Synthesis of Linear Antenna Array for 4G Mobile Communication Systems

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***Abstract*— This paper presents the application of optimization methods for the synthesis of a linear array to operate in the frequency range of 4G technology in Brazil. The desired pattern shall exhibit squared-cosecant shape, so as to provide uniform distribution of power inside the base station cell and to reduce co-channel interference. Such an array is well suited to operate as a radio base station of mobile communications systems. The synthesis is performed by a combination of optimization methods: the genetic algorithm, which is used for the initial global search, and the sequential quadratic programming, which is applied for local refinement of the solution. This approach allows faster convergence than using only one kind of optimization method. The technique is demonstrated for a linear array of isotropic elements and, subsequently, for an array of *E*-shaped microstrip antennas.**

Keywords*— *Beam shaping; genetic algorithm; sequencial quadratic programming; 4G systems; E-shaped microstrip antennas.

I. INTRODUCTION

The appearance of mobile radio systems has brought great convenience and flexibility, as they allow the communication among mobile users. The capacity of such systems is mainly limited by the signal-to-interference ratio produced by co-channel adjacent cells [1]. Some techniques have been proposed for reducing co-channel interference aiming at increasing the system capacity, such as cell splitting and sectoring [2], [3]. In [4], an approach based on shaping the antenna pattern has been proposed, where a radiation pattern characteristic following a squared cosecant function was considered to provide uniform distribution of power along the cell. Radiated power in the direction of co-channel cells was also minimized, hence reducing interference. The drawback of the formulation proposed in [4] is that the synthesized pattern was obtained considering an array composed of isotropic elements. For practical purposes, however, the array elements do not have isotropic radiation characteristic. Therefore, the real element pattern must be considered in order to obtain reliable results.

In contrast to [4], this paper presents the synthesis of a linear array by considering the radiation characteristic of the array element during the optimization process. The array element is an *E*-shaped microstrip antenna designed to operate in the frequency range allocated for the 4G technology in Brazil, which corresponds to the band of 2.50-2.69 GHz. As a special feature, the array shall exhibit squared cosecant-shaped radiation pattern.

In the next section, the design of the *E*-shaped antenna will be discussed. In Section III, the optimization techniques will be briefly introduced. Section IV presents numerical results for the synthesis of linear arrays composed of isotropic and *E*-shaped elements. Comparison between both results is shown to demonstrate the need of including the real radiation pattern of the array element in the optimization process.

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II. DESIGN OF AN *E*-SHAPED ANTENNA FOR 4G MOBILE COMMUNICATION SYSTEMS

Microstrip antennas have been widely applied in several areas the last years, such as satellite navigation, and aerospace and communication systems. Some interesting features, such as low cost for mass production, low weight, and ease of construction and integration with other microwave elements, make the microstrip technology a potential candidate for the proposed application.

It is well known that ordinary rectangularly or circularly shaped microstrip antennas are inherently narrow-banded. In order to counter this drawback, several techniques have been proposed in the literature. One easy technique to broaden the impedance bandwidth is to introduce a pair of slits in the rectangular patch, hence resulting in an *E*-shaped radiator. This allows the excitation of two modes that resonate in slightly different frequencies [5]-[7]. The *E*-shaped antenna is schematically shown in Fig. 1 and Fig. 2.

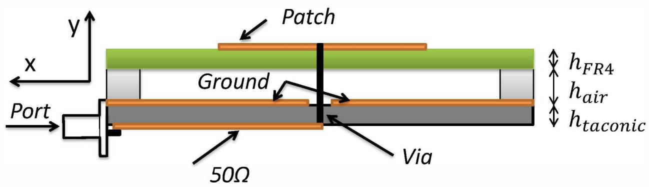

Fig. 1. Side view of the *E*-shaped antenna.

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In this design, two laminates were used: FR4 and Taconic TLC-338. An air layer 4.8-mm thick was inserted between both laminates in order to increase the bandwidth. The FR4 laminate has the following characteristics: thickness (h_{FR4}) of 1.54 mm, dielectric constant of 4.1 and dielectric loss tangent of 0.02. The Taconic TLC-338 features are: thickness ($h_{taconic}$) of 1.524 mm, dielectric constant of 3.56 and dielectric loss tangent of 0.0034.

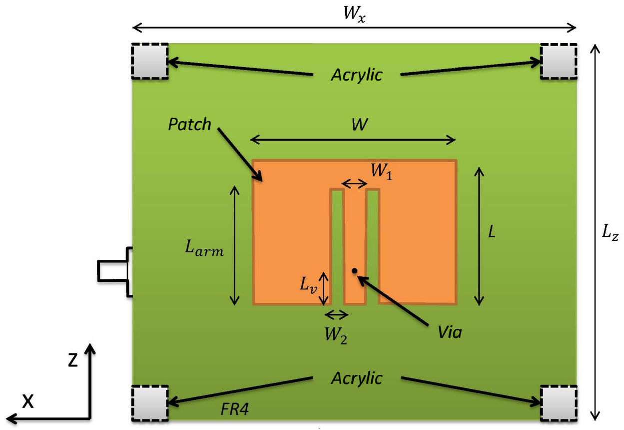

Fig. 2. Top view of the *E*-shaped antenna.

The antenna has been optimized using ANSYS *HFSS* software and the following dimensions were finally obtained: W = 53.82 mm, L= 38.37 mm, W_1= 3.48 mm, W_2 = 5.89 mm, L_v= 9 mm, L_{arm}= 30.82 mm, W_x= 117.22 mm and L_z = 101.77 mm. The via diameter (d_{via}) was 1.1 mm. In order to assess the sensitivity to the fabrication imperfections, the air layer thickness was varied. The results can be viewed in Fig. 3, where one can verify that the bandwidth operation is very sensitive to the change of the air layer thickness. Variation in the dielectric constant of the FR4 laminate has also been studied, but only a slight frequency deviation has been observed.

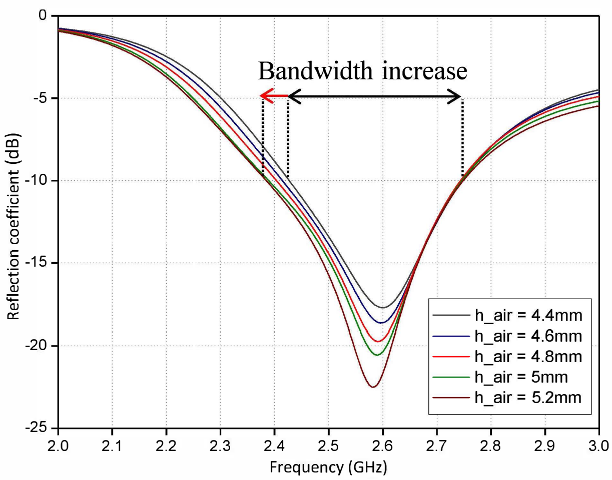

Fig. 3. Influence of the thickness of the air layer on the reflection coefficient.

During the design process, the antenna dimensions were optimized so that a wider operation bandwidth than the one specified for 4G systems was obtained. Hence, small deviations in the fabrication process do not affect the antenna performance strongly.

The *E*-shaped antenna radiation patterns at 2.595 GHz were calculated with *HFSS* and are shown in Fig. 4 and Fig. 5. In the *E*-plane, the radiation pattern exhibits larger radiation intensity below the normal direction related to the antenna surface. Since the desired squared cosecant-shaped pattern concentrates most of the energy below the horizon, this feature facilitates the synthesis process. The antenna was fabricated and a photo of the prototype is shown in Fig 6.

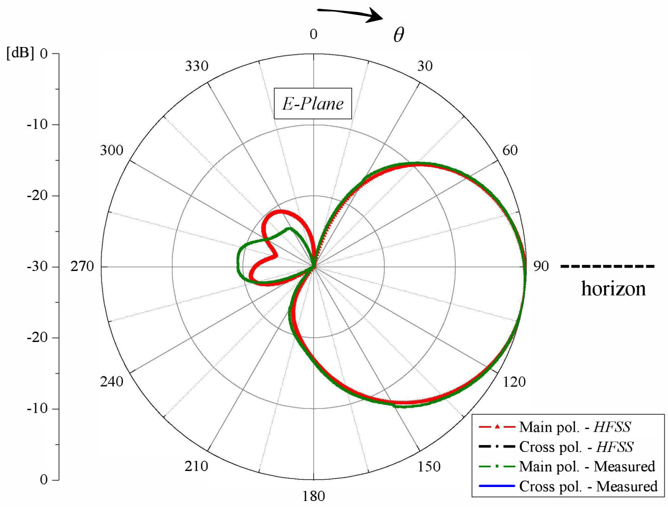

Fig. 4. Radiation pattern of the *E*-shaped antenna in the *zy* plane.

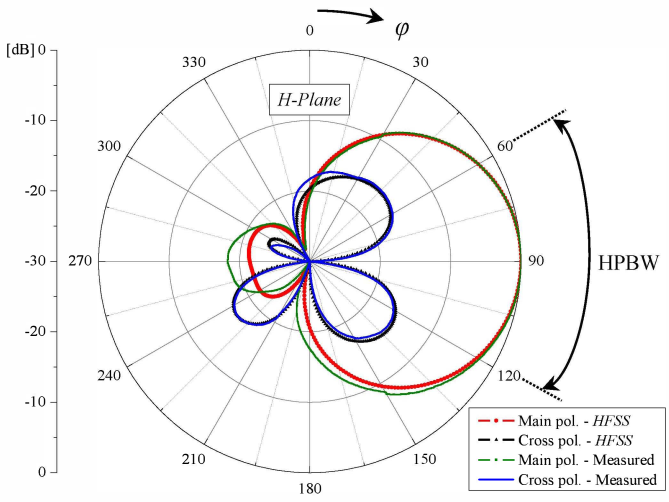

Fig. 5. Radiation pattern of the *E*-shaped antenna in the *xy* plane.

In Fig. 7, the simulated and measured results for the reflection coefficient are plotted. Very good agreement between the results is verified. The simulated and measured results for the input impedance in the range 2 - 3 GHz are shown in Fig. 8 in Smith chart format. The wide bandwidth is obtained because the impedance curves exhibit a loop that turns around the center of the Smith chart.

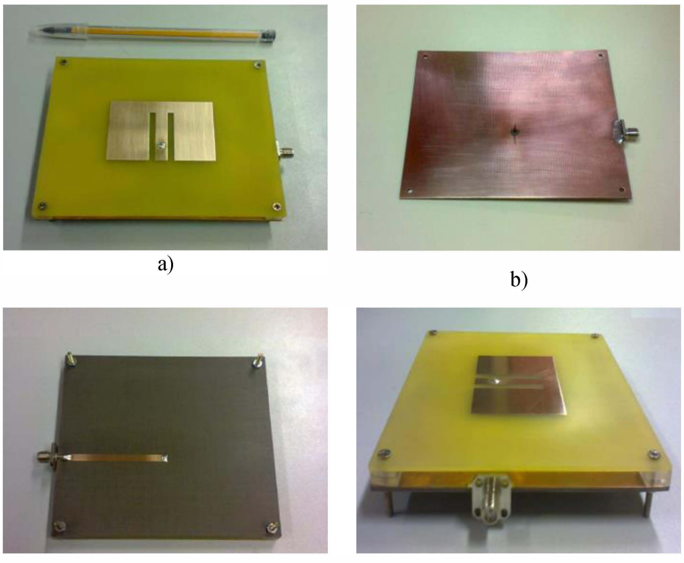

Fig. 6. Fabricated prototype: a) Top view; b) Ground plane; c) bottom view prior to the assembly and d) Final configuration of the *E*-shaped antenna.

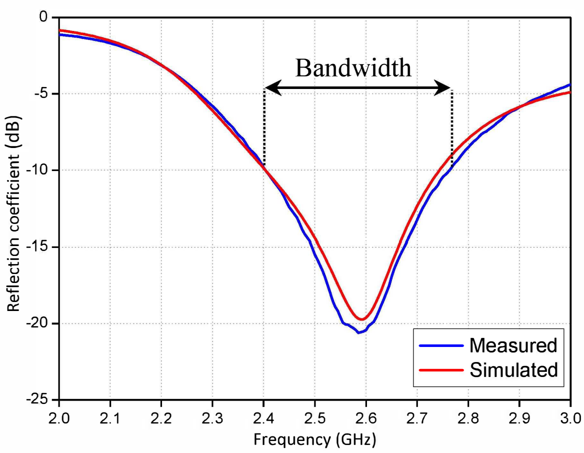

Fig.7. Simulated and measured reflection coefficient (S_{11}) at the antenna input.

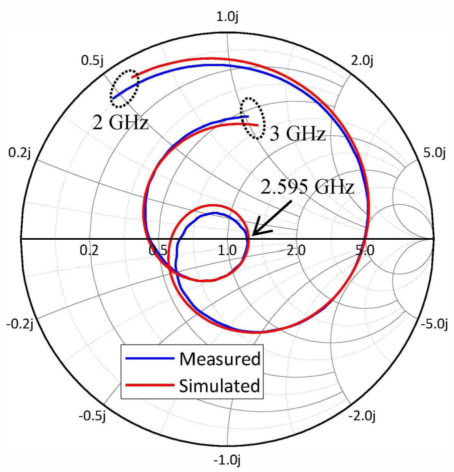

Fig. 8. Simulated and measured input impedance.

III. GENETIC ALGORITHM AND SEQUENCIAL QUADRATIC PROGRAMMING

In this paper, two optimization methods have been employed to find the excitation coefficients (amplitudes and phases) of linear arrays. Initially, genetic algorithm (GA) was used to search for a set of coefficients that correspond approximately to the specified pattern, then sequential quadratic programming (SQP) was applied to refine the solution. The SQP is always started with the best individual obtained by the GA.

The concept of the GA was formulated by Holland and applied in [8] to obtain the excitation coefficients of arrays, aiming at steering the main beam in a specific direction and to control the sidelobe levels (SLL). The SQP is a method of nonlinear programming, which performs optimizations based on gradients and approximates the sequential nonlinear programming problem as a quadratic programming problem [9]. This allows finding the best local solution.

In the optimization with GA, individuals in the population are treated as potential solutions to the problem, where an initial population suffers crossover and mutation over generations. With the evolution of the individuals, the GA finds better results to solve the problem. The fitness of each individual is given based on the evaluation of its chromosome as a cost function. The cost function used in this study for both methods is based on the error, calculated by

$$e(\theta) = S(\theta) - E(\theta), \theta \in [0°, 180°] \,, \tag{1}$$

where $S(\theta)$ is the desired pattern as shown in Fig. 9 and $E(\theta)$ is the obtained pattern (dependent on the chromosome of each individual), given by

$$E(\theta) = w \,.\, v(\theta). \tag{2}$$

In (2), w is the excitation vector (chromosome) of the N-element array, denoted by,

$$w = [w_1 \;\; w_2 \;\; w_3 \; .... \;\; w_N \,] \,, \tag{3}$$

whereas $v(\theta)$ is the vector associated with the pattern of the array elements, defined by

$$v(\theta) = \begin{bmatrix} g_1(\theta).\, e^{j\psi_1} \\ g_2(\theta).\, e^{j\psi_2} \\ g_3(\theta).\, e^{j\psi_3} \\ \vdots \\ g_N(\theta).\, e^{j\psi_N} \end{bmatrix}, \tag{4}$$

with g_i representing the pattern of each array element.

The estimated pattern $E(\theta)$ is analyzed in two regions: the first one (m = 1) is defined by the region of the side lobes, whereas the second one (m = 2) is the region of the csc² $(\theta - 90°)$ function. In Fig. 9, the region of the side lobes is defined between θ_a and θ_b, whereas the cosecant region is defined between θ_c and θ_d , and θ_n is the normalization angle of the cosecant function. The angle θ_n is introduced due to the fact that the cosecant function as defined in Fig. 9 tends to infinity as θ approaches 90°. The mean square error for the side lobe region (R_1) and for the region of the squared cosecant function (R_2) is given by

$$R_m = \left[\frac{1}{L_m}\sum_{i=1}^{L_m}|e(\theta_i)|^2\right]^{\frac{1}{2}} . \quad (5)$$

Considering the number of samples L_1 in the sidelobe region and L_2 in the cosecant region, the final cost function is computed by

$$fitness = R_1.P_1 + R_2.P_2 , \quad (6)$$

where P_1 and P_2 are the weights defined for each region.

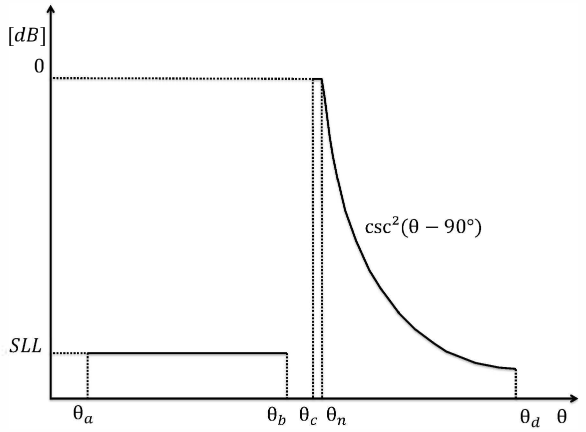

Fig. 9. Normalized desired pattern for uniform power distribution in the cell.

IV. RESULTS

Initially, the number of elements necessary to obtain the desired radiation pattern has been determined. For this purpose, linear arrays with 10, 15 and 20 isotropic elements uniformly spaced (0.5 λ) along the vertical direction were considered.

The GA was initialized with a population of 70 individuals, with crossover probability of 85%, mutation of 8%, and 7% of individuals treated as elite. The limits of variation of the genes were set to [0, 1] for the amplitude and [-π, π] for the phase. The linear arrays with isotropic elements were optimized over 500 generations with the GA, followed by the SQP started with the best individual. The side lobe regions were defined between [0°, 68°], [0°, 76°] and [0°, 84°] for the arrays with 10, 15 and 20 elements respectively, with maximum allowed SLL 42 dB below the maximum of the pattern. The squared cosecant region was defined in the interval [92°, 180°] and the normalization angle of the cosecant function was set to 95°. Both regions have been assigned with equal weights, i.e. $P_1 = P_2$ in (6). The desired pattern and the optimization results are shown in Fig. 10.

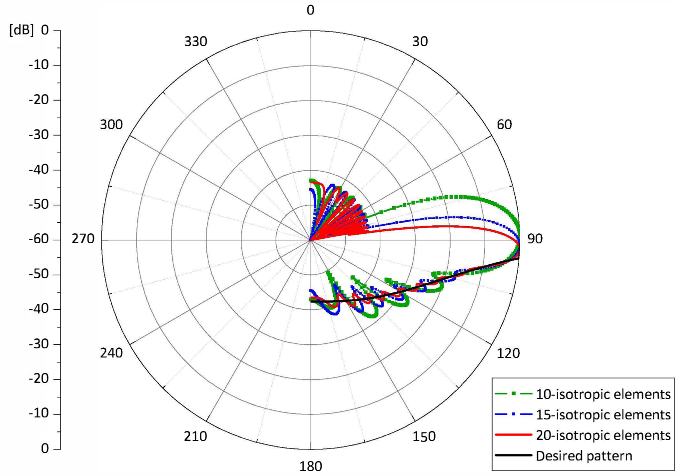

Fig.10. Radiation patterns for different number of the elements.

The arrays with 10, 15 and 20 isotropic elements produced patterns with ripple in the squared-cosecant region of 9.22 dB, 5.19 dB and 2.2 dB respectively. Moreover, it can be seen that the power radiated above the horizon increases when the elements number of the array is reduced.

In order to obtain a nearly constant power distribution in the cell of a given base station, the maximum ripple of less than 1 dB has been set as the maximum deviation from the ideal squared-cosecant function. After further analyses, it came out that this is only fulfilled with a linear array with at least 24 elements. An array with 24 elements results in the following dimensions: W_x= 117.22 mm and L_z = 1383.97 mm. These dimensions are similar to those of commercial radio base antennas.

By setting up the squared cosecant region in the interval [92°, 180°], the normalization angle of the cosecant function to 95° and the side lobe region was defined between [0°, 86°] for the optimization algorithms. The desired pattern and the optimization result for the array composed of isotropic elements are shown in Fig. 11, where the fulfillment of all requirements is verified.

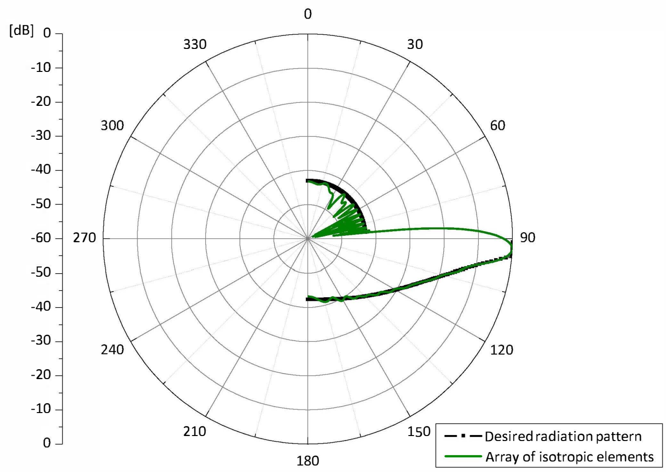

Fig. 11. Synthesized radiation pattern of the linear array with isotropic elements.

In order to demonstrate the influence of the array element radiation characteristic on the synthesized pattern, the isotropic elements were replaced by the *E*-shaped antenna described in Section II. The resulting radiation pattern, by exciting the array elements with the same coefficients as obtained with the isotropic elements, is shown in Fig. 12, where one can observe that the power level is strongly reduced as θ approaches 180°. The array pattern does not satisfy the specifications anymore.

A new optimization was carried out now considering the *E*-shaped antenna as array element. The same simulation setup was used for GA and SQP. The final result is shown in Fig. 13, where a much better representation of the desired pattern has been obtained.

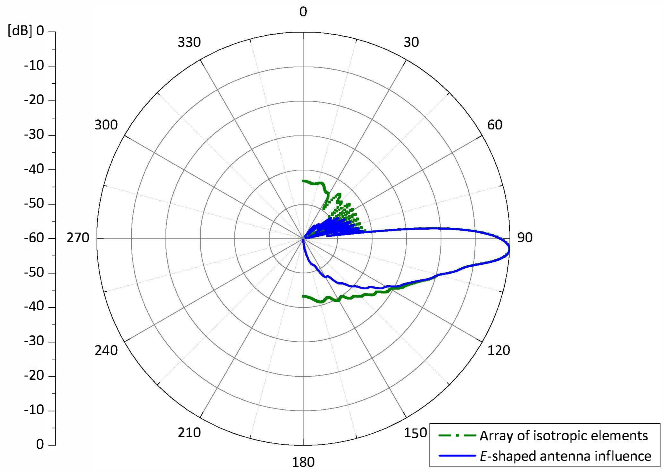

Fig. 12. Influence of the *E*-shaped antenna pattern on the synthesized array considering the coefficients determined with isotropic elements.

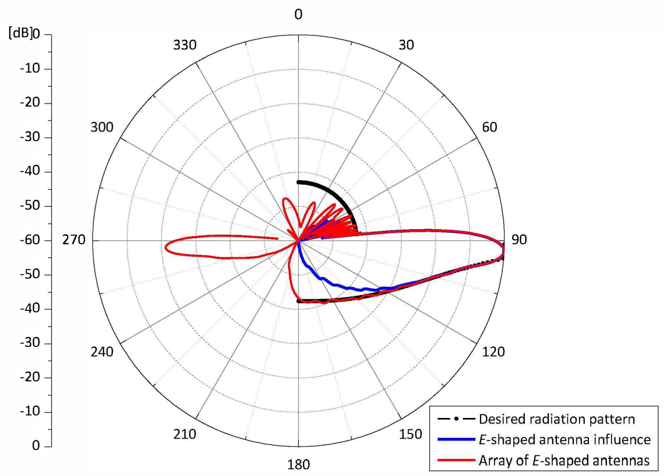

Fig. 13. Synthesized radiation patterns of the linear array with *E*-shaped antennas.

V. CONCLUSION

This paper presented the synthesis of squared cosecant-shaped pattern obtained with a linear antenna array. The main goal was to obtain uniform distribution of power along the coverage area of a radio base station and reduce the co-channel interference. In addition, a linear array with 24 elements was used to obtain squared cosecant-shaped radiation pattern with a ripple less than 1 dB. It should be pointed out that less elements can be used to synthesize a squared-cosecant shaped pattern, but with the drawback of introducing additional ripple in the radiation pattern.

The synthesized pattern exhibits low power levels in the angular region above the horizon, where the power radiated in the direction of co-channel cells was reduced by 3 dB. This contribution demonstrated that the radiation characteristic of the array element influences the final pattern and must be considered since the beginning of the optimization process.

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