Session 1A6a
Steerable Reflect-array Antennas

Principles of Synthesis of Steerable Reflect-array Antennas

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Abstract—The synthesis of steerable reflect-array antennas, as a synthesis of any antenna, should be realized by solving two problems: external antenna problem and internal antenna problem. The first one includes the investigation of the antenna radiation pattern. The second one concerns a design of the structure providing the required amplitude-and-phase distribution along the antenna aperture. The first problem is based on the theory of antenna array. The distance between the small reflectors and geometry of the array are responsible for the directivity, the beam width, and the side lobe level of the antenna. The internal antenna problem consists of finding the phase shift required for each small reflector. Remarkable feature of a tunable reflector made as a microstrip vibrator or a patch in combination with a tunable device is that the almost 360° phase shift of the reflected wave can be provided with only one tunable device (varactor diodes or ferroelectric tunable capacitors). Commonly, simulation of the phase shift required is performed with a numerical technique. The design of reflect-array antenna can be sufficiently simplified, if the simulation by the numerical technique is amplified with a correct analytical model.

1. Introduction

Reflect-array antennas are being developed during many years [1–3]. Recently the reflect-arrays were suggested as structures with an electronically steered radiation pattern [4, 5]. Such a steerable reflect-array antenna can be used as a low cost version of a phased array antenna for a wide commercial application. The synthesis of steerable reflect-array antennas, as a synthesis of any antenna, should be realized by solving two problems: external antenna problem and internal antenna problem. The first one includes the investigation of the antenna radiation pattern. The second one concerns a design of the structure providing the required amplitude-and-phase distribution along the antenna aperture. Any reflect-array antenna consists of primary radiator or illuminator and a reflecting surface. The primary radiator provides the reasonable amplitude distribution with a minimum spillover loss. Sophisticated design technique was used to diminish a constructive space occupied by the radiator [6]. The reflected surface is covered by a large number of small reflectors in form of microstrip vibrators or patches. The vibrators or patches are connected with tunable devices (varactor diodes [4] or ferroelectric tunable capacitors [7]), which serve for controlling the phase of the wave reflected by each small reflector. The goal of this paper is to characterize the main stages of synthesis and design of a steerable reflect-array antenna.

2. External Antenna Problem

The scheme of a typical reflect-array antenna is shown in Fig. 1. The system of patches provides transformation of a spherical wave front of the primary radiator into the plane wave front in the antenna aperture. The main beam width of an antenna and the directivity of the antenna, which are required, determine the size of the antenna aperture.

In Fig. 2, radiation pattern of a circle aperture with homogeneous field distribution as a function of a generalized angle function u is shown [8]. ($u = kR\sin\theta, k = 2\pi/\lambda$, where λ is wavelength is free space, R is radius of the circle aperture). In the case of homogeneous field distribution the main beam width in degree is $\Delta\theta^\circ = 59 \cdot \lambda/2R$ and level of the first side lobe is -18 dB. The antenna directivity is $D = 4\pi(\pi R/\lambda)^2$. If the field distribution decays to edge of the aperture, the main beam width is higher and the directivity is lower, the first side lobe level being decreased.

Forming the radiation pattern is drastically influenced by the inhomogeneity of the phase distribution over the radiating aperture. In the case of the reflect-array antennas the phase inhomogeneity can be provoked by inaccuracy of realization of size and position of the elementary radiators (patches). Let us consider a statistical estimation of the inaccuracy of the array realization [9].

The array radiation pattern can be presented as follows:

$$\Phi(\theta,\varphi) = \sum_{i=1}^{m} A_{0,i}\varphi_i(\theta,\varphi) \tag{1}$$

where $A_{0,i}$ is the optimized current amplitude of i-th elementary radiator, $\varphi_i(\theta,\varphi)$ is the pattern of i-th elementary radiator taking into account the position of the center of its phase pattern, m is the number of elementary

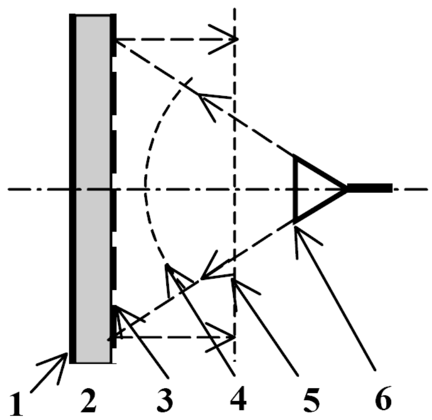


Figure 1: Scheme of a typical reflect-array antenna with: ground plane (1), dielectric layer (2), patch reflector (3), spherical wave front (4), plane wave front (5), primary radiator (6).

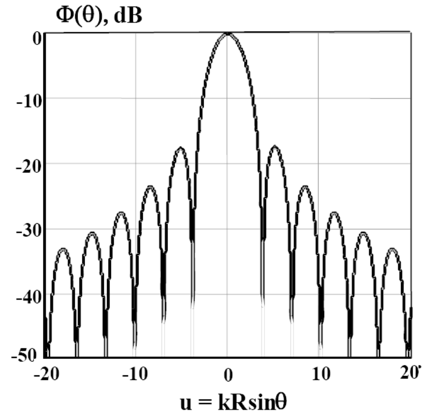


Figure 2: Radiation pattern of a circle aperture as a function of generalized angle function u.

radiator in the array.

If the real current distribution differs from the optimized one

$$A_i = A_{i,0} + \Delta_i, \tag{2}$$

the mean-square-error of amplitude/phase distribution over whole array is

$$\beta_{mse} = \left| \sum_{i=1}^{m} |\Delta_i|^2 / \sum_{i=1}^{m} |A_i, 0|^2 \right|^{1/2}. \tag{3}$$

The value β_{mse} can be used to find decay of the array directivity (by the factor g) and increase of the side lobe by ξ_{sl}:

$$g = \frac{1}{1+\beta_{mse}^2}, \qquad \xi_{sl} = \frac{3\beta_{mse}}{\sqrt{m}} \tag{4}$$

Let us suppose for example: $m = 2000$, $|A_{i,0}| = 1$ for all i, 30% of radiators are characterized by the phase error of 90°. Simple calculation gives $\beta_{mse} \cong 0.85$, $g = 0.58$ (decrease of the directivity in 2.4 dB), $\xi_{sl} = 0.06$ (increase of the side lobe level up to -12 dB).

3. Internal Antenna Problem

The primary radiator provides feed of the reflector patches with amplitude distribution required and with a minimum spillover loss. The efficiency of the primary radiator can be determined by the following equation:

$$\eta(\rho,\gamma) = \int_0^{\alpha(\rho)} [F(\theta,\gamma)]^2 \sin\theta d\theta / \int_0^{\pi/2} [F(\theta,\gamma)]^2 \sin\theta d\theta, \qquad F(\theta,\gamma) = [\cos(\theta)]^y \tag{5}$$

where $\rho = F/2R$, $\cot(\alpha(\rho)) = 2\rho$, F is the focal distance of the patch reflector, R is the radius of the aperture. $F(\theta,\gamma)$ is the radiation pattern of the primary radiator, the exponent γ determines the directivity of the primary radiator.

Table 1 illustrates the efficiency of the primary radiator expressed in dB. The data presented are followed by the conclusion that the preferable values of F and γ are $F \leq R, \gamma \geq 1$.

The phase of waves reflected by the patch mirror has to meet two principal demands: 1) Transformation of spherical wave front given by the primary radiator into a plane wave, 2) Providing phase gradient along the array, which corresponds to the beam deflection required.

The distance between the patches lies in the range $s = (0.63 - 0.67)\lambda$. If the linear size of the array (2R) is much higher than the wavelength (λ), the total phase shift change along the array can be much higher than 360°. In this case the phase distribution is corrected by reset of the phase by n times 360° where $n = 1, 2, 3, \ldots$ Such a phase correction is well known as a characteristic feature of Fresnel mirror.

The phase shift of the wave reflected by a patch depends on the patch size [1–4, 6]. That is illustrated by Fig. 3 for the operational frequency $f = 10\,\mathrm{GHz}$. Commonly, simulations of the required phase shift are performed with a numerical technique. The result of the phase shift simulation is used for designing the reflect-array antenna. In photo (Fig. 4) one can see the distribution of the patch sizes over the mirror array.

Table 1: The primary radiator efficiency $\eta(\rho, \gamma)$ in m dB.

ρ	0.2	0.3	0.4	0.5	0.6
$\gamma = 1.5$	-0.09	-0.31	-0.71	-1.21	-1.82
$\gamma = 1.0$	-0.23	-0.63	-1.20	-1.90	-2.60
$\gamma = 0.5$	-0.65	-1.22	-2.18	-3.00	-3.92

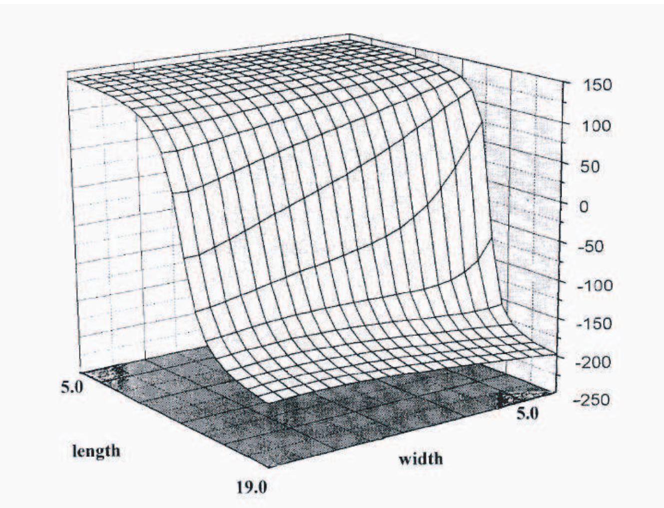


Figure 3: The phase shift of wave reflected by a patch as a function of length and width of the patch in mm for dielectric constant of the substrate $\varepsilon_{\mathrm{sab}} = 1.06$.

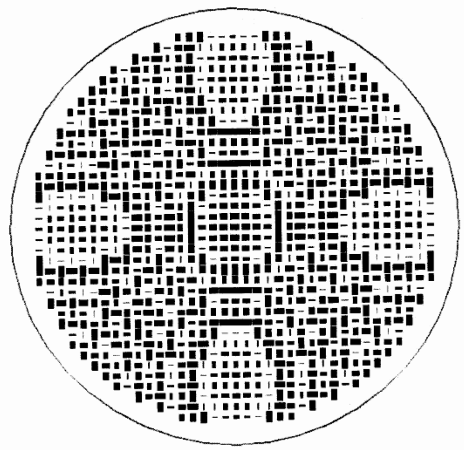
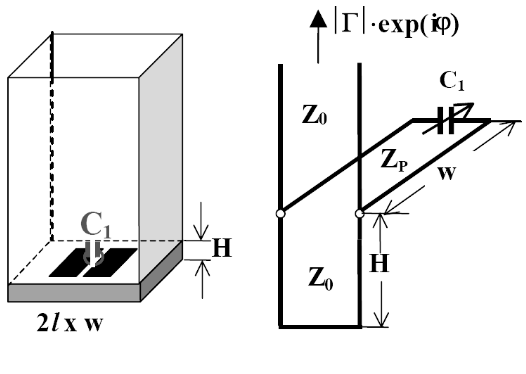


Figure 4: Distribution of the patch positions and sizes in array.

4. The Steerable Patch Array

The phase shifts (Fig. 3) and an appropriate design of the patch array (Fig. 4) correspond to a nonsteerable array with a fixed position of the main beam. In order to control the position of the main beam of radiation pattern, the phase shift of the reflection coefficient of each patch in the array should be controlled. The tunable devices (semiconductor varactor diodes [4] or ferroelectric tunable capacitors [7]) should be included in each patch. The state of the tunable device serves for controlling the phase of wave reflected by each small reflector. Remarkable feature of a tunable reflector made as a microstrip vibrator or a patch in combination with a tunable device is that the 360° phase shift of the reflected wave can be provided with only one tunable device. It should be reminded that for a realization of a digital transmission-type phase shifter one needs at least 8 tunable devices [9]. The optimum phase shift of each tunable reflector must be found as a result of a correct simulation and can be realized by applying to each tunable device the appropriate biasing voltage. The result of the phase shift simulation must be included in the driving program of the biasing voltage controller.

The problems mentioned above can be sufficiently simplified, if the simulation by the numerical technique is replaced by using a correct analytical model. The problems mentioned above can be sufficiently simplified, if the simulation by the numerical technique is replaced by using a correct analytical model. A scheme of a tunable patch is shown in Fig. 5. The sketch drawn in Fig. 5(a) presents a single patch located in a virtual waveguide confined by electrical and magnetic walls. The patch is considered as a microstrip vibrator loaded by the tunable capacitor. The microstrip is formed on a dielectric substrate with a conductive ground plane. Microwave current in the vibrator is induced by the incident wave. The current distribution along the vibrator is found on a basis of solution to telegraph equations using the method of induced electromotive forth [8]. The equivalent schematic is shown in Fig. 5(b). Two type of resonances can be observed in the circuit shown in Fig. 5(b). Firstly, fundamental resonance, which corresponds to infinite impedance of the vibrator, which



a) *b*)

Figure 5: Scheme of a tunable patch. a) Single patch located in a virtual waveguide confined by electrical an magnetic walls and b) Equivalent schematic.

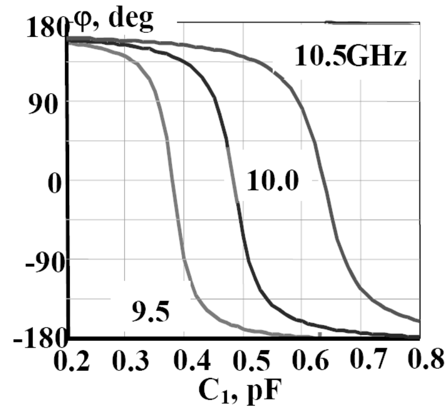


Figure 6: Simulations of the reflected wave phase shift in the framework of the schematic analytical model.

is presented by the transmission line stub with length l and characteristic impedance Z_P. In this case one observe the reflection from the ground plane through the substrate; the phase of the reflection coefficient is near to $\pm 180^\circ$. Secondly, anti-resonance, which corresponds to parallel resonance of two transmission line stubs (l, Z_P) and (H, Z_0). In this case one observes the zero phase of reflection from the plane, in which the vibrator is located. Thus, change of capacitance of the tunable capacitor makes possible to obtain the change of the reflection phase approximately in the range $+180 \ldots 0 \ldots -180$ degrees. In Fig. 6, results of simulations in the framework of schematic analytical model are presented [10]. The following data were taken: the square virtual waveguide $20 \times 20\,\mathrm{mm}^2$ with the substrate $H = 1\,\mathrm{mm}$, $\varepsilon_S = 3.0$; dimensions of the patch: $w = 2\,\mathrm{mm}$, $2l = 9\,\mathrm{mm}$, operational frequency $f = 9.5 - 10.5\,\mathrm{GHz}$.

Simulations based on the schematic analytical model can be used for the design of a steerable reflect-array and for developing a driving program of the biasing voltage controller. Some fitting parameters of the analytical model can be found using a comparison with the full-wave analysis simulation.

5. Conclusion

The development of simple and correct theoretical models and elaboration of material components of reflect-array antenna is an urgent problem, which solution is important for realization of a cheap steerable antenna for mass production.

REFERENCES

1. Huang, J., *IEEE Antennas Propagation Soc. Int. Symp. Dig.*, 612–615, 1991.
2. Javor, R. D., X.-D. Wu, and K. Chang, *IEEE Trans. Antennas Prop.*, Vol. 43, 932–939, Nov. 1995.
3. Chang, D. C. and M. C. Huang, *IEEE Trans. Antennas Propagat.*, Vol. 43, 829–834, Aug. 1995.
4. Boccia, L., F. Venneri, et al., *Proc. IEEE Antennas Propagat. Soc. Int. Symp.*, Vol. 3, 132–135, 2002.
5. Sievenpiper, D. F., J. H. Schaffner, et al., *IEEE Trans. Antennas Prop.*, Vol. 51, 2713–2723, Oct. 2003.
6. Pilz, D. and W. Menzel, *Electron. Lett.*, Vol. 34, No. 9, 832–833, April. 1998.
7. Romanofsky, R., J. Bernhard, et al., *IEEE MTT-S*, 00TH 1D$_1$, 2000.
8. Balanis, C. A., *Advanced Engineering Electromagnetics*, John Wiley, 1989.
9. Vendik, O. G. and M. D. Parnes, *Antennas with Electrical Scanning (Introduction to Theory)*, Ed. by L. D. Bakhrakh (in Russian), Science-Press, Moscow, 2002.
10. Vendik, O. G. and M. S. Gashinova, *St. Petersburg Electrotechnical University*, to be published.

Tunable Impedance Surfaces for Low-cost Beam Steering and Conformal Antennas

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HRL Laboratories has developed a tunable impedance surface that enables a low-cost electronic beam steering. It consists of a lattice of resonant elements that are interconnected with their neighbors by varactor diodes. The surface provides a frequency dependent phase shift for reflected waves, which depends on the resonance frequency of the resonators. By varying the reverse bias voltage on the varactors, we can locally tune the resonance frequency, creating a programmable phase shifting surface. This is used to steer or focus microwave energy.

The tunable surface can also be used with a conformal feed. In this implementation, we excite surface currents with a small antenna located directly adjacent to the surface. We tune the surface to create a periodic grating structure, and the surface waves leak off at an angle determined by the period of the surface impedance. This implementation eliminates the protruding structure required by a space feed, resulting in a complete antenna structure that is electrically thin. We have shown that this concept can also be used on curved surfaces, so the antenna can be conformal to arbitrary shapes.

Unique Issues and Features of a Scanning Reflectarray Antenna Based on Ferroelectric Thin Film Phase Shifters

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Evolving high data rate communications systems demand greater attention to subtle aspects of information theory and electromagnetic engineering. As the ratio of signaling bandwidth to carrier frequency decreases, less familiar phenomenon influence system performance. Some interesting effects are expected to appear if the trend toward wide-band scanning phased array antennas and bandwidth-efficient, high-speed modulators continues. Indeed there is a growing demand for efficient, low-cost phased array antennas. The reflectarray is an alternative to directly-radiating phased array antennas and promises higher efficiency at reduced cost. The ferroelectric reflectarray involves phase shifters based on coupled microstrip patterned on $Ba_xSr_{1-x}TiO_3$ films, that are laser ablated onto $LaAlO_3$ substrates. These devices outperform their semiconductor counterparts from X-through and K-band frequencies. There are special issues associated with the implementation of a scanning reflectarray antenna, especially one realized with thin film ferroelectric phase shifters. This paper will discuss these issues which include modulo 2π effects and phase shifter transient effects on bit error rate, scattering from the ground plane, relevance of phase shifter loss and presentation of a novel hybrid ferroelectric/semiconductor phase shifter, and the effect of mild radiation exposure on phase shifter performance.

Design of a Steerable Reflect-array Antenna with Semiconductor Tunable Varactor Diodes

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Abstract—The demonstrator of a steerable reflect-array antenna was designed as a system of dipoles loaded by varactor diodes. The microwave response of a dipole loaded by varactor has been simulated in closed form based on equivalent circuit approach. The circuit analytical model has been verified by the full-wave analysis. Change of the varactor capacitance in the range of 0.3–1.3 pF was provided by biasing voltage 0–20 V. The array consists of 20 dipoles structured as two parallel lines. The operational frequency is 11 GHz, the length of the dipole is 9.2 mm, spacing between dipoles is 18 mm. The double-side metallized PTFE with $\varepsilon = 2.8$ and thickness of 1 mm was used as a substrate. Dipole structure was manufactured by a photolithographic process and formed with surface mounted varactor diodes. The radiation pattern of the array is characterized by the width of the main beam $\approx 8^\circ$, the side lobe level -$(12 \div 20)$ dB, the steering range $\pm 15^\circ$. Control voltage was set manually with variable resistors separately for each varactor. The fine alignment of the control voltage for each varactors turned out to be very important. Inherited data are used for correction of the operational principle of a varactor steerable antenna controller.

1. Introduction

The reflect-array antennas are being studied many years and some theoretical and experimental results have been obtained [1–3]. The possibility to obtain an electron steering of the radiation pattern of such an antenna is currently under investigation [4, 5]. The goal of this paper is to discussed an experimental realization of a steerable reflect-array antenna demonstrator designed as a system of dipoles loaded by varactor diodes. The GaAs varactors MA46H070 produced by MACOM Inc. were tried. Control voltage applied to each varactor in the array should provide the phase shift along the dipole structures, which is necessary for transformation of a spherical phase front of a prime radiator into the plane phase front with the required declination. One should simulate the distribution of the phase shift of reflection coefficient of each dipole and find the dependence of the phase shift on the biasing voltage applied to the varactor loading the dipole. A set of error sources complicates solution to the problem. The following errors should be taken into account. 1) Incorrectness in simulation of the phase shift of reflection coefficient as a function of the dipole sizes and value of the varactor capacitance. 2) Dispersion of the dependence of the varactor capacitance on the applied biasing voltage. 3) Fabrication errors in dimensions of the design components.

In order to overcome difficulties mentioned above, the experience of designing and examination of different version of steerable reflect-array antenna should be accumulated and used for developing the design procedure.

2. Design of the Demonstrator

In Fig. 1 structure of the array under investigation is shown. The array consists of 20 dipoles structured as two parallel lines. Spacing between lines and spacing between dipoles in the lines are 18 mm. Each dipole is loaded by a tunable varactor. The dipole length is 9.2 mm, The doubleside metallized PTFE with $\varepsilon = 2.8$ and thickness of 1 mm was used as a substrate. A *dc-rf* filter and *dc*-biasing strip lines are arranged for each dipole. Fig. 2 shows the equivalent diagram of the bench, which was used for formation of the control voltages. The control voltages were set manually with variable resistors separately for each varactor. The fine alignment of the varactor tuning was find to be very important. That can be explained by a sharp dependence of the phase of the dipole reflection coefficient on value the biasing voltage. Fig. 3 illustrates this dependence, which was simulated for the dipole considered. Fig. 4 shows the derivative of the reflection coefficient phase with respect to the biasing voltage as a function of the biasing voltage. One can see that at the section of curve (Fig. 3) with the highest slope, the setting the phase shift with accuracy $\pm 45^\circ$ requires the accuracy of setting the biasing voltage better than ±0.25 V. That relates not only to the accuracy of the biasing voltage control, but to repitability of the varactor Volt-Farad characteristics.

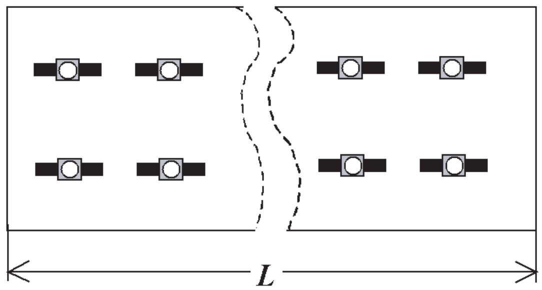

Figure 1: Structure of the array under investigation. The array consists of 20 dipoles structured as two parallel lines. The total length $L = 200$ mm.

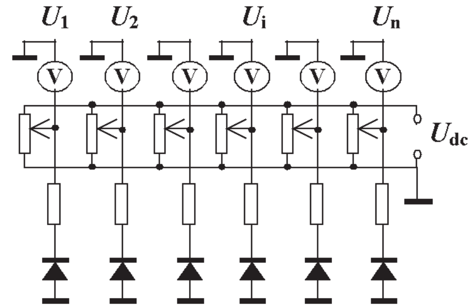

Figure 2: Scheme for formation of the control voltages. The voltages applied to varactors are separately registered.

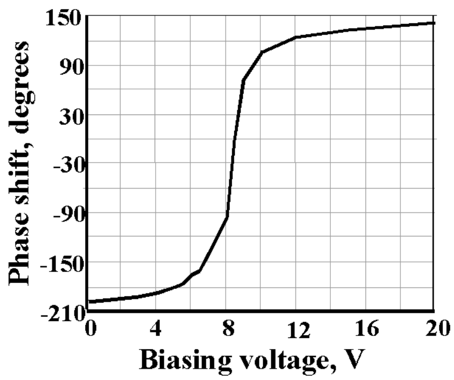

Figure 3: Dependence of the phase of the dipole reflection coefficient on biasing voltage.

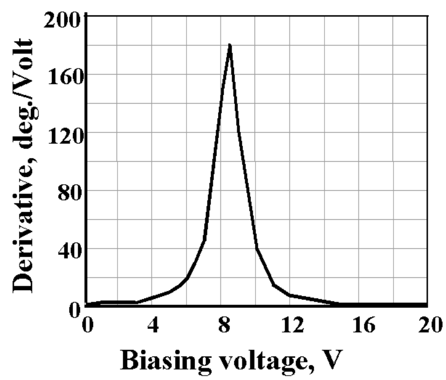

Figure 4: Derivative of the reflection coefficient phase with respect to the biasing voltage.

3. Some Experimental Results

The prime radiator was placed in a distance from the surface of the array of 100 mm. The open end of a rectangular waveguide with the transverse section $23 \times 5\,\mathrm{mm}^2$ was used as the prime radiator.

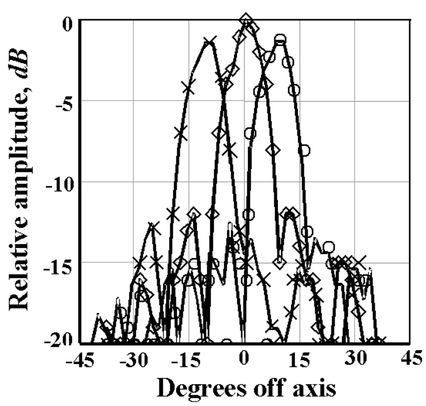

Figure 5: Radiation pattern of the steerable reflect-array (three positions).

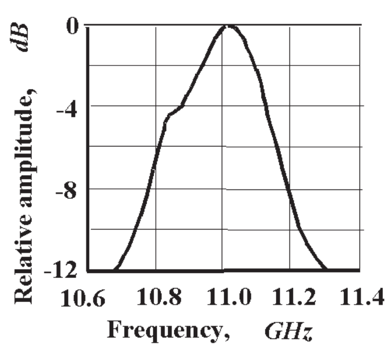

Figure 6: Amplitude-frequency response of the steerable reflect-array.

The required phase distribution along the array was simulated in closed form based on equivalent circuit approach. The adequacy of the circuit analytical model has been verified by the full-wave analysis. The simulation of the required phase distribution was followed by the simulation of distribution of the biasing voltage applied to all varactors.

Performance of the radiation pattern of the array at the frequency 11 GHz is shown in Fig. 5. Three positions of the main beam are shown: $-10°$; $0°$; $+10°$. The amplitudes of the radiation pattern in all positions are practically the same. While measuring the radiation pattern, the biasing voltage distribution was slightly corrected to obtain maximum of the signal.

The gain of the antenna was not measured, because the efficiency of the prime radiator had not been optimized.

Fig. 6 gives the amplitude-frequency response of array in the center position of the main beam. The frequency band of the antenna is about 2% at the level of -3 dB. Such a narrow frequency band can be referred to a high quality factor of a resonant tank formed by the dipole and the tunable varactor. It may be assumed that the reactive parameters of the pair dipole-varactor can be optimized and the frequency band can be extended.

4. Conclusion

A steerable reflect-array antenna demonstrator was experimentally realized. The results of the demonstrator investigation are in agreement with the theory of the steerable antenna arrays [6, 7]. That gives possibility to make a confident conclusion that the steerable reflect-array antenna based on application of the tunable varactors can be designed and manufactured. Such an antenna can be offered as a cheap version of steerable antenna for a mass production.

REFERENCES

1. Chang, D. C. and M. C. Huang, *IEEE Trans. Antennas Propagat.*, Vol. 43, 829–834, Aug. 1995.
2. Pilz, D. and W. Menzel, *Electron. Lett.*, Vol. 34, No. 9, 832–833, April 1998.
3. Boccia, L., F. Venneri, et al., *Proc. IEEE Antennas Propagat. Soc. Int. Symp.*, Vol. 3, 132–135, 2002.
4. Romanofsky, R., J. Bernhard, et al., *IEEE MTT-S*, 00TH 1D_1, 2000.
5. Sievenpiper, D. F., J. H. Schaffner, et al., *IEEE Trans. Antennas Prop.*, Vol. 51, 2713–2723, Oct. 2003.
6. Balanis, C. A., "Advanced engineering electromagnetics," *John Wiley* & *Sons*, 1989.
7. Vendik, O. G. and M. D. Parnes, "Antennas with electrical scanning (introduction to theory)," Ed. by L. D. Bakhrakh (in Russian), Moscow, *Science-Press*, 2002.

Modeling of Low-profile Reflect Array Antenna

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Abstract—The theoretical analysis of a planar reflect-array antenna consisting of a rectangular microstrip patch radiators is presented. Such an antenna is to be designed to convert spherical wave radiated by feed horn antenna into plane wave by phasing of reflected wave due to adjusting the patch sizes and arranging them by principle of Fresnel mirror. The modelling of array antenna is based on the modelling of elementary equivalent waveguide cell consisting patch radiator at the interface between superstrate and grounded substrate layers. Spectral Domain Approach (SDA) of Method of Moments is used to analyse the characteristics of elementary waveguide cell. Theoretical and experimental results are compared.

1. Introduction

Microstrip reflectarray antenna exploits operational principles of traditional parabolic reflector antenna and microstrip patch phased array [1–4]. Such a combination allows eliminating two disadvantages of both standard antennas. For microstrip patch antennas there is a common difficulty to overcome a 30 dB gain limit of phased array because of lossy feeding network. The conventional high-gain antennas are the parabolic reflectors. Being the very efficient radiators they are bulky and massive. A flat microstrip refletarray is being developed as a compact high-gain antenna [2]. Low loss is conditioned by the sizes of most radiators are far from ones of resonant half-wavelength radiators (used usually with transmission lines for phase adjustment) as well as by absence of feeding networks. Feeding of the printed radiators is realized quasi-optically.

Such an antenna is to be designed to convert spherical wave radiated by feed horn antenna into plane wave by phasing of reflected wave due to adjusting the patch sizes and arranging them by principle of Fresnel mirror.

The basic configuration of antenna includes a feed horn antenna and a printed reflectarray. Rectangular patches arranged in a planar aperture based on metal backed substrate will reradiate illuminated energy into space. Each radiator's phase is adjusted to make total reradiated field cophasal and concentrated in a specific direction. The phasing method is to use a variable size patches to form the front of reflected wave.

A folded version of such reflectarrays has been proposed and realized [1]. The dual polarization properties of rectangular patch array enable to remove the feed element from the focal point in front of antenna and to place it at the backside of antenna with polarizing grid for reradiation.

In present paper we describe a code developed for the design of multilayer printed reflectarrays, which adjusts the sizes of patches to achieve a progressive phase for dual linear polarisation according to both twisting and focusing requirements.

2. Theory

To build a procedure for calculation of a reflection coefficient of linear polarized wave normally incident on a patch radiator we assume that no coupling between adjacent patches takes place. Such a situation is valid for normal incidence on infinite periodical array and gives us a reasonable approximation for most part of finite nonperiodic array if the distance between adjacent radiators is big enough.

Figure 1 illustrates the normal incidence of the plane wave on a single microstrip patch. With two pair of opposite perfect electric and magnetic walls corresponding to the case of non-interacting radiators we can consider the whole structure as a stack of elementary TEM (in z direction) waveguides. Due to partial filling of such a waveguide with dielectric substrate, air superstrate and current carrying layers an existence of pure TEM mode represents another assumption and such a waveguide could be called as a quasi-TEM waveguide. To calculate the reflection coefficient of fundamental TEM mode we use a standard Spectral Domain Approach of Method of Moment.

General relation between electric fields in the plane of microstrip radiator is

$$\tilde{\bar{\mathbf{E}}}_{tot} = \tilde{\bar{\bar{\mathbf{G}}}} \cdot \tilde{\bar{\mathbf{J}}} + \tilde{\bar{\mathbf{E}}}_{inc}(1 + \tilde{\bar{\bar{\mathbf{\Gamma}}}}) \tag{1}$$

where $\tilde{\bar{\mathbf{E}}}_{tot}$ is vector of total tangential E-field (superposition of scattered and incident field), $\tilde{\bar{\bar{\mathbf{G}}}}$ is Green's dyad for elementary waveguide, $\tilde{\bar{\mathbf{J}}}$ is surface current density excited in microstrip dipole, $\tilde{\bar{\mathbf{E}}}_{inc}$ is vector of incident

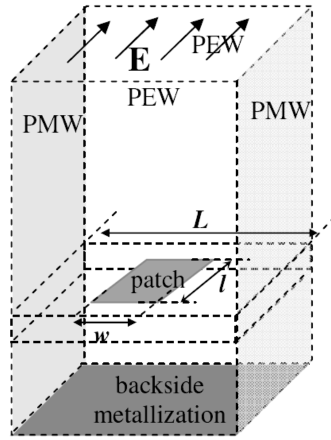


Figure 1: Equivalent elementary wavegude.

linear-polarized along the microstrip radiator side and $\tilde{\bar{\bar{\mathbf{\Gamma}}}}$ is a tensor of reflection coefficients. Tilde denotes operating in Fourier domain. $\tilde{\bar{\bar{\mathbf{\Gamma}}}}$ can be easily derived by analogy with determination of Green's dyad using standard immitance approach.

To approximate unknown current density one has to expand both components of $\tilde{\bar{\mathbf{J}}}$ by series of chosen basis functions.

Assuming that polarization of the incident wave is parallel to one side of rectangular patch, along y-direction in instance, we can suppose that y-polarized electric field excite only y-component of current density:

$$\tilde{\bar{\mathbf{J}}} = J_y \cdot \bar{\mathbf{e}}_y \tag{2}$$

Then equation (1) transforms to scalar one:

$$\tilde{E}_{y,tot} = \tilde{G}_{yy}\tilde{J}_y + \tilde{E}_{y,inc}(l + \tilde{G}_{yy}) \tag{3}$$

where

$$J_y = \sum_n A_n \varphi_n(x,y) = \sum_n A_n \varphi_n^x(x)\varphi_n^y(y) \tag{4}$$

In order to take into account a priory known current density distribution with zero value at edges parallel to electric walls and singularity at the edges parallel to magnetic walls, we suggested a set of following separable expanding functions:

$$\varphi_n(x,y) = \begin{cases} \frac{\cos\left(\frac{4n\pi x}{w}\right)\sin\left(\frac{(2n+1)\pi}{l}\left(\frac{l}{2}-y\right)\right)}{w\sqrt{1-\left(\frac{2x}{w}\right)^2}} & |x| \le \frac{w}{2}, |y| \le \frac{l}{2} \\ 0 & |x| \ge \frac{w}{2}, |y| \ge \frac{l}{2} \end{cases} \tag{5}$$

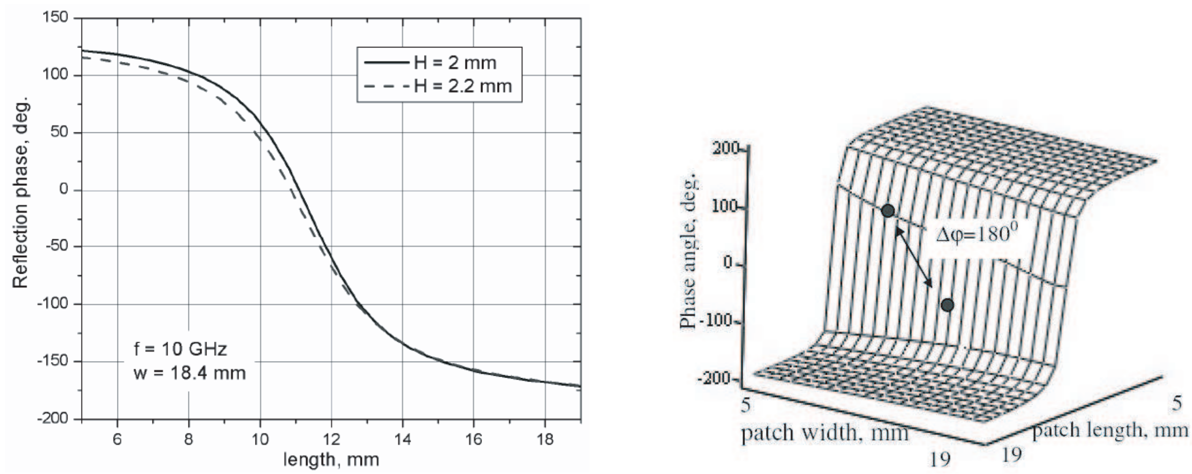
Here l is the length of rectangular patch, i. e., the length of a side corresponding to the direction of polarization, w is the width, and n is the number of basis function.

In order to find the vector of unknown coefficient $\{A_n\}$, Galerkin's procedure has to be implemented with weighting functions the same as the expansion functions. Then the phase of reflected field can easily be found.

Modifying (1) by addition of electric field $\tilde{\bar{\mathbf{E}}}_{tr} = R_s \cdot \tilde{\bar{\mathbf{J}}}$ we can take into account the losses due to finite conductivity of patches by using of equivalent surface impedance R_s.

3. Result and Discussion

Figure 2 demonstrates the results of calculation of phase angle depending on length of patch for two different thickness of single substrate. In order to adjust the sizes of patches to achieve a progressive phase for dual linear polarisation according to both twisting and focusing requirements the calculation of phase for polarizations along the both side of patches has to be done. If the dimensions of the patches are chosen in such a way that the absolute difference between the reflection angle of the two perpendicular polarization is 180° then the polarization of the wave incident on the patch tilted by 45° with respect to polarization of incident wave will be twisted by 90°.



(a) (b)

Figure 2: Results of calculation of phase angle depending on patch dimensions (a) Phase angle versus length of patch (different thickness of the substrate H), (b) Surface plot of phase angle in dependence on both dimensions of patch. Frequency 10 GHz, thickness of substrate $H = 2\,\mathrm{mm}$, dielectric permittivity 1.06, $L = 21\,\mathrm{mm}$, range of patch sizes 5–19 mm.

The phases of the reflected wave have to be attainable in the range of 0–360°. However the range of real array is, usually, less then 360 degree [4]. There are two opposite factor which influence the performance of antenna: the less the thickness of substrate and the bigger ε the greater range of obtainable phase, but the greater slope of curve and the higher technological requirements for patch size precision. Besides, for twisting reflectarray there is a gap of unreliazable phases. In Figure 3 the results of calculation of phase for bilayered substrate ($H_{upper} = 0.12\,\mathrm{MM}$ (dacron, $\varepsilon_r = 3.2$,), $H_{lower} = 1\,\mathrm{MM}$ (foam, $\varepsilon_r = 1.06$)) at 25 GHz are presented. In order to cover these two gaps about 55° each, one needs to use near values of patch sizes what will result in general error of phase approximately of 28°.

The validity of presented approach is confirmed by comparison with results of simulation by MS CST and Ansoft HFSS (Figure 4). Small discrepancy between data may be explained by more accuracy of methods of simulation based on FE and FDTD. However for acquisition of large arrays of patch dimensions the using MoM based code is more efficient.

Finally the developed code has been used for design of reflectarray antenna with diameter 300 mm operating at 25 GHz. More than 400 radiators were used to form a designed gain of 36 dB. Measured gain is around 32 dB, half-power beam width 2.7°, sidelobes are around 16°. Difference between measured and calculated characteristics can be partly explained by mentioned deviation of patch sizes to cover the unreliazable phase gap. Another reason of discrepancy is due to loss of some part of energy at the edge region of antennas plate. Moreover, the quasi-optical method to build a Fresnel reflector requires considering the dependence of phase angle also on angle of oblique incidence of wave on the edge patches.

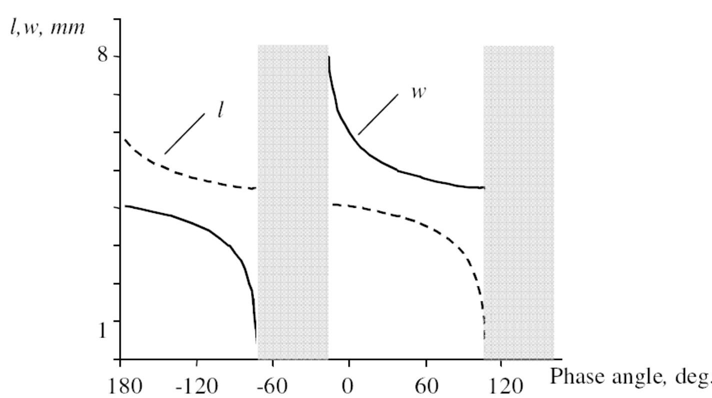


Figure 3: Dependence of phase angle on patch dimensions.

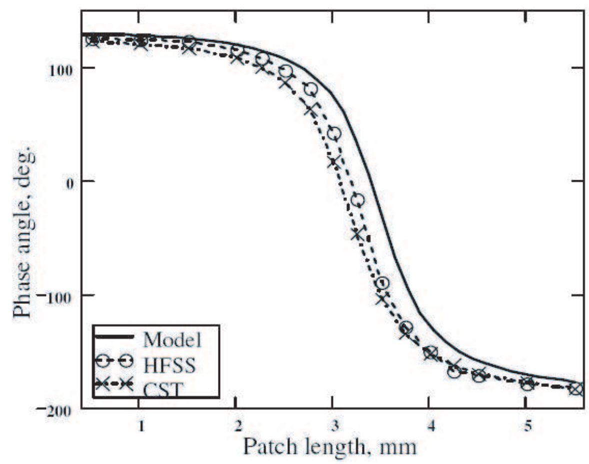


Figure 4: Comparison of simulation by presented model and by using microwave simulation packages. $H = 0.787\,\mathrm{mm}$, $\varepsilon_r = 2.22$, $L = 6\,\mathrm{mm}$, range of patch sizes from 0.6 to 5.6 mm, $f = 25\,\mathrm{GHz}$.

4. Conclusion

The design code based on the Spectral Domain Method of Moments (MoM) for lossy mutilayer periodic structures and normal incidence of a plane wave has been developed. Specific entire-domain basis functions have been proposed to achieve a high convergence and accuracy of MoM. Code developed is very efficient because it combines high calculation speed and high accuracy of full-wave analysis and is very promising for acquisition of large arrays of patch dimensions.

To validate the design method, a series of reflectarray antennas operating at different frequencies with and without twisting effect have been designed and manufactured [6]. A good agreement was obtained between predicted and measured radiation patterns for both polarizations. The measured gains were not less then 32 dB.

REFERENCES

1. Pilz, D. and W. Menzel, *Electron. Lett.*, Vol. 34, No. 9, 832–833, April 1998.
2. Menzel, W., D. Pilz, and M. AI- Tikrit, *IEEE Antenna's and Propagation Magazine*, Vol. 44, No. 3, 25–29, June 2002.
3. Encinar, J. A., *IEEE Trans. Antennas Propagat.*, Vol. 49, 1403–1410, October 2001.
4. Tsai, F.-C. E. and M. E. Bialkowski, *IEEE Trans. Antennas Propagat.*, Vol. 51, No. 10, 2953-2962, October 2003.
5. Pilz, D. and W. Menzel, *Asia-Pacific Microwave Conference Proc.*, Hong Kong, 225–227, December 2–5, 1997.
6. Parnes, M. D., V. D. Korolkov, M. S. Gashinova, I. A. Kolmakov, Ya. A. Kolmakov, O. G. Vendik, and T. V. Rossii, *Radioelectronics*, Vol. 1, Russian, 56–59, 2005.

Increasing Efficiency or Bandwidth of Electrically Small Transmit Antennas by Impedance Matching with Non-foster Circuits

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We have previously reported [1] the experimental realization of a stable, low-power high-Q VHF *non-Fostercapacitor*. This is an active one-port circuit element whose reactance-versus-frequency slope is *negative*. Our experimental verification showed that a non-Foster capacitor can reduce mismatch loss of electrically-small *receive* antennas over substantial bandwidths, thereby improving their sensitivity.

When used to cancel the reactance of an electrically-small *transmit* antenna, a non-Foster capacitor will be subjected to excessively high voltage in order to radiate even a moderate amount of power. We have previously reported [2] that we can mitigate this high voltage first by resonating the antenna with a conventional passive reactance, and then by canceling the reactance of this combination with a non-Foster tuned circuit. We also demonstrated the performance of stable non-Foster tuned-circuits. Initially, our tuned circuits were configured for operation in a simple, but relatively inefficient, class-A bias mode. Subsequently, we configured a composite device as a parallel connection of an NPN and a PNP transistor to operate in a more efficient class-B mode. Both class-A and class-B circuits were constructed with low-power devices.

Now, we report on our efforts to increase the power output capability and efficiency of these circuits to render them practical for realistic transmit antenna applications. We show that non-Foster transmit matching of electrically-small antennas is much more broadband than any passive matching circuit. The power efficiency of the antenna and its class-B biased non-Foster matching circuits is also better than that achievable for a passively matched transmit antenna.

REFERENCES

1. *IASTED International Conference on Antennas, Radar and Wave Propagation*, Banff, Canada, Arp. 2005.
2. *Twenty-Ninth Annual Antenna Applications Symposium*, Allerton Park IL, Sept. 21–23.