in the feed aperture . The multiplier $(1+\cos(\psi_P))$ takes account of the radiation pattern of every Huygens element. The phase of radiation field in aperture radiation analysis is determined by distance ($\rho^* + \Delta l$), i.e. corresponding phase delay is equal to $\frac{2\cdot\pi}{\lambda_0}\cdot(\rho^* + \Delta l) = \beta\cdot(\rho^* + \Delta l)$, where λ_0 is free-space wavelength. An inclination angle of electric field vector to the parabolic antenna aperture plane is ϑ ($\cos(\vartheta) = \frac{\zeta}{\Delta l}$). This angle ϑ can have a noticeable effect if the feed is defocused. All other multipliers in expression (13) are standard during integration of a parabolic antenna aperture area [13], [14]. Angle ϕ is an angular coordinate in the plane of the parabolic antenna aperture, and angle Δ defines a direction of an electric field intensity vector relative to axis x. The direction of field beam radiated by Huygens element in the parabolic antenna aperture relative to a paraboloid axis (z), is given as θ.

Geometric parameters of beams in the parabolic antenna structure are founded in a single coordinate system (fig. 12). At this, transverse coordinates (x, y) for the feed aperture area are designated as (x_{20}, y_{20}), for the parabolic antenna aperture area – as (x_1', y_1') and for points on the reflector surface – as (x_1, y_1, z_1). Thus, according to fig. 12, we receive:

$$\rho^* = \sqrt{\left(x' - x_1\right)^2 + \left(y' - y_1\right)^2 + \left(D - z_1\right)^2}\ ; \quad (14)$$

$$\Delta l = \sqrt{\left(x_1' - x_1\right)^2 + \left(y_1' - y_1\right)^2 + \left(\zeta_0 - z_1\right)^2}\ ; \quad (15)$$

$$tg(\psi_P) = \frac{\sqrt{(x' - x_1)^2 + (y' - y_1)^2}}{D - z_1}. \quad (16)$$

However, to determine coordinates of a point in an antenna aperture (x_1', y_1', ζ_0) it is necessary to find coordinates of the point (x_0, y_0, F), where the beam reflected from the parabolic reflector passes the focal plane. An equation of the reflected beam is built on the basis of the known equations of an incident ray, an equation of normal to the parabolic reflector and assuming that $\phi_v = \phi_p$.

Using this numerical method, there were chosen such sizes of feed horns of the radiometric channel and their arrangement that can form 32 radiation patterns overlapping angle 31° between maximums of radiation pattern of extreme feeds and with the width of every radiation pattern in the altitude plane at level of 3dB: $2\theta_{0,5} = 1°$.

Fig. 13 presents normalized radiation pattern of the integrated parabolic antenna of the objects monitoring system for 16 feed horns of the radiometric channel – solid curves, and for the radar channel – a dashed curve.

Complexification of passive location systems with an active radar system in a single electrodynamic structure allows to form the radio thermal images of investigated objects, to determine coordinates of the objects and the distance to them.

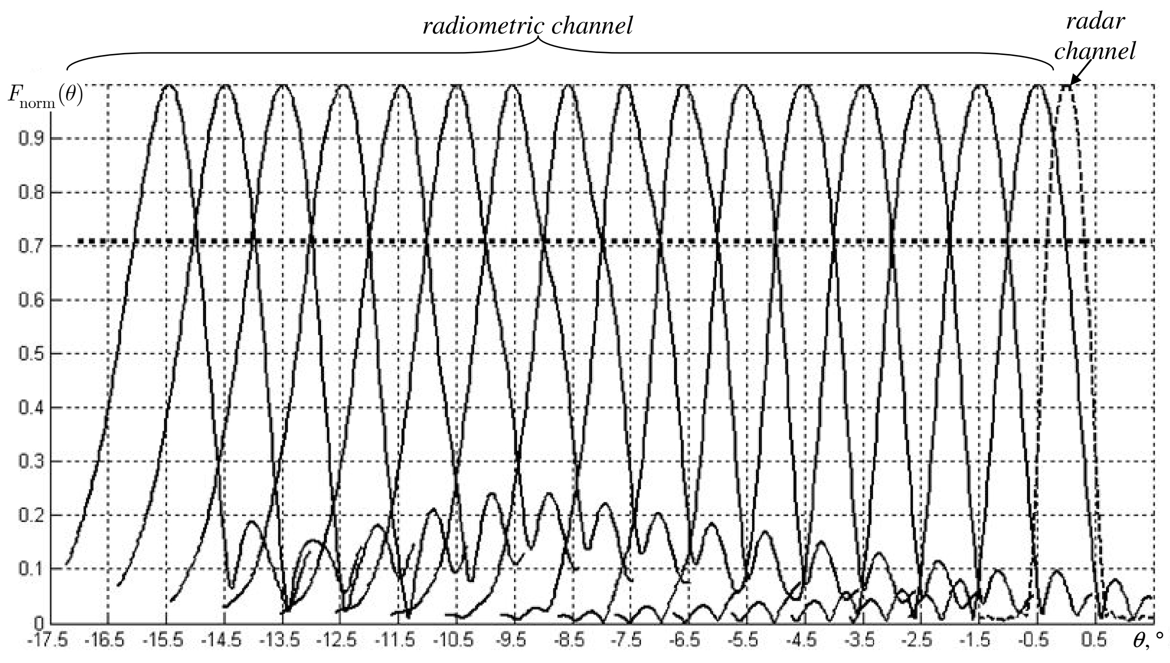

Fig. 13. Normalized radiation patterns of the integrated parabolic antenna of the objects monitoring system.

4. Integrated strip antennas-oscillator

Last time increased interest to integrated active antennas that were produced by planar process and are realizing in the way of functional and constructional integration semiconductor active elements with strip antennas. As a result of this kind of integration we get multifunctional knot that provide generating, emanation and accepting and strengthen of waves.

Also, antenna executes function of reactive element microwave circuit.

At the expense of nonlinearity characteristics active element antennas-oscillators (AO) realize transformation accepted electromagnetic vibration and execute synchronous detecting.

Inherently developed transistor antennas-oscillator included modify circuit the Clapp`s oscillator for microwave range that meet a claim of providing high stability frequency generated vibration in the best way and constructive-technologic method of group integrate technology.

Constructive integration strip antennas with microwave oscillator board can manage by two ways: planar and volumetric.

In planar performance radiator and topology of oscillator are located in the same board plane with joint screen. This is a kind of high-technology performance but in this case a process of tuning antennas-oscillator become complicated because of operator is situated in electromagnetic field radiation. Also, coefficient of area using antennas-oscillator is decreasing a bit.

In volumetric performance radiator and oscillator board are located in parallel planes which are separated by dielectric and screen and consequently volumetric integrate microwave construction realize.

On basis of this complex integration strip antennas and oscillator basic circuit were developed circuits and transistor antennas-oscillator constructions of centimetric range on different kind of antennas (print, vibrator, resonator or dielectric) which do function of reactive element microwave oscillator.

Topology of transistor antennas-oscillator planar performance with print resonator antenna in rectangular shape is characterized on fig.14.

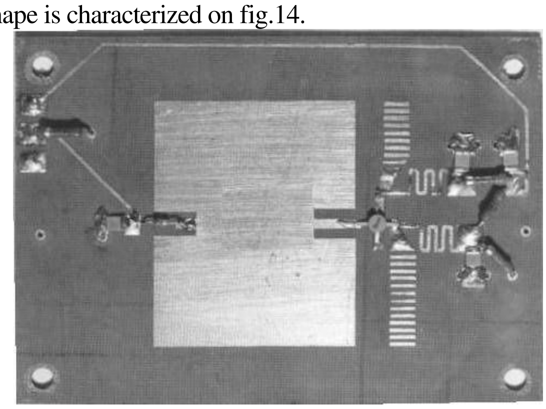

Fig. 14. Antennas-oscillator board planar type.

Volumetric (double side) antennas-oscillator construction on slot-resonator antenna is shown on fig. 15.

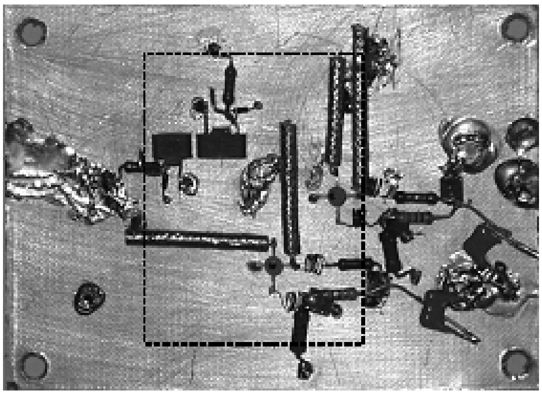

Fig. 15. Antennas-oscillator board volumetric type

Radiation power of antennas-oscillator define of transistor type and for 2T640A-2 is from 5 MW to 50 MW in depend on mode its work and working frequency define as sizes (resonance frequency) of radiator and as parameters of reactive elements antennas-oscillator. Developed antennas–oscillator on teflon are working on frequencies from 0,9 GHz to 6 GHz. Charts of direction and polarization characteristics define as type used antenna.

Significant parameter of antennas-oscillator as open self-oscillations system is stability of frequency generated vibration. It depends on conditions of waves` spread in space and its changing (e.g. in object placing) to lead to changing of frequency generated vibration at the expanse of changing incoming antenna's impedance. This property is useful in using antennas-oscillator in autodyne mode during realization radio waves sensor.

When conditions of spreading waves in free space are unchangeable instability frequency of vibration antennas-oscillator is $1 \cdot 10^{-3}/1 \cdot 10^{-4}$ in temperature vibration range from -20ºC to +50ºC.

For the purpose of increase stability frequency are developed a transistor antennas-oscillator with dielectric resonators (DR). DR connected with radiator in the way it not to radiate EMW and work in reflecting wave mode (tightening frequency), as high-durable resonator.

Topology AO with DR is shown on fig. 16.

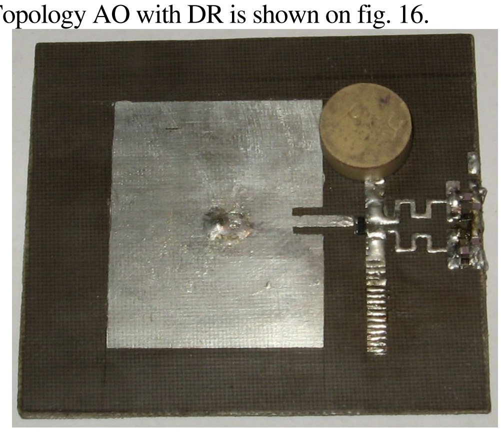

Fig. 16. Antennas-oscillators with dielectric resonators

The results of searching showed that using DR provides increase stability of frequency on rank above comparative with variants of AO that were considered early.

Last time specialists pay more heed to dielectric resonator antennas. Therefore, was developed and searched AO based on dielectric resonator antenna (DRA). Its topology is described lower on fig. 17.

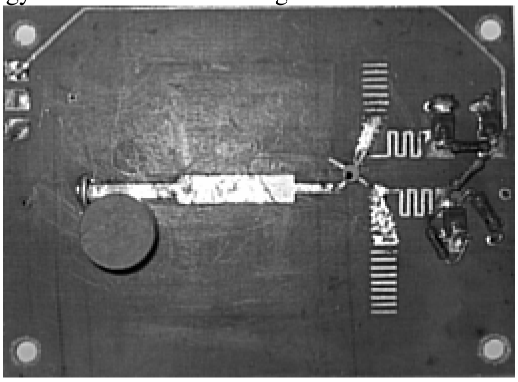

Fig. 17. Dielectric resonator antenna-oscillator

Dielectric AO (DAO) has spectral line width on level -30 dB about 15 kHz, that's in 4-5 times less than in AO on strip radiators without stabilizing DR.

Relative instability of frequency vibration DAO is $1{,}3\cdot10^{-5}$ in temperature changing in limits of 70ºC. Energy characteristics DAO are analogous to characteristics DAO on strip resonator antennas.

Construction AO on strip resonator radiators allow its following integration with detector microwave section. Based on this circuit was developed receiving-driving module with synchronic detector on fig. 14. Such microwave module has better sensitivity characteristics in comparing with autodyne sensor on AO.

Developed AO are using for realization radio-waves sensors protective and special system, also are perspective for construction active phased antennas grates.

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