INTRODUCTION

It is well known that antenna systems of airborne radars make substantial contribution into the radar cross sections (RCSs) of aircrafts [1]. Application of phased antenna arrays (PAAs) in modern radars poses the problem of detailed investigation of the scattering properties of such devices. Since modern PAAs are very complex devices, the most reliable data on their scattering properties can be obtained only by measuring the characteristics of real antennas. However, adequate electrodynamic models are especially valuable at the stages of the design and development of prototypes. As applied to the PAAs, the model must primarily allow reconstruction of the structure of the scattering pattern (SP) estimation of the maxima of diffraction lobes.

There are many studies devoted to calculation of the SPs of reflector antennas, because such antennas were used in most aircrafts. Usually, computational algorithms used the method of physical optics [2]. Modern aircrafts are equipped with the antenna systems containing PAAs and the field scattered by such antennas should be calculated with the use of combined methods taking into account the discrete character of the apertures of such antennas. A general technique for solution of the scattering problems for antennas consisting of periodic arrays of radiators was described in [3].

Modeling of the scattering by antenna devices with consideration for the radomes or shields has not been adequately studied in the literature. At the same time, formulation of this problem is important for lowering the radar visibility.

In this study, we consider construction of one of possible models that allow correct prediction of the lobe structure of the SP of an antenna array and take into account the influence of radome on the SP.

1. SOLUTION OF THE DIFFRACTION PROBLEM FOR PHASED ANTENNA ARRAYS

Formulation of the diffraction problem for a PAA with slot radiators is shown in Fig. 1. In a real electronically scanned array with slot radiators, the radiating panel consists of slots in a plane metal screen. The slots are loaded by a complex feeding system, which contains feeders, phase shifters, switches, and other devices. In the considered model of the PAA, the properties of the slot load are taken into account with the use of weakly reflecting screen (substrate) $S_{\text{sub}}$ (Fig. 1a), which is placed at distance $d$ behind the aperture. The electromagnetic properties of this screen are described by surface impedance $Z_{\text{sub}}$ (Selection of parameters $Z_{\text{sub}}$ and $d$ is considered below.). Let plane electromagnetic wave $\hat{E}_{\text{inc}}$ be incident onto the antenna from direction $(\theta_0, \phi_0)$ (Fig. 1b). It is necessary to find the field scattered into the far zone.

If antenna aperture $S_A$ has large wave dimensions, the scattered field can be calculated in the physical-optics approximation with integration of equivalent surface electric and magnetic currents, whose densities are assumed to be equal to those of the currents on an infinite grating illuminated by a plane wave, over surface $S_A$. This approach can readily be implemented in the absence of a radome. However, in the presence of a radome, the primary (incident) field on the antenna aperture is not plane. Hence, the solution to the model diffraction problem for equivalent currents on an infinite array cannot be directly used. In order to avoid these difficulties, we will use the following approach.

Let us introduce dummy surface $S_1$, which is placed at distance $d_1$ in front of the aperture. This surface is chosen so that, in the case of infinite dimensions of the array, the scattered field can be represented by the sum
of only propagating Floquet harmonics. Then, solving the model problem, we can determine partial reflection coefficients (RCs) for propagating Floquet harmonics in plane $S_1$ and, consequently, represent the total field as the sum of the incident field and a finite number of the plane waves of reflected harmonics. Then, in calculation of the PAA scattered field, currents should be integrated over dummy aperture $S_a$.

Thus, the first step in construction of this model is determining the RCs of each propagating Floquet harmonic in plane $S_1$. These coefficients are determined from solution of the diffraction problem for a plane electromagnetic wave incident onto an infinite grating whose design is identical to the model of the panel of the PAA radiators (see Fig. 1a). There is only a finite number of propagating harmonics. This number is determined by the ratio of the array period and the wavelength. The propagation coefficients of these harmonics along the $x$ axis, $y$ axis, and $z$ axis, respectively, are calculated from the formulas

$$\alpha_m = \frac{2\pi}{a} m + k_0 \sin \theta_0 \cos \varphi_0;$$
$$\beta_n = \frac{2\pi}{b} n + k_0 \sin \theta_0 \sin \varphi_0;$$
$$\gamma_{mn} = \sqrt{k_0^2 - \alpha_m^2 - \beta_n^2},$$

where $k_0 = 2\pi/\lambda$ is the propagation coefficient of the incident wave, $\lambda$ is the wavelength, $a$ and $b$ are the array periods along the $x$ and $y$ axes, and angles $\theta_0$ and $\varphi_0$ determine the direction of the incident primary field.

A harmonic is propagating in the case of simultaneous satisfaction of the following relationships:

$$\text{Re} \gamma_{mn} > 0, \quad \text{Im} \gamma_{mn} = 0.$$  

Diffraction of a plane wave by an infinite periodic system of slots was considered in many studies [4, 5]. For slots with a sufficiently complicated shape, the diffraction problem is usually solved with the method of integral equations (IEs). In this paper, we will use the IEs for the components of the density of the electric field over one period of a perforated metal surface. (A general technique for derivation of such equations is presented in [4].) The integral equations are solved numerically with the use of the Galerkin method and consideration for the weakly reflecting screen is performed with the method of long lines [6] at the stage of calculation of the IE kernels. The expressions for these kernels contain parameters $Z_{sub}$ and $d$.

Let us consider two principal polarizations of the incident wave, namely, parallel and perpendicular polarizations. We denote the RCs of propagating harmonics for the matched and cross polarizations, respectively, by $R_{mn\|}$, $R_{mn\perp}$, $R_{mn\perp}^c$, and $R_{mn\|}^c$.

Let us introduce equivalent surface currents in antenna aperture $S_a$:

$$\mathbf{j}_a = \mathbf{n} \times \mathbf{E}_{tot}; \quad \mathbf{M}_a = \mathbf{E}_{tot} \times \mathbf{n},$$

where $\mathbf{n}$ is the normal to the antenna surface and $\mathbf{E}_{tot}$ and $\mathbf{H}_{tot}$ are the total electric and magnetic fields. The total fields are the sums of incident fields $\mathbf{E}_{inc}$ and $\mathbf{H}_{inc}$ and reflected fields $\mathbf{E}_{refl}$ and $\mathbf{H}_{refl}$:

$$\mathbf{E}_{tot} = \mathbf{E}_{inc} + \mathbf{E}_{refl}; \quad \mathbf{H}_{tot} = \mathbf{H}_{inc} + \mathbf{H}_{refl}.$$