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STEERABLE REFLECTIVE HIGH-GAIN ANTENNA ARRAY BASED ON LOADED DIPOLE SCATTERERS

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We present the results of studying experimentally a constructively simple and inexpensive centimeter-wave antenna having a radiation pattern, whose shape can be controlled. Specifically, the main lobe of the pattern can be scanned in a wide angular range, its width can be changed, and minima of the pattern can be formed in specified directions. The feed being the only active element of the antenna is located in front of the mirror made up by a system of steerable passive scatterers, which are dipole antennas loaded by semiconducting diodes. The capacity of the diode loads is changed under the action of controlling signals, which set specific bias voltages at each diode. The studies aimed at testing experimentally the operability of the antenna and assessing the practically achievable parameters of the proposed antenna design. The developed laboratory model of the antenna operates at frequencies near 2.4 GHz with a gain of more than 21 dBi and the possibility of scanning the main lobe of the radiation pattern in the horizontal and vertical planes to $\pm 60^\circ$ and $\pm 15^\circ$, respectively, for the antenna with overall dimensions $100 \times 60 \times 30$ cm.

1. INTRODUCTION

High-gain antennas, whose radiation and transmission direction can be controlled electronically, seem to be rather promising from the viewpoint of application in up-to-date wireless communication systems. Such antennas allow one to increase the communication range even in cases, where the relative position of the source and the radiation receiver changes with time and make it possible, specifically:

— to establish continuous stable connection of the desired quality with mobile objects;

— to compensate for mechanical deviations of the antenna structures, which can be caused by different undesirable factors (wind, collapse of support structures, etc.).

— to ensure space-time separation of the communication channels;

— to reconfigure promptly the topology of a network consisting of many nodes.

If an antenna system makes it possible to use electric controls to set not only the direction of the maximum of the radiation power or the radiation pattern, but also the directions of the minimum of the emitted power (reception efficiency), then it becomes possible to eliminate noise, which also allows one to improve the quality of the communication via the wireless network.

However, despite all the valuable properties, which have been mentioned above, the antennas with electronically controlled radiation patterns so far have not been used widely in wireless communication systems available in the mass market of telecommunication equipment (Wi-Fi, WiMAX, 3G networks, etc.). There have been only individual attempts at creating commercial antenna systems based on such solutions, for example, one proposed by ViVATO [1, 2]. Mainly, they are related to the use of antennas with controlled diagrams on the side of base stations. Cases of using such antennas on the client side are virtually unknown.

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The main reason for this situation is the great cost of such devices, which is connected with a high price of microwave elements used in most up-to-date antenna systems with controlled radiation patterns, which are based on the use of phased antenna arrays (PARs).

It is known that to obtain high gains of antenna arrays, one needs an aperture having a great electrical size and, correspondingly, a great number of antenna radiators distributed over this aperture. In accordance with the most frequently used concept of the PAR architecture, a separate waveguide feeds each radiator with microwave signals received from the source. In this case, the channel of each radiator contains a separate phase control unit, i.e., phase rotator. Despite the advance in the development of microwave materials and the technology of designing the corresponding units, it is still rather expensive to produce great numbers of individual microwave circuits. The above-said factors are augmented by the difficulties of the arrangement of PAR elements, as well as the power loss in long waveguide lines. Compensation for the loss requires the use of amplifiers, which makes the final cost of the antenna array even higher.

It is possible to minimize the required number of expensive microwave elements, reduce the power loss, and lower the cost of the antenna system significantly while retaining the advantages of the antennas with electronically controlled radiation patterns in the framework of the concept of antenna arrays with spatial feeding of its elements being passive scatterers with controlled parameters. Changes in the parameters of the scatterers allow one to vary the phase and amplitude of the fields, which are reradiated by them, which makes it possible to form the required radiation pattern of the entire set of scatterers.

The development and studies of this type of antennas started rather long ago [3–7]. Different versions of the antenna design were proposed for different wavelength ranges.

For centimeter- and millimeter-wavelength ranges, the so-called reflective antenna arrays, i.e., antennas formed by a flat reflecting surface [3–6], were developed. A great number of passive reflector antennas are installed on this surface (as a rule, they are microstrip antennas with different designs). Microwave transmission lines are not connected to these antennas. The reflector antennas should reradiate the field, which is incident on them, with the phase required to ensure that the entire system forms a plane wave front propagating in the specified direction in the far zone. The possibility of adjusting the phase of the field reradiated by the field scatterer by means of changing its parameters allows one to control the radiation pattern of the entire system. The parameters of scatterers in the reflector arrays are changed by means of changing the parameters of the controlling elements, of which they are comprised. In this case, the steerable elements can be PIN diodes, which allow one to switch the system elements on and off discretely, thus controlling discretely the phase of the reradiated field, as well as variable-capacitance diodes, which provide the possibility to adjust the load parameters smoothly when the bias voltage is changed, thus allowing one to tune the phase of the rereflected field gradually.

Approximately at the same time when the first works about the reflector antenna arrays were published, an antenna was proposed, which consisted of steerable scattering elements installed in space around the active antenna feed, rather than on a plane. In this case, as in the case of reflector antenna arrays, the change in the parameters of the scatterers allows one to form the phase of secondary fields which ensure the required radiation pattern of the entire system.

Such antenna arrays with steerable scatterers were called ESPAR antennas (Electronically Steerable Parasitic Array Radiators). Most ESPAR antennas [8–13] are based on the scheme proposed in early works [14–16]: a spike antenna is situated over a metal plane and fed via a coaxial cable. This antenna is surrounded by metal rods connected to the metal plane via diode loads. The bias voltage applied to the diodes is determined by the load capacity. The size of the rods is approximately equal to a quarter of the operating wavelength. As a rule, the diameter of the circle formed by the scatterers does not exceed the operating wavelength. Therefore, the amplification gains of the ESPAR antennas, which were studied earlier, rarely reach 10 dBi and range from 5 dBi [17] to 9 dBi [12] on the average. An appealing feature of the antennas with space arrangement of the scatterers is greater development flexibility as compared with the case of planar arrangement, which allows one to develop antennas with unique parameters. An example of this development is ESPAR antennas which allow one to perform circular scanning of the radiation pattern maximum.

The simplicity of the design and manufacture of the above-described antennas, as well as the absence of costly microwave elements in them, are their extremely appealing features. However, for quite a long time such antenna arrays were significantly less widespread than the classic PARs [18, 19]. The reason for that was insufficient refinement of the design of such antennas, which was due to the complexity of their analysis and optimization, as well as with the necessity of creating non-standard technical and algorithmic solutions to control such antenna arrays. Insufficient development of computer technology and microelectronics made the developers prefer the classical PAR schemes.

The situation changed only in the recent years. The fast advance in the functionality and accessibility of computing machinery allowed one to perform series of numerical experiments, which were required to solve the problems of design optimization for the antennas based on scatterer arrays, as well as create control systems for these antennas where easily accessible microprocessors are used. The novelty of the antenna systems using steerable scatterers actualizes both the creation of such antennas for specific applications (primarily, in the field of wireless communication systems, evidently) and the development of effective methods for their theoretical and experimental analysis.

In this paper, we propose the antenna design which ensures a high gain and a steerable radiation pattern. The antenna was manufactured with the use of the scattering dipole elements, which are typically used in ESPAR and are located in space separately. They form an electrically big mirror, which is typical of reflector antenna arrays. The parameters of the scatterer are varied by changing the impedance of the semiconducting diode connected to the scatterer. Actually, in this case, the functions of the radiator and phase rotator are combined in one simple and inexpensive element. The antenna design which is proposed in this paper provides ample opportunities for varying the spatial positions of the scatterers, thus changing the shape of the mirror, the number of the scatterers that comprise it, and the distance between the scatterers. The proposed design also permits easy replacement of the scatterers. The above-described distinctive features of the antenna design, when they are combined with the existing possibilities of the numerical simulation, offer wide possibilities of optimizing the antenna parameters to obtain the desired characteristics. The developed antenna is easy to manufacture and inexpensive.

The developed laboratory model of the antenna operates at the frequencies near 2.4 GHz with the gain exceeding 21 dBi. This allows one to scan the main lobe of the radiation pattern in the horizontal and vertical planes to $\pm 60°$ and $\pm 15°$, respectively, for the the antenna with overall dimensions $100 \times 60 \times 30$ cm.

2. ANTENNA DESIGN

Figure 1 shows the conceptual diagram of the antenna under consideration. The antenna consists of a mirror, which is formed by steerable scatterers, and a radiator. Electric dipoles loaded at the center with a capacitor impedance whose value can vary are chosen as steerable scatterers. Varying of the load impedance allows one to tune the phase of the wave scattered by the oscillator. Simultaneously, the amplitude of the scattered field changes. The proposed antenna design (in which the scatterers are located in space, and not on a plane) makes it possible to vary the relative positions arbitrarily. It simplifies the tuning of the antenna and expands the possibilities of optimizing its structure in order to obtain these or those characteristics.

Unfortunately, when the load impedance changes, the current amplitude changes simultaneously with the change in the current phase, which does not allow one to use the scatterer for generation of the secondary radiation with an arbitrary phase. To achieve phase tuning in the 360° range, one should use a combination of several layers of scatterers (Figure 1 shows the three-layer structure). In this case, the propagation of the exciting and scattered waves between the layers ensures the required continuous phase shift, which is added to the phase shift ensured by the variation of the load impedance. The scatterers which form the mirror are not independent: since strong electrodynamic coupling exists between them. Therefore, the scattering properties of each mirror element are determined not only by the load of this element, but also by the loads of the adjacent scatterers.

Practical implementation of the above-described antenna concept requires, first of all, choosing of

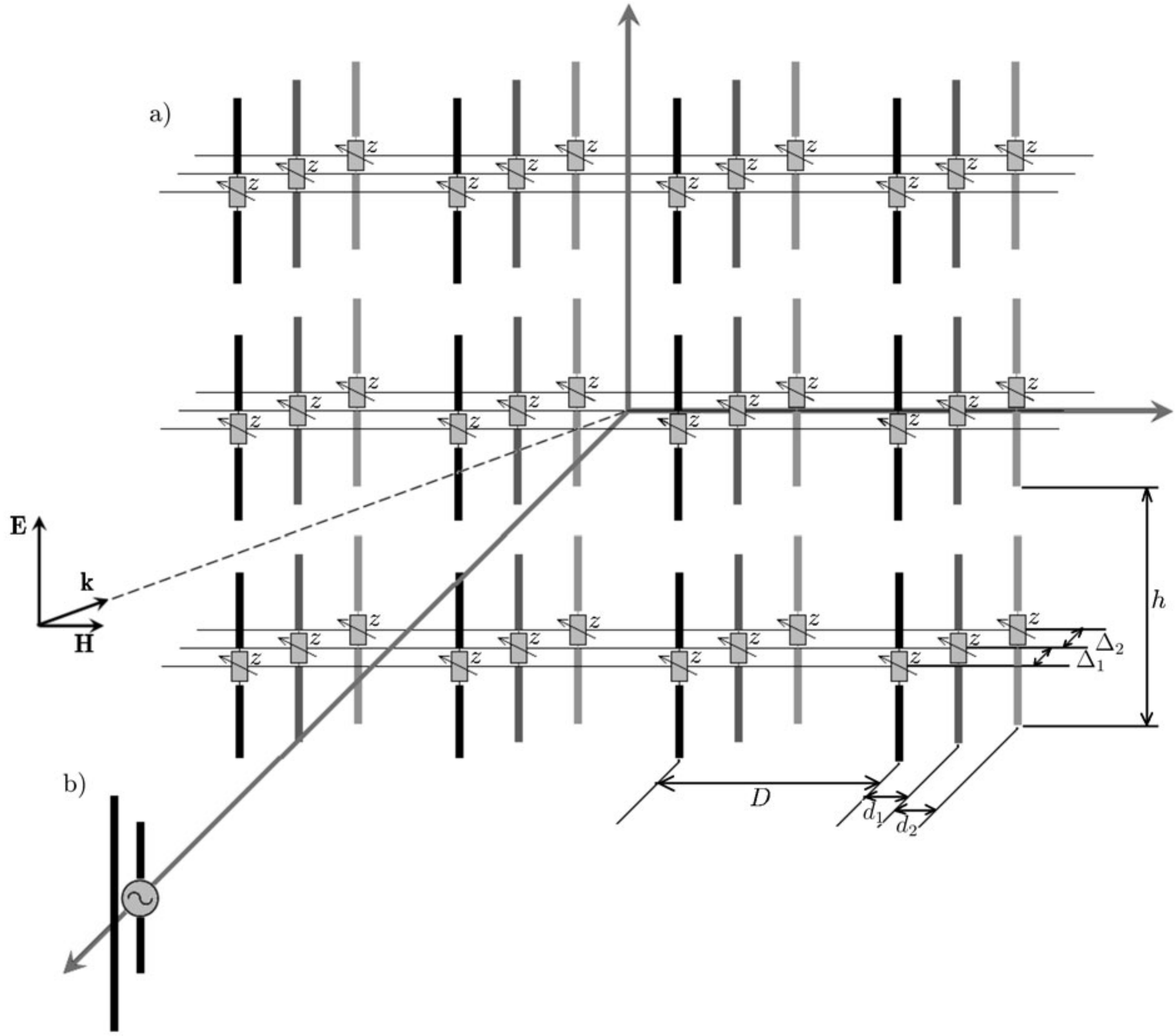

Fig. 1. Concept scheme of the antenna under consideration: the three-layer mirror of the antenna (*a*) and the radiator (*b*).

a specific design of the scatterer, i.e., the electric dipole, as well as determining of the distance between the scatterers (see Fig. 1). Additionally, one should determine the structure of the radiator and its position relative to the scatterers. The possibility of changing many parameters even in the framework of the antenna concept, which has been chosen already, and the corresponding wide range of the changes in its parameters results in the necessity of defining the requirements imposed on the antenna concretely. In this paper, as this requirement, we chose the approximation of the parameters of the antenna under consideration to the parameters of directed antennas (parabolic antenna, Yagi-Uda antenna, etc.), which are widely used currently to organize long-range wireless communication channels, specifically, the achievement of a gain of 18–24 dBi for frequencies near 2.4 GHz. In this case, an additional requirement imposed on the antenna under consideration is the possibility to maintain a high antenna gain in the angle range $\pm 60^\circ$ in the horizontal plane during the tuning of the radiation pattern. The dimensions of the antenna have been chosen to be similar to the dimensions of the parabolic antenna with a radiation gain of 24 dBi, namely, 100×60 cm. In the process of the experiments, the parabolic antenna with these dimensions was used as the reference one.

The design of the scatterer used in the experimental model of the antenna is shown in Fig. 2*a*. The scatterer is a single-sided printed board and consists of dipole branches (*1*), an impedance transformer being a long line (*a*), a variable-capacitance (vc) diode (*3*) connected to the long line, shunt chokes (*4*) separating

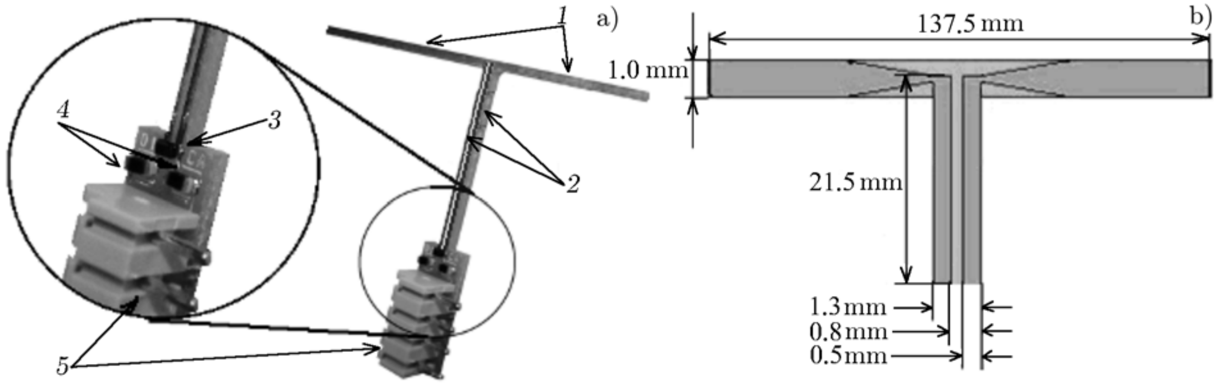

Fig. 2. Scheme of the scatterer (*a*) and its geometric dimensions (*b*): dipole *1*, impedance transformer *2* (long line), variable-capacitance diode *3*, shunt chokes *4*, plug connected to the controlled line *5*.

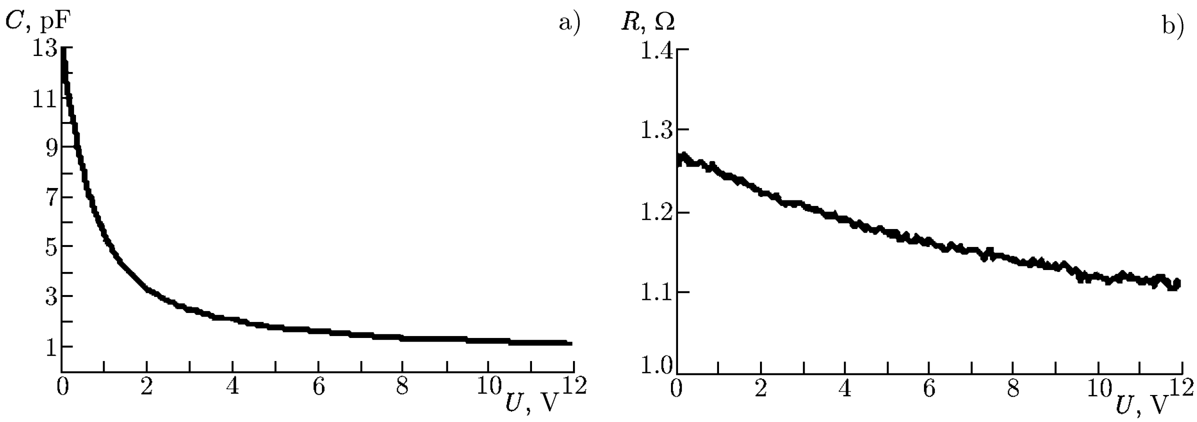

Fig. 3. Experimental characteristics of the vc diode: dependences of the capacity C on the bias voltage U (*a*) and of the active resistance R on the bias voltage (*b*).

the high-voltage part of the scatterers and the control lines (*5*), which feed the bias voltage to the vc diode, and a contact unit connecting the control lines to the antenna control system. The long line (impedance transformer) is introduced to the structure to widen the range of changes in the load impedance at the dipole input.

The geometric parameters of the scatterer were chosen basing on the data of the numerical simulation performed in the HFSS software environment. It was assumed that the parameters of the scatterer material coincided with the parameters of foil-clad textolite FR4. The thicknesses of the dielectric substrate and the metal were 1.5 and 0.05 mm, respectively, and the dielectric permittivity and the loss tangent of textolite are 4.9 and 0.03, respectively.

The role of the controllable load of the scatterer was played by the MA4ST1240 ODS1279 vc diode with the following parameters claimed by the manufacturer: when the bias voltage changed in the 0–12 V range, the vc diode capacity changed from 13 to 1.1 pF, and the active resistance changed from 1.2 to 1.1 Ohm. Figure 3 presents the results of studying the vc diode parameters experimentally, namely, the dependences of the capacity C and active resistance R (Figs. 3a and 3b, respectively) on the bias voltage U. We chose the ILC 0603 throttles with an inductance of 68 nH$\pm$10% as the elements through which the control line was connected to the scatterer.

The numerical experiments were aimed at determining the optimal geometric dimensions of the scatterer (including the parameters of the long line, i.e., the impedance transformer). The problem of searching for the optimal dimensions of the scatterer under conditions of its electrodynamic coupling with the rest scatterers that form the mirror was solved nonstrictly with additional assumptions made. To

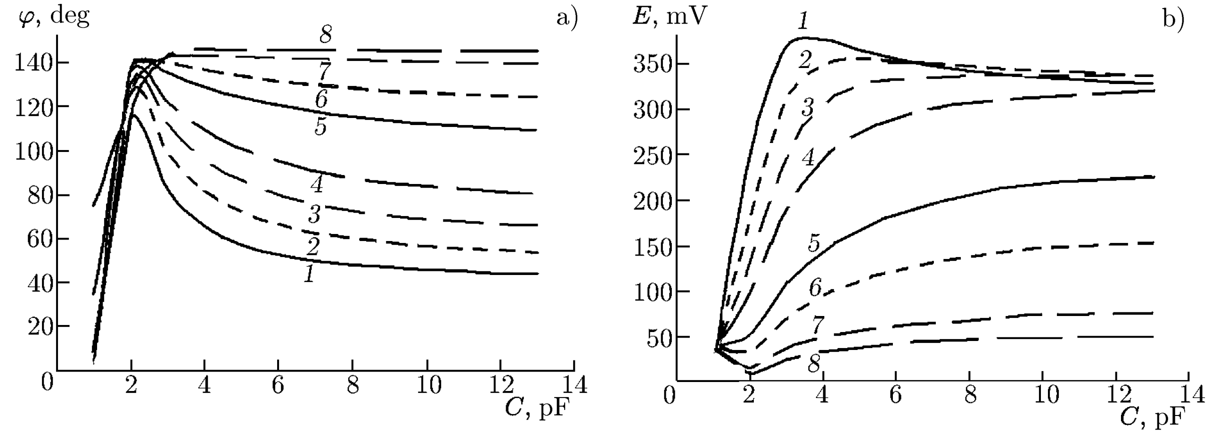

Fig. 4. Phase and amplitude characteristics (*a* and *b*, respectively) of the scatterers for different lengths of the dipole branches: curves *1* correspond to the dipole branch length equal to 1.15λ, *2* to 1.1λ, *3* to 1.05λ, *4* to 1.0λ, *5* to 0.9λ, *6* to 0.8λ, *7* to 0.6λ, and *8* to 0.5λ.

simulate approximately the phase and amplitude of the parameters of the scatterer operating as a mirror component, we studied a set of 17 scatterers situated at distance less than half of the wavelength λ from each other (0.48λ). The series was excited by a plane wave with the electric-field vector being parallel to the axis of the scatterers and the wave vector being perpendicular to the plane of the set. We studied the dependence of the amplitude and phase of the reradiated field in the far zone of the array (at the maximum of the radiation pattern of this set of scatterers) on the load capacities. In this case, the values of the capacity were set equal for all elements of the set.

This approach to determination of the properties of the scatterer operating in a real mirror does not allow for the mutual influence of the elements located in adjacent layers and rows, as well as the influence of the changes in the load impedances on the parameters of the mirror. With the state-of-the-art computational potential, the account for the above-said factors requires too much time to solve the problem of optimization of the scatterer geometry. The proposed approach can be justified by the fact that the main influence on the operation of each scatterer, which forms the mirror, is exerted by its immediate neighbors. In this case, it is logical to assume that the neighboring scatterers in the mirror have almost equal load impedances, which is necessary to ensure continuity of the changes in the current phases in the scatterers along the mirror. The geometric dimensions of the long line, i.e., the impedance transformer, are shown in Fig. 2*b*. They were found basing on the results of an individual series of numerical experiments aimed at finding such dimensions of the long line, which ensure the maximum range of the load impedance tuning.

Figure 4 shows the phase and amplitude parameters of the scatterers calculated within the framework of the above-described model. The plots labeled *1–8* correspond to different lengths of the scatterer arms. The maximum tuning range of the reflected-field phase corresponds to the arm length being equal 1.1 of the wavelength in free space and amounts to 110°. In this case, the amplitude of the rereflected field reaches maximum values, and the range of its variations amounts to less than 2 dB for the changes in the load capacity from 2 to 13 pF.

Basing on the obtained results of the numerical experiment, the scatterer having the geometric dimensions shown in Fig. 2 was chosen as an antenna-forming element.

At the next step of the numerical simulation, we studied the antenna based on the scheme shown in Fig. 1 and comprised by the scatterers with the above-specified parameters. This step was aimed at determining the distances between the scatterers, which would allows one to achieve the required values of the antenna gain and the tuning range of the main maximum of the radiation pattern. For each fixed location of the scatterers, we determined the capacities of the diodes, which would ensure the maximum of the antenna gain in several angles in the above-said range. To find the distributions of the capacities for different directions of the wave incidence, we solved the problem of multi-dimensional optimization using

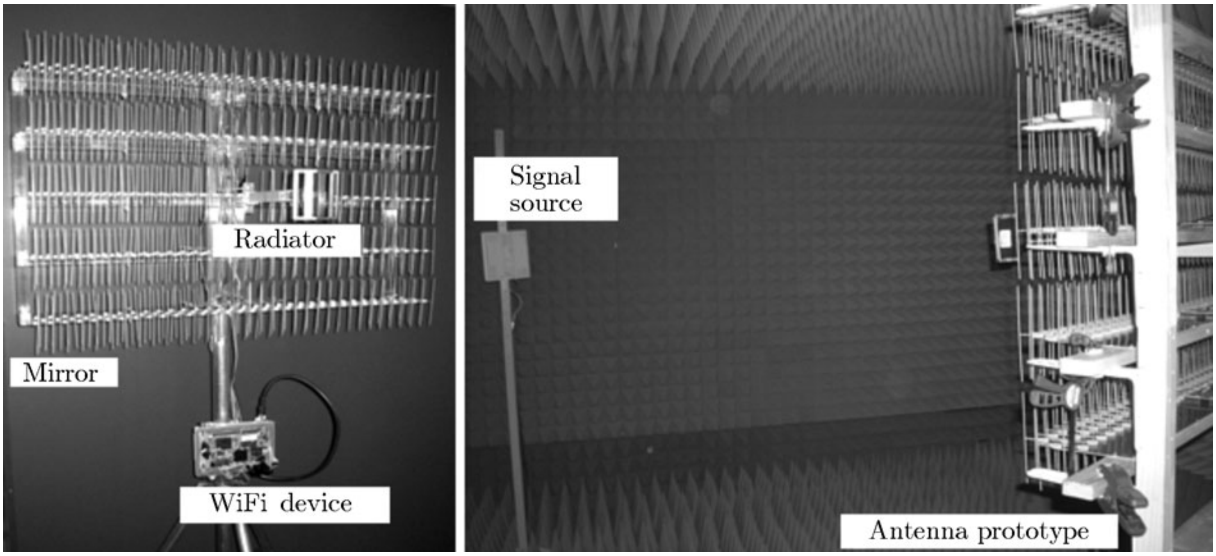

Fig. 5. Antenna prototype in an anechoic chamber.

the evolutionary genetic algorithm [20]. The results of the numerical experiments demonstrated that the optimal distance between the scatterers (from the viewpoint of the maximum tuning range of the phase of the reradiated field) ranges from 3 to 6 cm. In this case, the phase tuning range can reach 100° (at $d = 3$ cm). The optimal distance between the layers of the scatterers in one row varies from 20 to 30 mm, and the phase tuning range for the three-layer structure can reach 270°–360°.

3. LABORATORY MODEL OF THE ANTENNA

The laboratory model of the antenna was created using the above-described scatterers and basing on the multi-layer scheme shown in Fig. 1 (antennas with different numbers of layers and scatterers, and different spatial locations of the layers and scatterers were studied in different experiments). To fix the scatterers at certain spatial points, special holders were used, which served simultaneously to feed the bias voltages to vc diodes of the scatterers. The scatterers are fixed on the holders by means of end connectors (element *5* in Fig. 2). The design of the holders allow one to vary the distance between the scatterers in a layer, as well as the distance between the layers.

The bias voltages are formed at the scatterers by means of control boards connected to a PC. The boards receiving command codes allow one to specify the distribution of the bias voltages at the vc diodes of all scatterers (up to 500) which form the mirror. The voltage can vary in the 0–15 V range with an increment of 0.5 V. Up to five 100-channel control boards can be used in the laboratory setup.

The radiator being a back-radiation antenna consists of an active dipole and a reflector and is located in front of the mirror formed by the set of scatterers. The width of the main lobe in the radiator's radiation pattern with regard to the -3 dB level is 60°. This radiator is similar to that used in the standard parabolic antenna having a gain of 24 dBi and a mirror having a size of 60×100 cm and a focal distance of 30 cm. The location of the radiator with regard to the first row of the scatterers can vary. Then, the focal distance of the parabolic reference antenna was chosen as the initial distance.

All the antenna experiments were performed in an anechoic antenna chamber.

The field that radiated the antenna was produced by means of a horn in the far zone of the antenna under consideration. The photos of the antenna in the chamber are shown in Fig. 5. The right-hand photo shows the signal source and the studied antenna prototype. Both devices were connected to the network analyzer that measured the transmission coefficient S_{12}. The measurement of this coefficient allowed one to monitor the level of the received signal and its dependence on the antenna rotation angle (the radiation

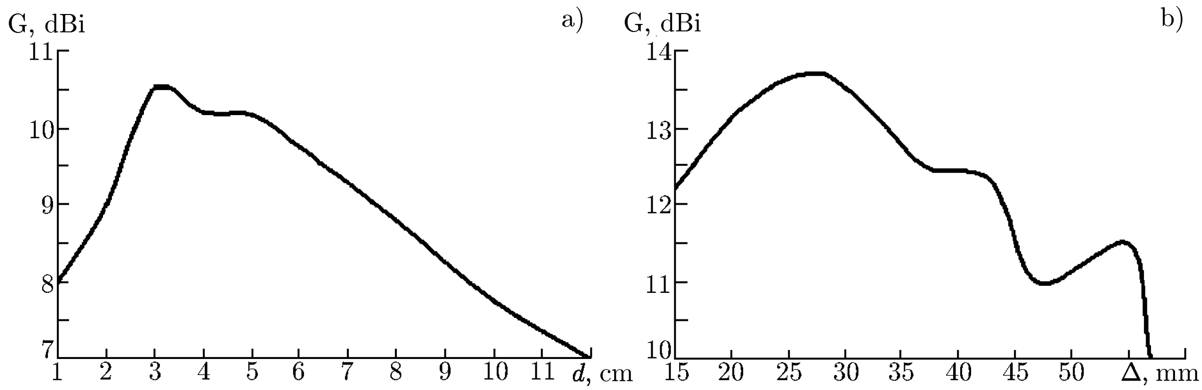

Fig. 6. Antenna gain (G) as a function of the distance d between the scatterers (*a)* and the distance Δ between the layers of the scatterers (*b*).

pattern). The antenna was turned in the azimuthal plane and tilted in the vertical plane by means of a turning table.

Electric tuning of the antenna, i.e., the choice of the bias voltage for all vc diodes of the scatterers forming the antenna mirror was performed in the automatic regime with respect to the criteria of the maximum power of the signal received by the antenna. For tuning, we used the algorithms of multi-dimensional optimization [20–22] (the Nelder–Mead method and different variants of genetic algorithms), which were realized in software installed on the computer which controlled the antenna.

The values of the bias voltage, which were found by using the optimization algorithms, were stored and could be used directly as required to perform the electric tuning of the antenna again.

4. ANTENNA TUNING

To achieve the required parameters of the antenna, its structure was refined manually by choosing the distances between its elements. The distances between the scatterers, which were found in the numerical experiments, were considered reference ones.

The structure of the antenna was refined in several stages. At the first stage, the optimal (in terms of the maximum antenna gain) distance between the scatterers in the layer was found. To determine this parameter, an antenna consisting of one row and one layer of the scatterers was assembled. Basing on the found optimal distributions of the bias voltage at distances between all array elements ranging from 1 to 12 cm (the wavelength λ is 12.5 cm) and a fixed row length of 100 cm, the dependence of the antenna gain in the central direction on the distance d between the scatterers was plotted (see Fig. 6a). It is seen that at $d = 3$ cm, the maximum of the gain of a single-row single-layer antenna is achieved, which is equal to 10.5 dBi. An increase in the density of the array elements leads to an increase in the mutual influence and deterioration of the reflection parameters (narrowing of the phase tuning range), whereas an increase in the distance between the scatterers leads to a decrease in the efficiency of interception of the radiation from the active element of the system.

At the second stage, we determined the optimal (from the viewpoint of the maximum antenna gain) distance between the layers of the scatterers in one row. To determine this parameter, we assembled an antenna, which consisted of one row and two layers of the scatterers, with the fixed distance $d = 3$ cm between the scatterers in each layer. Figure 6b shows a plot of the antenna gain in the central direction as a function of the distance Δ between the layers, which changes from 15 mm to 58 mm with an increment of 2 mm. The maximum gain 13.5 dBi was reached at $\Delta = 27$ mm. For a three-layer antenna, at the same distance between the layers $\Delta = 27$ mm, the maximum of the antenna gain amounts to 14 dBi. At such distances, smooth tuning of the phase along the mirror is ensured for the wave reflected by the mirror, and the negative effect of the mutual influence of the antenna array elements is minimized.

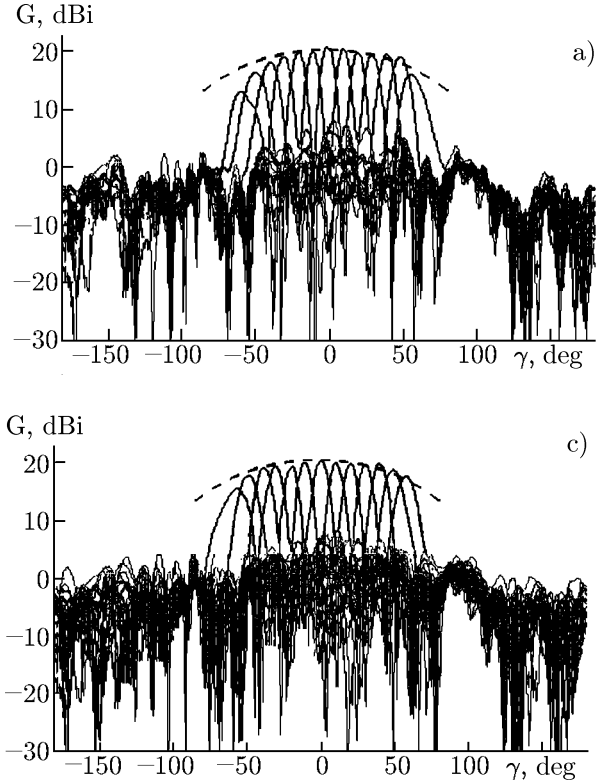

Fig. 7. Antenna radiation patterns in the case of electric tuning in different azimuthal directions: elevation angle -10° (*a*), elevation angle 0° (*b*), and elevation angle 10° (*c*). Solid lines correspond to the experimental data, and the dashed line is the cosinusoid decline of the antenna gain.

In the process of manual antenna tuning, we noted good agreement of the experimentally obtained values of d and Δ and the recommended values found by numerical simulation (see Sec. 1).

At the next stage of the experiments, the maximum value of the antenna gain was found as a function of the number of rows used in the antenna. The maximum gain was equal to 14 dBi for one row, 16.2 dBi for two rows, 18.9 dBi for three rows, 20 dBi for four rows, and 21.2 dBi for five rows. The distance between the centers of the rows was 15 cm, which exceeds the wavelengths by 20%. A fivefold increase in the number of rows and, consequently, in the antenna area corresponds to a fivefold increase in the antenna gain (to 7 dB approximately), i.e., addition of new rows does not lead to a significant additional mutual influence of the elements, which distorts their phase-amplitude characteristics.

5. FORMATION AND OPTIMIZATION OF THE RADIATION PATTERN

The series of experiments, the results of which are presented in this section, was aimed at compiling the table of distributions of the control voltages corresponding to different directions of the main lobe of the radiation pattern and ensuring the maximum antenna gain in specified directions.

The experiment was performed in an anechoic antenna chamber. In its process, the antenna was turned to a specific angle in the horizontal and vertical planes, after which the optimization software was launched to find the bias voltages at the scatterer diodes, at which the maximum gain for this direction was achieved. This procedure was repeated until the table was completed, which allowed tuning the antenna in the angle interval $\pm 60^\circ$ in the horizontal plane and $\pm 15^\circ$ in the vertical plane with a step of 5°.

Figure 7 shows the antenna radiation patterns during its electric tuning in different azimuthal radiations γ at elevation angles -10°, 0°, and 10°. The experimental values are shown by solid lines, and the theoretical cosinusoid decline of the antenna gain (due to a decrease in the effective antenna area), by a dashed line. It is seen from the plots in Fig. 7 that the experimental data agree sufficiently well with the theoretical ones in all directions excluding $\gamma = -60^\circ$ for all elevation angles. This is explained by the experimental conditions of the antenna optimization. The low level of the gain in this direction was caused by the doorway in the antenna chamber, which was the reason for parasitic reflections in this situation.

The experiments on simultaneous formation of maxima and minima of the radiation pattern in specified directions were also performed in an antenna chamber. Two radio signal sources were used, which simulated a useful signal and interference. In this case, the optimization problem was aimed at achieving the maximum difference in the power of two signals (signal-to-noise ratio). It was found as a result of the experiments that for a satisfactory antenna gain (about 18–21 dBi) in the direction of the useful-signal source, the interference signal level was about -20 dBi.

The obtained results are tentative and demonstrate only the wider possibilities of controlling the radiation pattern of the antenna under consideration. Additional studies are required to find out whether it is possible to suppress interference depending on how close to each other the sources of the useful signal and the interference are.

6. STUDY OF THE BROADBANDEDNESS OF THE ANTENNA SYSTEM

When an antenna is used in wide-band wireless-communication systems, e.g., IEEE 802.11 Wi-Fi or IEEE 802.16 WiMAX, the width of the frequency band is one of its most important parameters.

For example, 11 frequency channels are assigned for communications using the IEEE 802.11 WiFi protocol at a frequency of 2.4 GHz. Each channel has a width of 22 MHz, and the total frequency band is equal to 72 MHz. Thus, it is necessary to ensure that at least within one frequency channel the gain of the tuned antenna does not change, or its variations do not exceed 1 dB. The best result will be a change the gain of the tuned antenna by less than 1 dB in the chosen direction in all frequency channels simultaneously, or the possibility to tune the antenna electrically in each frequency channel (with the gain changing by less than 1 dB within each channel).

The plots which characterize the broadbandedness of the antenna are shown in Fig. 8. The solid line shows the frequency dependence of the gain of the antenna tuned to a frequency of 2.415 GHz in the direction of an elevation angle of 0° and an azimuth of 0°. The bandwidth (with respect to the -1-dB level) is equal to 47 MHz, which corresponds to four frequency channels (from the first to the fourth). The dashed line corresponds to a similar dependence for the antenna tuned to a frequency of 2.415 GHz in the direction of an elevation angle of 0° and an azimuth of 30°. In this case, the bandwidth amounts to 45 MHz, which also corresponds to four frequency channels. Significant aberrations of the antenna gain are observed at frequencies going beyond the 50 MHz band. The maximum deviations are 3.5 dB for an elevation angle of 0° and an azimuth of 0°, and 4.5 dB for an elevation angle of 0° and an azimuth of 30°.

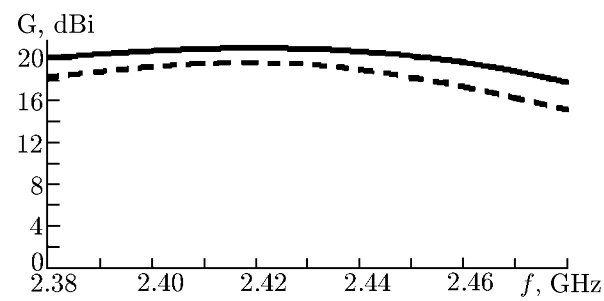

Fig. 8. Frequency dependences of the antenna gain for the antenna tuned to a frequency of 2.415 GHz, an elevation angle of 0°, and an azimuth of 0° (solid line), and for the tuned to a frequency of 2.415 GHz, an elevation angle of 0°, and an azimuth of 30° (dashed line).

Analysis shows that in the case of electric tuning of the antenna system having this design, one can achieve a uniform frequency characteristic over the entire range (with variations in the antenna gain being less than 1 dB), but with a level of the antenna gain being less by 1.5–2 dB than the maximum value for this direction.

7. FIELD TESTS OF THE ANTENNA SYSTEM

The field tests of the antenna system were performed by establishing a wireless communication channel between points located within the municipality of Nizhny Novgorod (in 2008) on the roofs and top floors of residential and office buildings, as well as between several points spaced up to 7 km in two municipalities separated by a river (in 2009).

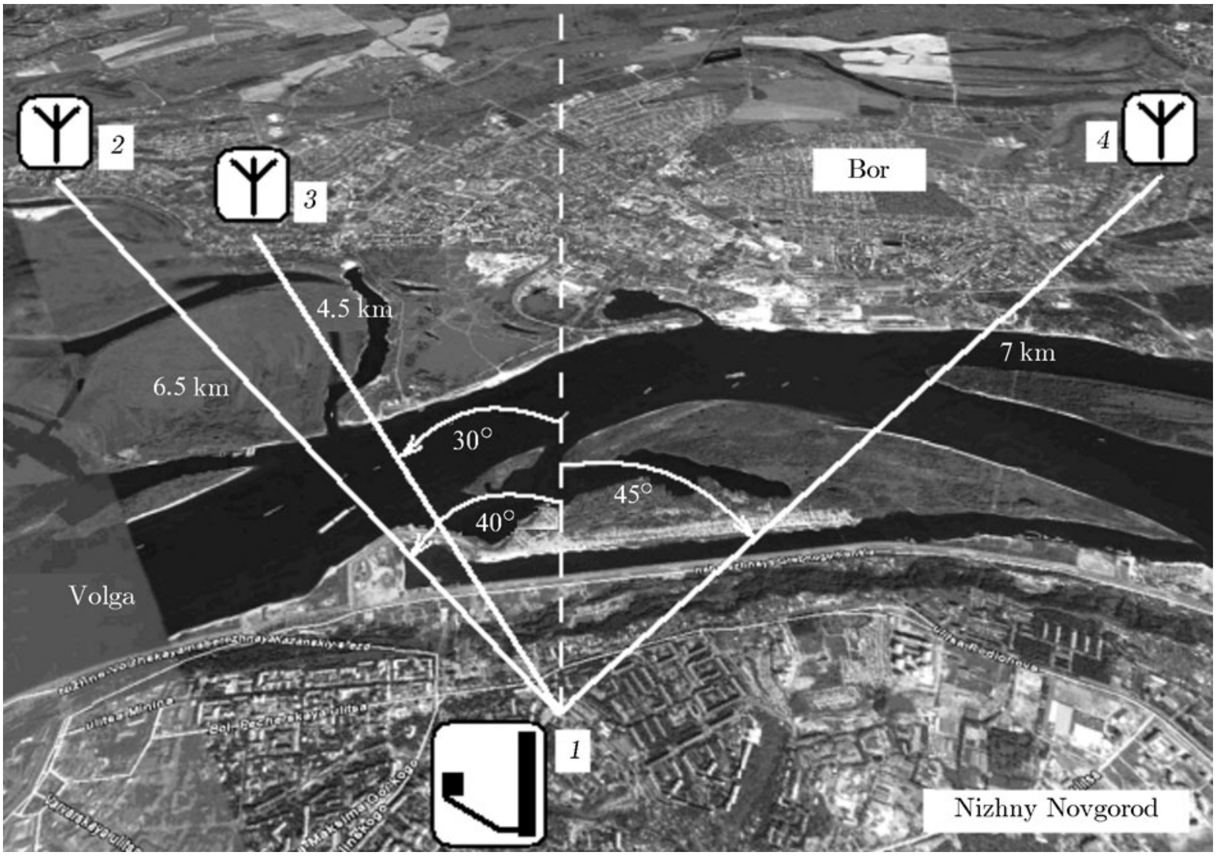

Fig. 9. Scheme of communication channels across the river: the antenna under consideration (*1*) and the communication points (*2–4*).

The scheme of communication channels across the river is shown in Fig. 9. The antenna under consideration, which was connected to a spectrum analyzer or a special Wi-Fi receiver, was located in the city of Nizhny Novgorod on the ninth floor of an office building (*1*), at a height of 30 m above the ground. Standard parabolic antennas with a directive gain of 24 dBi, which were connected to the generators producing microwave signals at a frequency of 2.4 GHz or to special Wi-Fi transmitters, were installed at three points in the town of Bor, namely, at a maternity hospital (*2*), a school (*3*), and a private house (*4*).

The essence of the experiment was electrical tuning of the antenna radiation pattern to the signal source, further establishing of the communication over the IEEE 802.11b (Wi-Fi) protocol and measuring of the maximum rate of the data transfer rate over this protocol using the antenna under consideration. To measure the carrying capacity of the communication channel, the antennas were connected to wireless-communication devices based on the Gateworks Avila Network Processing System (GW2348-4) with network adapters Ubiquity XR2. The duration of copying of a 10-megabyte file over the FTP protocol from point *1* to points *2*, *3*, and *4* was measured. The measurements were performed for different frequency channels. The transfer rate amounted from 6.17 megabit/s to 7.53 megabit/s (6.85 megabit/s ±10%), which is comparable with the results of the measurements [23] performed in similar conditions with the use of standard wireless communication means without the possibility of controlling the main lobe of the radiation pattern.

An additional series of field tests of the antenna was performed in the municipality of Berkley (USA) in the period from 2007 to 2009. The experiments were aimed at finding whether it was possible to locate Wi-Fi access points within the territory of the city using the antenna installed at a height of 40 m and scanning several building blocks (10 km^2) in the interval of azimuthal angles ±60°. In the process of the experiments, we managed to detect signals from more than fifty access points and determine their azimuths. The knowledge of angular coordinates of the access points allowed us to establish connection consecutively with any of them. Similar results were obtained by rotating mechanically a reference parabolic antenna with a gain of 24 dBi.

8. CONCLUSIONS

This paper demonstrates the possibility of creating an antenna with a high antenna gain and a steerable radiation pattern, which is based on a mirror formed by dipole scatterers with variable parameters. The simplicity of the antenna structure and minimization of the number of microwave elements allow one to expect that such antennas can be used effectively in mass-production up-to-date wireless communication systems. Moreover, the proposed design of the antenna can be used in both more and less high-frequency ranges.

The proposed and studied antenna design solves the task of achieving the required parameters. However, it also allows further optimization (choice of scatterer types, refinement of their geometric dimensions and mutual spatial positioning with detailed account for the influence of different values of the load impedances on the neighboring scatterers, etc.). The possibility of the antenna's forming maxima and zeros of the radiation pattern simultaneously, as well as minimizing the level of side lobes requires a separate study.

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