Phase Synthesis of Conformal Array Antennas

Leo I. Vaskelainen

***Abstract*—A new phase synthesis method for planar and conformal arrays is presented. The method is based on approximate linearized equations for small phase changes in the excitations, which can be determined by the least-squares method and used iteratively to obtain the final solutions for the excitation phases. Several different kinds of contoured beams can be used in the same array by using only phase shifters in the elements, as demonstrated in the calculated synthesis examples.**

***Index Terms*—Antenna arrays, antenna pattern synthesis, conformal antennas.**

I. Introduction

THE phase-synthesis problem is defined here as a problem in which the amplitudes of the excitations of an array are fixed (but not necessarily all ones) and the optimum values for the phases of the excitations are sought for synthesizing some shaped-gain function. This problem is highly nonlinear and usually a very general optimization method is needed. This kind of synthesis method is described by Bucci *et al.* [1]–[4]. Other general approaches usable for phase synthesis are the genetic algorithm [5] and simulated annealing techniques [6].

A related problem in which the excitation amplitudes are not fixed, but either they are constrained between certain values or the variation of the nearby excitations is minimized, is discussed in [7]–[9]. In linear equi-spaced arrays with shaped beams with filled nulls, there are several solutions which have the same amplitude in the array factor, but the distribution of the excitation values is changed by selecting between the two possible values of the complex roots of the array factor polynomial. A simple genetic algorithm can be used to optimize the excitation distribution [7], [8], and the nulls in the sidelobe region can be filled for a better excitation distribution as described by Rodriguez and Ares [9]. Buckley has also introduced a method for the design shaped beam antenna patterns of a linear array with constrained current elements [10]. An excitation constrained synthesis method has been introduced for circular arrays [11] and a genetic algorithm is used to design thinned and digitally phase-shifted pencil-beam antennas [12], [13].

Generally, because of the local minima, the phase-synthesis problem is more conveniently solved by using methods which Johnson and Rahmat-Samii [14] call global techniques (random walk, simulated annealing, genetic algorithm). In the field synthesis problem (in which the phase of the destination function is fixed) the problem is linear and a least-squares solution can be used. The least-squares solution is closely related to more common local optimization techniques (conjugate gradient method, quasi-Newton method, etc.). The author has used the least-squares method iteratively to solve the power synthesis problem (in which the phase of the destination function is not constrained) [15]. In fact, the phase-synthesis problem can be solved by the iterative least-squares method [15], [16]. A weight correction method applied during the iteration usually (but not always) solves the problem of local minima [16].

In Section II of this paper, the author linearizes the problem and uses the least-squares method to solve the phase-synthesis problem. The approximate linearized equations are formed by using the same kind of linear approximations for differential phase changes as is used for phase synthesis of nulls [17], [21]. In directions, in which the value of the destination function is zero, a linear set of equations can be written for the differential phase changes. In other directions, approximate linear equations are written for the differential phase changes. A least-squares solution for this kind of set of linear equations can be used iteratively to find the final phase-synthesis result. The procedure for solving the problem is very similar to that used in the iterative least-squares synthesis method [15]. The solution converges quickly, but essentially the method is a local optimizaton method. The weight correction method can be used during the iteration, but it is not as effective as with nonconstrained excitation amplitudes.

The solution for the phase synthesis problem is usually ambiguous. The problems caused by this factor are discussed in Section III.

In Section IV, some examples of phase synthesis of a conformal array antenna are presented.

II. Theory

First, we write the directivity vector $\mathbf{E}_0 = [E_0(\theta_1, \varphi_1) \ \cdots \ E_0(\theta_K, \varphi_K)]^T$ of an array having N elements in K directions (θ_k, φ_k) in matrix form, using the same formulation as in [15]

$$\mathbf{E}_0 = \mathbf{X}_0 \mathbf{A}_0 \tag{1}$$

where A_0 is the excitation vector $[a_1 \ \cdots \ a_N]^T$ of the array (T is the transpose of a matrix) and the elements X_{ij} of the geometry-dependent matrix $\mathbf{X}_0$ are

$$X_{ij} = g_j(\theta_i, \varphi_i) \times e^{jk_0[\sin(\theta_i)\cos(\varphi_i)x_j + \sin(\theta_i)\sin(\varphi_i)y_j + \cos(\theta_i)z_j]}. \tag{2}$$

x_n, y_n and z_n are the coordinates of the position vectors of the array elements and $g_j(\theta, \varphi)$ are element radiation patterns, which do not need to be identical. k_0 is the wave number $2\pi/\lambda$.

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The phases of the values $\mathbf{E}_0$ and $\mathbf{A}_0$ should be chosen so that

$$(\mathrm{abs}(\mathbf{E}_0)-\rho_0\mathbf{D}_0)^T(\mathrm{abs}(\mathbf{E}_0)-\rho_0\mathbf{D}_0)=\mathrm{minimum} \quad (3)$$

where ρ_0 is an unknown constant, $\mathbf{D}_0$ is a vector giving the destination values for the directivity function, and abs-function takes the absolute values of the elements of a matrix.

We can fix the phase of one excitation value without invalidating the generality of the solution. If we choose the phase of excitation of the first element a_1 to be fixed (zero), we can divide the matrices in (1) into partial matrices

$$\mathbf{X}_0=[\mathbf{X}_1 \quad \mathbf{X}], \quad \mathbf{A}_0=[a_1 \quad \mathbf{A}^T]^T. \quad (4)$$

Next, we arrange the rows and columns of the matrices so that the rows corresponding to directions where $D_0(\theta_i,\varphi_i)=0$ are first and divide the matrices into submatrices in the same way

$$\mathbf{X}_0=\begin{bmatrix}\mathbf{X}_{1Z} & \mathbf{X}_Z\\ \mathbf{X}_{1N} & \mathbf{X}_N\end{bmatrix}, \quad \mathbf{D}_0=\begin{bmatrix}\mathbf{0}\\ \mathbf{D}_N\end{bmatrix}, \quad \mathbf{E}_0=\begin{bmatrix}\mathbf{E}_z\\ \mathbf{E}_N\end{bmatrix}. \quad (5)$$

If we now change the excitation values $\mathbf{A}$ by adding the small amounts $\mathbf{dA}$, which cause small changes $\mathbf{dE}_0=[\mathbf{dE}_Z^T \; \mathbf{dE}_N^T]^T$ in the directivity values $\mathbf{E}_0$, (1) can be written

$$\begin{bmatrix}\mathbf{E}_Z+\mathbf{dE}_Z\\ \mathbf{E}_N+\mathbf{dE}_N\end{bmatrix}=\begin{bmatrix}\mathbf{X}_{1Z} & \mathbf{X}_Z\\ \mathbf{X}_{1N} & \mathbf{X}_N\end{bmatrix}\begin{bmatrix}a_1\\ \mathbf{A}+\mathbf{dA}\end{bmatrix}. \quad (6)$$

Later on we also need the real and imaginary parts separated

$$\begin{aligned}\mathbf{X}_{1Z}&=\mathbf{X}_{1ZR}+j\mathbf{X}_{1ZI}, \quad \mathbf{X}_Z=\mathbf{X}_{ZR}+j\mathbf{X}_{ZI}\\ \mathbf{X}_{1N}&=\mathbf{X}_{1NR}+j\mathbf{X}_{1NI}, \quad \mathbf{X}_N=\mathbf{X}_{NR}+j\mathbf{X}_{NI}\\ a_1&=a_{1R}+ja_{1I}, \quad \mathbf{A}=\mathbf{A}_R+j\mathbf{A}_I.\end{aligned} \quad (7)$$

Three steps are needed for solving the approximate phases of the excitations.

Step 1) The two first terms of the Taylor expansions of the excitation value phase factors are used. As in [21], this means that the excitation changes can be written thus

$$\mathbf{dA}=j(\mathrm{diag}(\mathbf{A}_R)+j\,\mathrm{diag}(\mathbf{A}_I))\mathbf{p} \quad (8)$$

where $\mathrm{diag}(\mathbf{A})$-function is a square-matrix with vector $\mathbf{A}$ in its diagonal and $\mathbf{p}$ is a vector of unknown real-valued variables.

Step 2) From (6) and (8) $\mathbf{E}_Z+\mathbf{dE}_Z$, the new gain value in directions in which the destination gain is zero can be calculated using the definitions in (7). The requirement that $\mathbf{E}_Z+\mathbf{dE}_Z=\mathbf{0}$ leads to the equation

$$\begin{bmatrix}\Xi_{ZR}\\ -\Xi_{ZI}\end{bmatrix}\mathbf{p}=\begin{bmatrix}\Psi_{ZR}\\ \Psi_{ZI}\end{bmatrix} \quad (9)$$

where

$$\begin{aligned}\Xi_{ZR}&=\mathbf{X}_{ZR}\,\mathrm{diag}(\mathbf{A}_I)+\mathbf{X}_{ZI}\,\mathrm{diag}(\mathbf{A}_R)\\ \Xi_{ZI}&=\mathbf{X}_{ZR}\,\mathrm{diag}(\mathbf{A}_R)-\mathbf{X}_{ZI}\,\mathrm{diag}(\mathbf{A}_I)\end{aligned} \quad (10)$$

and

$$\begin{aligned}\Psi_{ZR}&=\mathbf{X}_{1ZR}a_{1R}-\mathbf{X}_{1ZI}a_{1I}+\mathbf{X}_{ZR}\mathbf{A}_R-\mathbf{X}_{ZI}\mathbf{A}_I\\ \Psi_{ZI}&=\mathbf{X}_{1ZR}a_{1I}+\mathbf{X}_{1ZI}a_{1R}+\mathbf{X}_{ZR}\mathbf{A}_I+\mathbf{X}_{ZI}\mathbf{A}_R.\end{aligned} \quad (11)$$

When the phase condition used in step 3 is considered apart when the destination gain function is zero (or very small), the overall nonlinearity of problem is reduced and numerical problems are avoided.

Step 3) The third step is to calculate the gain changes $\mathbf{dE}_N$ in directions in which the destination gain has nonzero values, calculate from these gain changes the components which are in phase with the gain values $\mathbf{E}_N$, and require that the absolute values of $\mathbf{E}_N$ added to these in-phase components of the gain changes equal the destination gain values $\mathbf{D}_N$ multiplied by the unknown constant ρ_0.

This is a linearized approximation and for small phase changes, this approximation causes a second-order error in the absolute value of the gain function, and during iteration the error decreases quickly.

Using (6), -(8)

$$\mathbf{dE}_N=-\Xi_{NR}\mathbf{p}+j\Xi_{NI}\mathbf{p} \quad (12)$$

where

$$\begin{aligned}\Xi_{NR}&=\mathbf{X}_{NR}\,\mathrm{diag}(\mathbf{A}_I)+\mathbf{X}_{NI}\,\mathrm{diag}(\mathbf{A}_R)\\ \Xi_{NI}&=\mathbf{X}_{NR}\,\mathrm{diag}(\mathbf{A}_R)-\mathbf{X}_{NI}\,\mathrm{diag}(\mathbf{A}_I).\end{aligned} \quad (13)$$

If we write the phasing matrices

$$\begin{aligned}\Theta_R&=\mathrm{diag}(\cos(\mathrm{angle}(\mathbf{E}_N)))\\ \Theta_I&=\mathrm{diag}(\sin(\mathrm{angle}(\mathbf{E}_N)))\end{aligned} \quad (14)$$

we can calculate the components $\mathbf{dE}_{NN}$ of $\mathbf{dE}_N$, which are in phase with $\mathbf{E}_N$ from (12)

$$\mathbf{dE}_{NN}=-\Phi_N\mathbf{p} \quad (15)$$

where

$$\Phi_N=\Theta_R\Xi_{NR}-\Theta_I\Xi_{NI}. \quad (16)$$

The angle function of a matrix takes the phases of the complex elements of the matrix. When the excitations are changed, the gain values $\mathbf{E}_N$ will also change, and it is not possible to require that the absolute values of $\mathbf{E}_N+\mathbf{dE}_N$ equal the destination values $\mathbf{D}_N$. What we can do is to approximate the absolute value of the changed gain function by computing $\mathrm{abs}(\mathbf{E}_N)+\mathbf{dE}_{NN}$ and require that these values equal the destination gain values $\mathbf{D}_N$ multiplied by the unknown constant ρ_0

$$\begin{aligned}\mathrm{abs}(\mathbf{E}_N)+\mathbf{dE}_{NN}&=\mathrm{abs}(\mathbf{E}_N)-(\Theta_R\Xi_{NR}-\Theta_I\Xi_{NI})\mathbf{p}\\ &=\rho_0\mathbf{D}_N\end{aligned} \quad (17)$$

which can be written thus

$$[\Phi_N \quad \mathbf{D}_N]\begin{bmatrix}\mathbf{p}\\ \rho_0\end{bmatrix}=\mathrm{abs}(\mathbf{E}_N). \quad (18)$$

Finally, we can combine (9) and (18) for the unknown $\mathbf{p}$ and ρ_0

$$\begin{bmatrix} \Xi_{ZR} & \mathbf{0} \\ -\Xi_{ZI} & \mathbf{0} \\ \Phi_N & \mathbf{D}_N \end{bmatrix} \begin{bmatrix} \mathbf{p} \\ \rho_0 \end{bmatrix} = \begin{bmatrix} \Psi_{ZR} \\ \Psi_{ZI} \\ \text{abs}(\mathbf{E}_N) \end{bmatrix}. \quad (19)$$

Equation (19) is now written in the same form $\mathbf{X}_p[\mathbf{p}^T \ \rho_0]^T = \mathbf{E}_p$ for the unknown $\mathbf{p}$ vector as (1) is for the $\mathbf{A}_0$ excitation values and the least-squares solution $[\mathbf{p}^T \ \rho_0]^T = (\mathbf{X}_p^H\mathbf{W}\mathbf{X}_p)^{-1}(\mathbf{X}_p^H\mathbf{W}\mathbf{E}_p)$ can be calculated readily. If the weight values are written for different directions, these weight values can be written in matrices $\mathbf{W}_Z$, which correspond to directions in which $D_0(\theta_i, \varphi_i) = 0$, and $\mathbf{W}_N$ corresponding directions in which $D_0(\theta_i, \varphi_i) \neq 0$. The weight matrix for solution of (19) is then

$$\mathbf{W} = \begin{bmatrix} \mathbf{W}_Z/2 & \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \mathbf{W}_Z/2 & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \mathbf{W}_N \end{bmatrix}. \quad (20)$$

As in [21], the elements of real-valued vector $\mathbf{p}$ can be used iteratively to determine the phases of the excitations by replacing $a_i \rightarrow a_i e^{j\arctan(p_i)}, i = 2 \ldots N$ and calculating new values for $\mathbf{p}$ using (19) with the corrected Φ_N and $\mathbf{E}_N$ matrices.

When (19) is used, the same kind of optimizations can be achieved as in normal iterative least-squares synthesis. Both polarization components (linear or circular) and several frequencies can be included in the synthesis as the author has done in [15]. The weight values can be normalized by the reference error (or desired error) values, and the weight values can be corrected according to these normalized error values during the iteration, as described in [15] and [16]. Of course, the decreased degree of freedom usually makes it very difficult to optimize several properties of the gain function using only the phases of the excitations. If one-mode elements are used, the element gain functions can be assumed to be independent of the mutual impedances and the mutual couplings can be described with a mutual impedance matrix or some other linear coefficients [18], [19] and the mutual couplings can be taken into account in the synthesis process by simply transforming the $\mathbf{X}_0$ matrix in (1) [18].

III. Discussion of the Ambiguity of the Solution

The determination of the phases of the excitations by the method described in Section II converges in few iterations to the nearest minimum of the minimizing function in (2). In the phase synthesis problem, this function very often may have several local minima. In some cases it may indeed have some solutions which are consistently acceptable when measured by the minimizing function.

The reason for this is that the solution for a phase synthesis problem usually has a phase distribution (for element excitations) in which different parts of a shaped beam are formed by using different parts of the array as "subarrays" phased in different directions. That kind of formation of subarrays can be done in different ways, leading to an ambiguous solution to the problem.

If the number of elements in the array is excessively great, there may be solutions which contain "holes" (very narrow nulls). These nulls may be so narrow that they don't have much impact on the minimizing function.

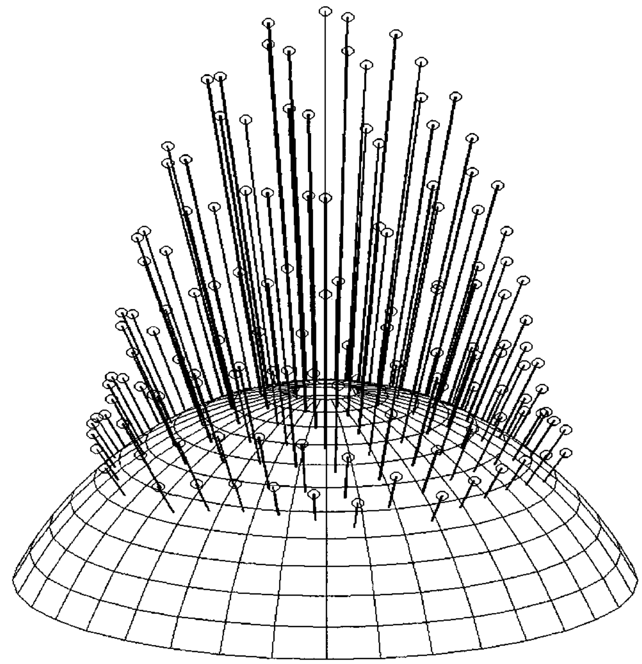

Fig. 1. 1136 circular patches on a sphere. The places, directions, and amplitudes of the vectors show the places and axial directions of the elements and the amplitudes of the excitations.

No theoretical method is presented in this paper for avoiding these kinds of local minima. Some principles have proved useful in practical synthesis problems.

- A reasonable guess for the phase distribution must be used. The starting values for the iteration can be formed by using subarrays pointing in different directions.
- The solution will be less ambiguous if the array is smaller or a broad-band solution is being sought. A solution in several frequencies and for a smaller partial array can be used for starting values in which case more elements can be added to the array before the final optimization.

At this point we can also note that the solution for conformal arrays is less ambiguous than the solution for planar arrays.

The examples in Section IV are calculated using these principles.

IV. Phase Synthesis Examples

All the examples are calculated using the same array geometry, in which 136 circular patch elements are located on the sphere in an icosahedron grid, as presented in Fig. 1. The space between elements is about 0.65λ and the axes of the elements are positioned to give the minimun cross polarization in direction $\theta = 90°$ and $\varphi = 0°$. The radius of the sphere is 6.458λ. The amplitude distribution of the excitations is cosine with a 0.1 pedestal, as seen in Fig. 1.

For elements, the circular-polarized circular-patch element model in [15, eqs. (A.1), (A.2)] is used.

In Figs. 2–4, three types of right-hand circular polarization (RHCP) gain functions are synthesized using this same geometry of the array and amplitude distribution. Triangular and

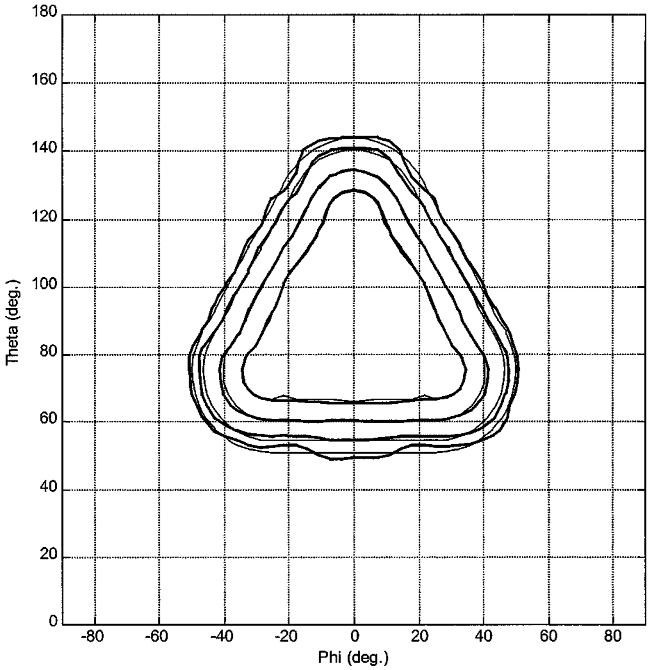

Fig. 2. Contour-plot of the gain function “TRIANGLE” (RHCP-component). Levels −20 dB, −15 dB, −8 dB, and −3 dB compared to the maximum gain are presented. Thin line for the destination function and thick line for the synthesized result.

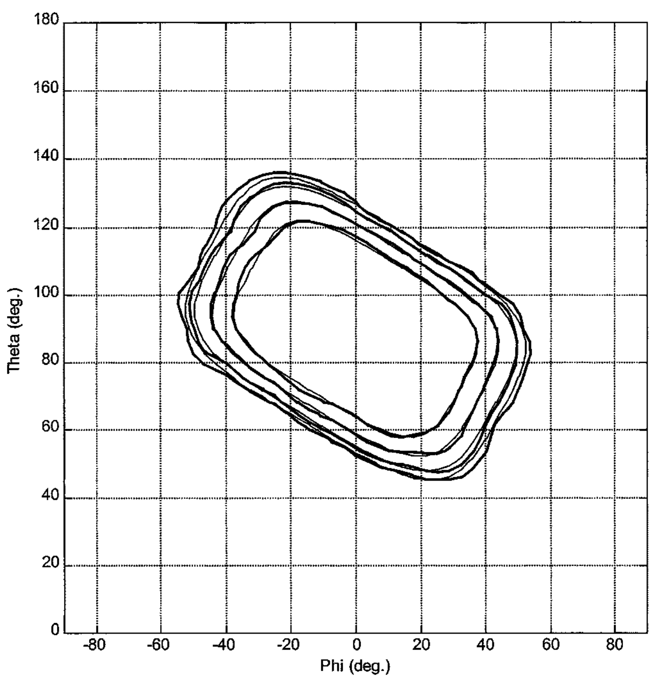

Fig. 3. Contour-plot of the gain function “RECTANGLE” (RHCP-component). Levels −20 dB, −15 dB, −8 dB, and −3 dB compared to the maximum gain are presented. Thin line for the destination function and thick line for the synthesized result.

rotated rectangular wide, contoured beams can be synthesized rather well by changing the phases of the excitations. In Fig. 4, a wide contoured beam in which a logo-shaped bulge about 6 dB high is added on the top of the gain function and is also successfully synthesized by using the phase changes of the excitations.

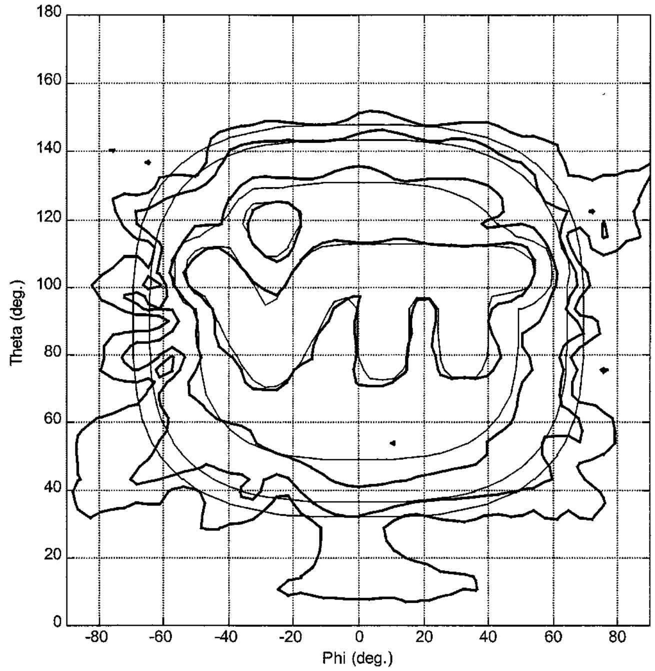

Fig. 4. Contour-plot of the gain function “VTTLOGO” (RHCP-component). Levels −20 dB, −15 dB, −8 dB, and −2 dB compared to the maximum gain are presented. Thin line for the destination function and thick line for the synthesized result.

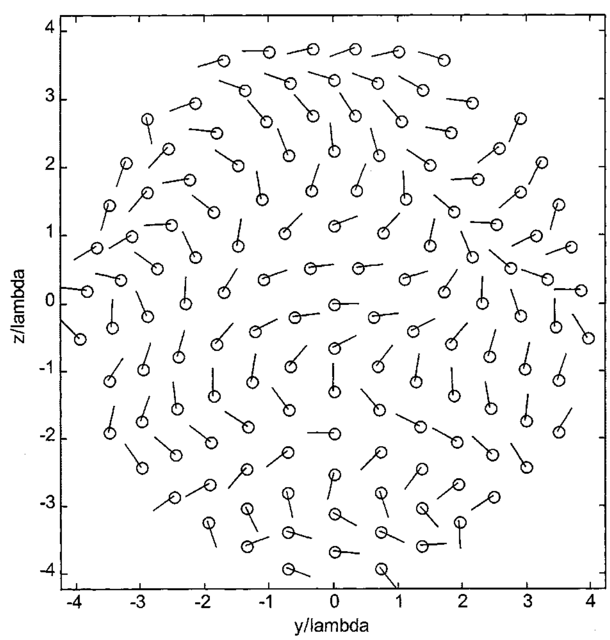

Fig. 5. The projections of locations of elements on y-z-plane. The directions of the vectors show the phases of the excitations for the phase-synthesized “TRIANGLE” gain function.

In all calculations of the examples the starting values for the final iteration are found by using pre-iterations in several steps.

1) The starting values for all phases of the excitations are zeros.
2) Thirty-one elements around the center of the array are selected and the phase synthesis is made for this small array using three frequencies in a 15% bandwidth.

3) Step 2 is repeated with 61 elements, 91 elements, 116 elements and finally with all elements. The starting values for the new elements are corrected by the phase changes of the nearest already synthesized element excitations.
4) Steps 2 and 3 are repeated, but using three frequencies in a 10% bandwith.
5) The result of step 4 is used as the starting value for the final phase synthesis.

In the synthesis of Fig. 2 ("TRIANGLE" gain function) six iterations are used in step 2, $9+10+12+15$ iterations in step 3, $6+9+10+12+20$ iterations in step 4. This procedure may not be optimal considering the calculation time (109 iterations with different sizes of arrays are used), but it gives reliably good results when the weight values correction is used during the iteration, as mentioned in Section II. In the final phase synthesis, only 18 iterations are needed. The number of iterations in calculation of Figs. 3 and 4 are rather similar.

No constraints on the cross-polarization components are used. The cross-polarization levels are -28.3 dB (Fig. 2), -25.6 dB (Fig. 3) and -22.6 dB (Fig. 4). Only very insignificant decrease of the cross-polarization level can be achieved when the both polarization components are included in the phase synthesis.

These examples demonstrate that the phase-synthesis method can be an effective tool in the design of planar or conformal array antennas. Different kinds of contoured beams can be produced in the same array by using only the phase shifters in the elements.

In Fig. 5, the phases of the excitations for the triangular beam are shown. The phase changes continuously and it can be seen that the different parts of the array are used to radiate in different directions so as to form the destination contoured beam.

V. Conclusion

A new iterative phase-synthesis method is described for planar and conformal arrays. This method is very straightforward and yields a fast convergence, but some additional procedures are needed to avoid ambiguous solutions. As in the iterative least-squares method, in this phase-synthesis method also several frequencies can be included in the synthesis problem, the synthesis can be performed for any polarization component and even the mutual couplings can, in some cases, be taken into account.

The calculated synthesis examples indicate that this method is an effective tool for practical phase synthesis problems. By using this method, different kind of contoured beams can be generated in the same array by using phase-only control in the elements.

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