Radiation Pattern of Multibeam Array Antenna with Digital Beamforming for Stratospheric Communication System: Statistical Simulation

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This paper presents the results of the numerical simulation of a multibeam active phased array antenna for a High Altitude Platform Station (HAPS). The simulation takes into account the random errors caused by the non-identity of the array elements and the inaccuracy of the antenna calibration. The results of our statistical simulation show that the strict requirements on the sidelobe envelope for HAPSs can be met when the amplitude and phase distribution errors are minor, a condition which may be achieved by using digital beamforming.

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I. INTRODUCTION

The High Altitude Platform Station (HAPS) will be an essential component of future communication systems. Such a communication system consists of a stratospheric platform HAPS (including airships, communication payloads, additional mission payloads, etc.), user terminals, several gateways, and the ground facilities for TT&C (Telemetry, Tracking, and Command). The system will provide direct user services through several spot beams and links to terrestrial networks (PSDN, PSTN, WWW, etc.) through the gateways [1], [2]. The HAPS will be located at an altitude of 20.6 km above sea level and will remain nominally stationary within a sphere with a radius of approximately 0.5 km [3].

For S band transmissions, [4] introduced a passive transponder that reduces the consumption power of the payload; however, with such a transponder, the link margin would not be sufficient for direct communication between users. A ground station with a high gain antenna would be needed to connect users.

This paper presents a multibeam system with beamformers to obtain a sufficient link margin to communicate with users directly and to reduce interference between co-channel cells and between other channel cells. The most important component of a HAPS is the multibeam active phased array antenna (APAA) with high directivity. The HAPS uses the APAA for S-band and in the future will use it for Ka-band. Each channel of the APAA contains a microwave amplifier—a high power amplifier (HPA) in the transmitting channel or a low noise amplifier (LNA) in the receiving channel—and an analog/digital converter (ADC) or a digital/analog converter (DAC).

The antenna has to meet a number of strict requirements. Antenna beams must have a high electromagnetic isolation. This leads to a stringent limitation on off-axis antenna radiation. Especially, a low sidelobe level (SLL) is required when a Code Division Multiple Access (CDMA) standard is used.

To satisfy these requirements and to design the antenna with a low SLL, it is necessary to realize with adequate accuracy a certain taper amplitude distribution in the antenna aperture. Owing to fabrication errors and dispersion of microwave amplifier parameters, array elements are not absolutely identical. This leads to random errors in the amplitude and phase distribution over the antenna aperture. While the antenna gain and beamwidth exhibit minor variations due to small random disturbances of the distribution, the antenna SLL can grow significantly because of the same errors. In order to reduce the influence of fabrication errors as well as amplifier dispersion, the array antenna must be calibrated. The overall accuracy of the amplitude and phase aperture distribution is determined by a quantization in phase shifters and by the quality of the calibration method. The antenna multibeam radiation pattern can be realized by an analog or digital beamformer. For example, the Butler matrix is a well known analogous beamformer. In implementing a weighted summation of analog signals from many array elements for each of many beams, such a beamformer represents a network of interlaced microwave devices. For our investigation, we had to process the complex gain values of the array elements with high accuracy. (Suppressing the sidelobes of the HAPS antenna required amplitude errors of less than 0.2 dB and phase errors of less than 1.5°.) Imperfections in the microwave devices caused by reflections and insertion loss prevented the analogue beamformer from providing this accuracy. The only way to achieve the required accuracy was to use digital beamforming (DBF) in which signals from the array radiators were converted directly to a digital form and translated to the digital signal processor where all the antenna beams were formed independently.

This paper considers the influence on the APAA radiation pattern of the amplitude-and-phase aperture distribution errors that remain after calibration. The APAA is very a complicated system that will only be realized in the future; hence, physical modeling of the APAA for a HAPS was not possible. Our aim was to present an APAA simulation with a numerical statistical algorithm for probabilistic evaluation of the SLL. To estimate the feasibility of realizing the APAA system, we used this algorithm to determine the required overall antenna calibration accuracy.

II. ANTENNA GEOMETRY

Construction of the antenna takes into account the size of the area served (an angular sector of about ± 70°) and consists of a system of several plane panels (a multi-faced antenna) whose beams are properly joined. Each panel itself is a multibeam APAA with DBF, and the statistical method is valid for each panel.

The first version of the planar APAA to be considered here is shown schematically in Fig. 1. The antenna represents a rectangular array of 14 × 14 elements. The element spacing is $d_x = d_y = 0.5\lambda$ along x and y axes (these axes lie on the aperture plane). This antenna geometry is acceptable for the initial stage of the development of the stratospheric communication system [1].

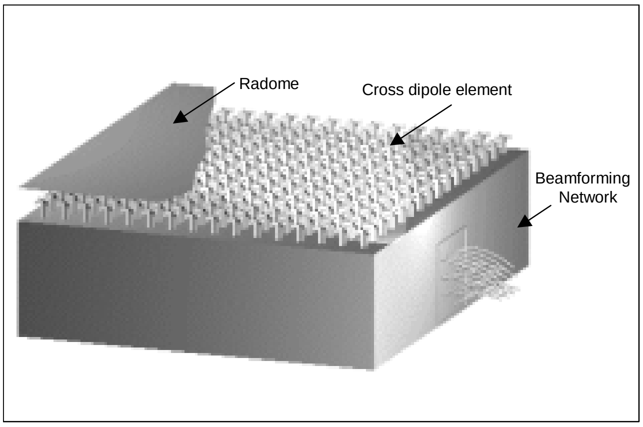

Fig. 1. General view of the first APAA version.

The radiators have a circular polarization. Each radiator represents a pair of crossed dipoles with a balun. The radiators, microwave amplifiers, and digital devices of the DBF form a monoblock. The inter-element connections are located inside the monoblock. The antenna is connected with a power supply and a control system by means of cables.

Radiators for phased array antennas are well known and are widely described in the literature. For example, radiators for

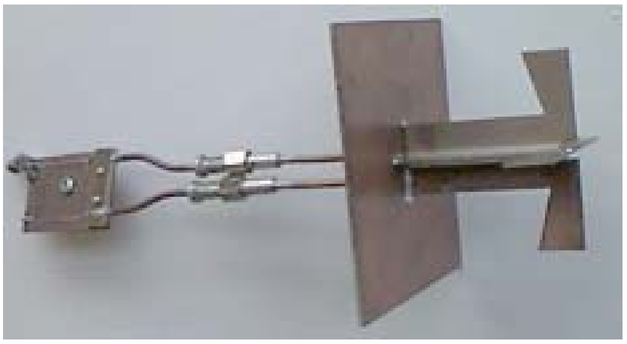

Fig. 2. The breadboard of radiating element for IMT-2000 frequency bands.

circular polarization are described in [5], [6]. The radiator contains two identical dipoles. Each one excites a linearly polarized wave. If radiators are placed in two orthogonal planes and the phase shift between two dipoles is 90°, a circularly polarized wave is formed. Figure 2 presents the cross-dipole breadboard.

For comparison, we consider the second version of the array antenna, which has 73 elements that are placed on a hexagonal grid so as to fill some circular aperture. The circular antenna aperture has a minimum SLL. Element spacing is d = 0.7 λ. Figure 3 shows the antenna geometry for the second version of the array antenna.

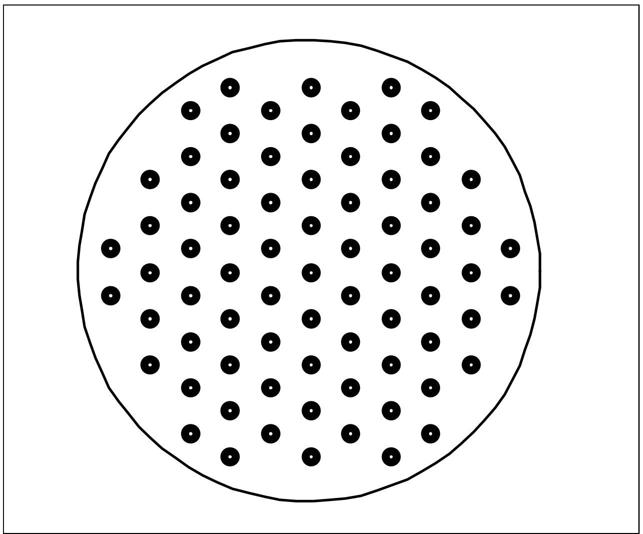

Fig. 3. Geometry of the second APAA version.

III. MAIN RELATIONS

In this section we define the notations and derive the main statistical relations for array antenna characteristics. On the basis of these relations, the numerical algorithm for statistical analysis of the antenna sidelobe envelope will be developed in section IV.

A field radiated by an APAA is the sum of the element fields; therefore its fluctuation (deviation from the average value) is determined by the error magnitude in each array channel. Usually, the number of elements in the antenna array is rather big, so errors in different channels fractionally countervail one another, and the total array field fluctuation may be much lower than the fluctuation of a field in each antenna channel. We will show that the reduction of the total array field dispersion in contrast with the dispersion in each channel is determined by the array gain.

Let's consider the array with radiators placed in points $\vec{r}_n$, $n=1...N$. Its radiated field is determined by the equation

$$E(\vec{v}) = \frac{f(\vec{v})}{A} \cdot \sum_{n=1}^{N} \{A_n \cdot \exp[ik \cdot (\vec{r}_n;\vec{v})]\}, \qquad (1)$$

where $f(\vec{v})$ is the radiator pattern, A_n is the amplitude of the n-th radiator, N is the total number of radiators in the array, $\vec{v}$ is the coordinate of the view point, $k = \frac{2\pi}{\lambda}$ is the wave number, and $A = \sum_{n=1}^{N} A_n$ is the normalized factor.

In practice, the greatest interest is on arrays having some center of symmetry, that is, the array structure is invariant to symmetrical transformation at this center. We confine our consideration to only such symmetrical arrays and accept that the coordinate system origin (the origin of vectors $\vec{r}_n$) coincides with the array symmetry center. As (1) reveals, when A_n is real, the array field is also real.

If there are random errors in antenna channels, the amplitude A_n is determined by the equation

$$A_n = A_n^0 \cdot \varepsilon_n = A_n^0 \cdot \rho_n \cdot e^{i\varphi_n}, \qquad (2)$$

where A_n^0 is the amplitude of the n-th radiator without errors, and ρ_n and φ_n are amplitude and phase errors, correspondingly.

Below, we take the normalized factor in (1) as $A = \sum_{n=1}^{N} A_n^0$.

Assume that amplitude (ρ_n) and phase (φ_n) errors in each antenna channel are independent, and owing to the calibration procedure, the phase errors have zero average values. For simplicity, we assume that a mutual coupling between array elements does not essentially change the element radiation pattern, the array factor, or consequently, the results of the simulation.

In addition, we assume that both ρ_n and φ_n have an identical distribution (regardless of channel number n) and the errors are independent from channel to channel.

To characterize the total channel error ε_n, we define its dispersion σ_ε as follows:

$$\langle \Delta\varepsilon_n \cdot \Delta\varepsilon_{n'}^* \rangle = \delta_{n,n'} \cdot \sigma_\varepsilon^2. \qquad (3)$$

Here, $\langle\ \rangle$ indicates the average of an ensemble of realizations, symbol * denotes a complex conjunction, and Δ is a deviation from the average value. The magnitude of σ_ε is determined by the dispersions of amplitude and phase errors ρ_n and φ_n and their distributions. When there are normal distributions and minor errors, the magnitude is described as follows: $\sigma_\varepsilon^2 = \sigma_\rho^2 + \sigma_\varphi^2$. The Kronecer delta symbol $\delta_{n,n'}$ shows that errors in different channels are uncorrelated.

Using (1)-(3), we can derive the formula for the average radiated power:

$$P_{av}(\vec{v}) = \langle E(\vec{v}) \cdot E(\vec{v})^* \rangle = \langle \varepsilon \rangle^2 \cdot P^0(\vec{v}) + \frac{\sigma_\varepsilon^2}{D} \cdot |f(\vec{v})|^2, \quad (4)$$

where $P^0(\vec{v})$ is the ideal power pattern for the case of no errors in aperture distribution, and $D = A^2 \Big/ \sum_{n=1}^{N} \left(A_n^0\right)^2 = \left(\sum_{n=1}^{N} A_n^0\right)^2 \Big/ \sum_{n=1}^{N} \left(A_n^0\right)^2$ is the array antenna directivity.

It is necessary to make several comments about (4) as follows:

First of all, (4) is similar to the well known equations of statistical antenna theory [5], [6], but it is valid for any distribution of random values ε_n (not only for normal distribution).

Eq. (4) shows that the average antenna pattern equals the sum of the ideal pattern without errors multiplied by $\langle \varepsilon \rangle^2$ and the random background $\frac{1}{D}\sigma_\varepsilon^2 |f(\vec{v})|^2$ determined by the dispersion of the random error in each channel, decreased by the value of the antenna directivity.

Eq. (4) is very convenient at the initial stage of antenna design because it allows us to estimate the required overall antenna calibration accuracy on the assumption of the desired antenna gain and required sidelobe envelope (and without any assumptions about antenna geometry, number of array elements, etc.).

Under the given assumptions, real and image components of the array field are uncorrelated, that is $\langle \mathrm{Re}\{\Delta E(\vec{v})\} \cdot \mathrm{Im}\{\Delta E(\vec{v})\} \rangle = 0$. This relation is helpful for further deriving the numerical algorithm.

IV. NUMERICAL ALGORITHM FOR STATISTICAL ANALYSIS

Here we present the numerical algorithm we will use for the statistical analysis of the antenna array sidelobes, which takes into account the aperture amplitude and phase distribution errors.

The algorithm includes an analysis of the ensemble of the random array patterns for the given maximum values of amplitude and phase distribution errors. Each pattern realization (1) is calculated by using N random amplitudes A_n determined by (2), where

$$\rho_n = 10^{B_0 \cdot \mu_n^a / 20}, \quad \varphi_n = \mu_n^\varphi \cdot \Phi_0 .$$

Random values $\mu_n^{a,\varphi}$ are independent and uniformly distributed over the interval [–1;+1]. Parameters B_0 and Φ_0 are maximum amplitude (dB) and phase (radian) errors, correspondingly.

We note the following about our choice of error distribution model. In practice, it is very difficult to evaluate the actual amplitude and phase errors distribution in antenna channels, but for our purpose it is not necessary to do this; we are interested in the overall array characteristics. The total array field equals the sum of a large number of independent random array radiator fields with identical properties. According to the central limit theorem of the probability theory, its distribution tends to a normal distribution regardless of the distribution in each antenna channel. Thus, our simple model of randomness is acceptable from an engineering perspective.

Further, in this paper, we consider the array pattern sections only. Thus, the vector $\vec{v} = [\sin\theta\cos\varphi, \sin\theta\cos\varphi, \cos\theta]^T$ depends on $\theta \in [-\pi/2; \pi/2]$ only ($\varphi = const$), and the argument $\vec{v}$ is replaced by θ below. The ideal element radiation pattern $f(\vec{v}) = \sqrt{\cos\theta}$ is taken into account.

The statistical evaluation of the average radiated power for N_l independent pattern realizations $E_l(\theta), l = 1...N_l$ is given by the equation

$$P_{av}(\theta) = \frac{1}{N_l} \sum_{l=1}^{N_l} |E_l(\theta)|^2 .$$

For the average fluctuation of power, we have the following expression:

$$P_d(\theta) = \sqrt{\frac{1}{N_l} \sum_{l=1}^{N_l} \left(P_{av}(\theta) - |E_l(\theta)|^2\right)^2} .$$

The upper sidelobe envelope is determined by the equation

$$P_{up}(\theta) = P_{av}(\theta) + P_d(\theta) .$$

Curve $P_{up}(\theta)$ has the following physical sense. As we mentioned above, the total array field $E(\theta)$ has a normal distribution. In our analysis, both the real and image part of the array field have normal distribution, and they are uncorrelated (see section III). Thus, the array field magnitude $R(\theta) = |E(\theta)|$ has a generalized Rayleigh distribution

$$\omega(R) = \frac{R}{\sigma^2} \exp\left\{-\frac{R^2 + b^2}{\sigma^2}\right\} \cdot I_0\left(\frac{bR}{\sigma^2}\right), \quad (5)$$

where $b = \sqrt{[\mathrm{Re}\langle E(\theta)\rangle]^2 + [\mathrm{Im}\langle E(\theta)\rangle]^2} = |\langle E(\theta)\rangle| = \langle R \rangle$

and I_0 is the modified Bessel function of zero kind.

The level $\langle R^2 \rangle + \sqrt{\langle (\Delta R^2)^2 \rangle}$ corresponds to $P_{up}(\theta)$ for arbitrary direction θ.

The probability that antenna power radiated in direction θ will not exceed the level $P_{up}(\theta)$, following from the law in (5), is equal to approximately 0.85.

The numerical evaluation of this probability for the ensemble of 2000 realizations gives the value 0.83–0.87 depending on the level of ideal power pattern $P^0(\theta)$.

Thus, the area below the curve $P_{up}(\theta)$ represents the domain where, at a given range of amplitude and phase errors, a random implementation of $P_l(\theta)$ is allocated. On a positional relationship of the curve $P_{up}(\theta)$ and the curve of the required sidelobe envelope, it is possible to decide whether the antenna meets the sidelobe radiation requirements.

The level $P_{up}(\theta)$ essentially depends on the maximum values of the amplitude and phase errors. Parameters B_0 and Φ_0, for which $P_{up}(\theta)$ is allocated below the required sidelobes envelope in the given beam steering range (or not too much above it), determine the necessary antenna calibration accuracy.

V. SIMULATION

In this section, two examples of statistical simulation of APAAs are considered.

The array in each example has an identical beamwidth (12°) and gain (approximately 23 dB). The number of elements, the element gain, and aperture distributions in both cases were chosen to obtain the same antenna parameters and satisfy the required reference radiation pattern exhibited in [2]. For each example of simulation, 2000 random samples in the ensemble were used.

The first APAA version

The amplitude distribution constitutes a tensor product of two amplitude distributions strongly tapered over x and y directions: $A_{nm} = A_n^x \cdot A_m^y$. Each distribution A_n^x and A_m^y corresponds to the $cos^2 x$ function. Simulation results of the array antenna pattern are shown in Fig. 4 for one major plane. The results for the other major plane are similar. The sidelobe level in diagonal planes is smaller as is known from antennas having a rectangular aperture.

Figure 4 presents three cases: an ideal radiation pattern for an array without random errors (Fig. 4(a)) and average patterns and upper sidelobe envelopes for two different sets of values, B_0 and Φ_0 (Figs. 4(b) and 4(c)). The region hatched is the domain where a random pattern realization is placed with a high probability. In addition, the figures show the curve of the random background (from (4)) and the worst realization with the maximum SLL.

Choosing the values of B_0 and Φ_0, we supposed that digital beamforming was used and that phase errors introduced the main contribution to the degradation of the antenna radiation pattern.

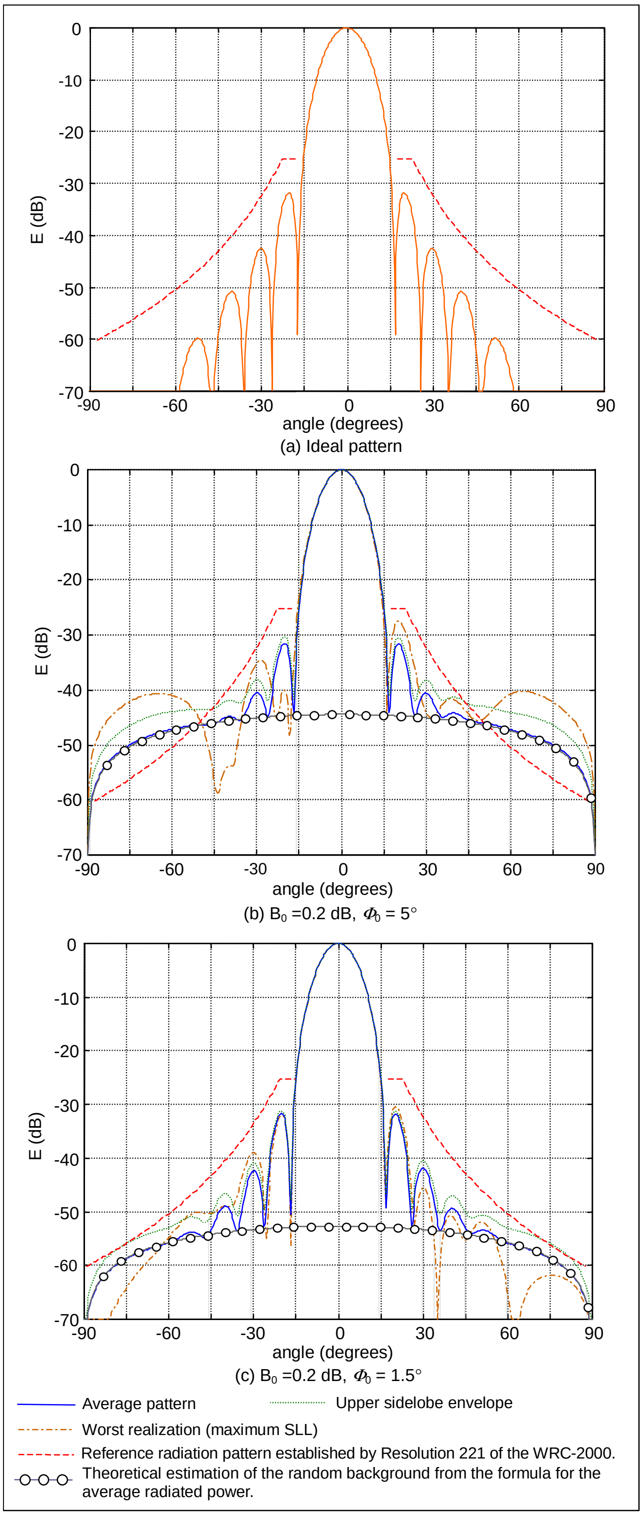

Fig. 4. Simulation patterns of the 14×14 element array.

The second APAA version

In this version, we selected a tapered amplitude distribution which has an axial symmetry and can be calculated according to the expression

$$A_n^0 = 0.58 + 0.42 \cdot \cos\left(\pi \frac{R_n}{R_{\max}}\right),$$

where R_n is the distance between the circle center and the n-th element $R_{\max} = \max_n R_n$. This distribution gives enough high aperture efficiency under the low sidelobe level. It provides satisfactory performances by using a smaller number of the antenna elements than the first antenna version.

Simulation results for this example are shown in Fig. 5 for one major plane. The results for other planes are similar.

Figures 4 and 5 demonstrate that if amplitude and phase errors are equal to 0.2 dB and 1.5° correspondingly, antenna performances are similar in both cases despite an essential difference in the quantity of radiating elements.

The explanation of this fact is as follows. In the first APAA, we used a very tapered amplitude distribution to obtain low aperture sidelobes, and in this case, the antenna had a low efficiency. In the second APAA, a weaker tapering of the amplitude distribution was used, and aperture sidelobe level was higher (without taking into account random errors). In addition, the aperture form of the second antenna was nearer to a circle. The sidelobe level of a circle aperture is lower than a rectangular aperture with the same tapering. The antenna gain was equal in both versions:

$$G\,(\text{dBi}) = 10 \log N + G_r(\text{dBi}) + L_1 + L_2,$$

where N is the total number of the radiators, G_r is the antenna element gain, L_1 is the efficiency determined by the amplitude taper for the chosen aperture geometry, and L_2 is the efficiency determined by random errors. In the first version we have $N = 196$, $G_r = 5$ dBi, and $L_1 = -4.2$ dB. Therefore, $G = 23.5$ dBi. In the second version, we obtained $N = 73$, $G_r = 6.9$ dBi, $L_1 = -1.7$ dB, and $G = 23.6$ dBi. In both cases $L_2 = -0.2$ dB for the same model of errors. The values of the element gain G_r are different in these arrays when we take into account the element spacing and mutual coupling radiators. Thus, the gain and beamwidth in both antennas are nearly the same. However, we determined the final sidelobe level mainly by the amplitude and phase errors. According to (4), the background of the sidelobes is determined by random errors, and it does not directly depend upon the total element number. This background level is connected with the directivity. We assumed that losses in the APAA are small and a value of the directivity coincides with the gain. Therefore, the use of very tapered amplitude distribution is not effective for these purposes.

Note that the APAA which meets the requirements on amplitude and phase errors and service conditions does not yet

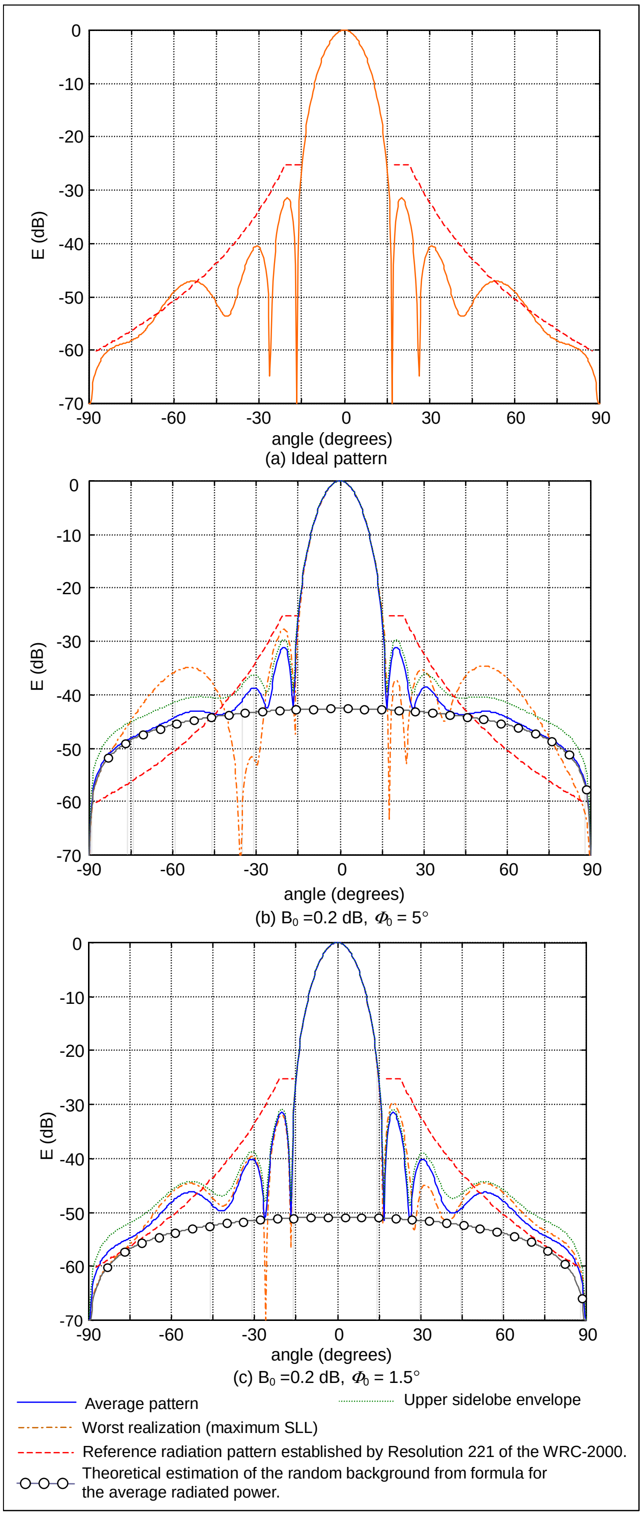

Fig. 5. Simulation patterns of 73 element array.

exist. The parameters of contemporary microwave amplifiers can provide a phase non-identity within a limit of 10°. Digital beamforming permits the compensation of a dispersion of APAA channel parameters with high precision (about 0.1° on phase and 0.1 dB on amplitude). To accomplish this, it is necessary to have a permanent precise calibration of array channels in order to insert phase errors into the processor memory. The requirements are near the accessible limit, which is reached by using modern methods of array antenna calibration [8]-[10]. A method of satellite antenna calibration with an RMS of about 1° is described in [9]. A feature of the stratospheric platform environment is that the angular fluctuations are due to blasts. Special optic-and-radio equipment with high operating speed should be developed to carry out permanent antenna calibration during APAA service.

All the results given here are for a single boresight beam. Owing to the directional properties of an array element, the steered antenna beam has a decreased gain and increased sidelobes according to the particular element pattern. However, if antenna calibration is performed for every beam of the antenna radiation pattern, the sidelobe component that is caused by the random errors could be the same as in case of the boresight beam. The digital beamformer as well as the calibration procedure permits this possibility.

VI. CONCLUSION

We developed an algorithm of statistical SLL envelope simulation for the HAPS antenna and obtained the tolerances of the antenna calibration. The implemented calculations show that rigorous requirements for determining array channel parameters are needed in order to realize the APAA for the Stratospheric Communication System. The admissible values of the amplitude and phase errors are near the limit of the modern calibration methods for phased array antennas.

It is essential that new approaches and technologies be developed to create an antenna for the Stratospheric Communication System.

We think that the creation of an experimental simplified communication system would be a good step toward the development of the HAPS. A similar approach is already under implementation in Japan [11]. The experimental system would be the basis for further development and refining of both antennas and other equipment.

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