ANTENNA ARRAYS WITH SYNTHESIZED FREQUENCY RESPONSE OF GAIN
ANTĚNNÍ ŘADY SE SYNTETIZOVANOU KMITOČTOVOU ZÁVISLOSTÍ ZISKU

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1 INTRODUCTION

Nowadays, the demands on communication technologies are increasing. The lack of frequency spectrum enforces us to move wireless technologies at higher frequencies. Therefore, the millimeter-wave frequency band plays an important role in modern communication systems. At higher frequencies, the wireless communication faces lower interferences, higher data rates, smaller sizes of antennas, etc.

At present, increasing demands are related to front ends. More and more attention is paid to multifunction devices (integrated modules for filtering and radiation performance, especially). These multifunction modules comprising filtering and radiation performance are called filtering antennas (filtennas).

In a communication system with a highly sensitive receiver, a band-pass filter has to be necessarily put in between the antenna and the first stage of the receiver since the band-pass filter can separate the required signal at the operating frequency from out-band signals. In order to make the design compact, an antenna and a band-pass filter can be integrated into a single module carrying out both the spatial pre-filtering and the spectral one. Hence, we require a properly designed filtering antenna again.

Demands on filtering antennas are not limited to spatial and spectral filtering only. Antennas are also required to exhibit a prescribed side-lobe level, impedance matching and polarization properties.

In the dissertation thesis, we present an overview of existing concepts of filtering antennas. Since a multi-objective synthesis of filtering dipole arrays has not been properly described in the open literature yet, we have deeply investigated potential multi-objective approaches to the synthesis of such arrays. In order to speed up the design, the synthesis is performed using idealized arrays for evaluating objective functions. In the following step, the design is refined in a full-wave solver. Finally, the designed antenna is fabricated and measured.
2 STATE-OF-THE-ART

Most filtering antennas are based on an integration of a frequency filter into the antenna structure. Many papers describe an integration of a band-pass filter into the feeding network of an antenna or an antenna array. Several papers deal with horn antennas with filtering behavior. A limited number of papers discuss the design of antenna arrays which can provide both the spectral filtering and the spatial one at the same time.

In the following sub-chapters, existing approaches are briefly analyzed.

2.1 Filtering horn antennas

Horn antennas can provide filtering functions, but small changes in the structure of the antenna have to be done. In order to create a filter in a horn, capacitive and inductive elements have to be created using discontinuities and metal obstacles. The obstacles can generate higher modes and specific modes of resonance. In case of H-plane horn antennas, the filtering function can be provided by an integrated band-pass filter. If discontinuities and metal obstacles cannot be used to create a filter, the role of a frequency filter can be played by a frequency selective surface (FSS) in the aperture of the horn antenna. A good radiation and filtering performance of a horn antenna can be also achieved by a proper design of a corrugated horn antenna. Such an antenna can reduce noise excited by feeds of a regular horn antenna [2] - [6].

2.2 Antennas with integrated filters

Several papers have described an integration of a filter into a planar antenna. A compact filtering patch antenna with a second-order quasi-elliptic response of gain belongs to most interesting approaches. Here, a U-shaped radiating patch is excited by a T-shaped resonator through an inset coupling structure. The inset coupling structure can be considered as an admittance inverter in the filter design. This filtering patch antenna was extended to an array where antenna elements together with a very compact feeding network behave as a third-order band-pass filter. The U-shaped patches play the role of both the radiator and the last stage of the filter [7], [8].

Some papers have been focused on antennas with high band-edge gain selectivity. Related structures have been created by a radiator and a band-pass filter on the same substrate. For example, the structure can be created by a printed meander-like antenna and an integrated quarter wavelength resonator. Here, the antenna performs not only functions of a radiator but also functions of the resonator of the band-pass filter [9]-[11].

A direct integration of the band-pass filter into the antenna feeding structure is the easiest way to create an antenna with filtering performance [12]-[15]. We can even create a reconfigurable filtenna by using a reconfigurable band-pass filter [16].

A slightly different approach of creating an antenna with filtering performance uses an Ultra Wide Band (UWB) antenna with implemented narrowband rejection elements for the reduction of interference. If interference is reduced properly, certain
narrowband services are suppressed appropriately [17], [18]. Radiation of the band-pass filter in the out-band is a disadvantage of this concept.

The level of gain can be reduced in the out-band by using a combination of antennas and band-pass Substrate Integrated Waveguide (SIW) filters [19] - [21].

2.3 Multilayer antennas and antenna arrays with filtering performance

Using multilayer technology, a good suppression of the filter radiation in the stop-band can be reached. An integration of a filter into a multilayer planar antenna was published in [21]-[23]. Here, filters were located on the back side of the antenna.

Low-profile arrays of printed dipoles with inherent filtering properties belong to multilayer antennas as well. Here, both the antenna and the band-pass filter are integrated into a small module, which is inserted into the unit cell of an array. The element of the array may be represented by a three-pole Chebychev filter with an implemented load resistance and the first resonator created by the radiating dipole completed by the backing reflector. A part of the filtering function is implemented via capacitively coupled dipoles. This capacitive connection provides an additional parameter for tuning the array [1].

Unfortunately, the described filtering antenna can excite common-mode resonances in E-plane scanning due to the unbalanced excitation of dipoles. Hence, the low-profile dipole array was improved by connecting an X-band hybrid-ring [24] - [26].

2.4 Synthesis of the antenna array

The synthesis problem of an antenna array is related to the calculation of the excitation and geometric configuration that can produce a desired pattern. Many methods have been adopted to achieve specified radiation patterns [27]-[30] for non-uniformly excited, non-uniformly spaced linear arrays.

2.5 Filtering antennas without filter

These filtering antennas contain no filters in the antenna structure. The filtering performance is obtained by a suitable setting of antenna elements individually.

A stacked patch antenna with the frequency response of reflection coefficient with sharp band edges, a good impedance matching in band of operation and high rejection in the out-band was developed. This concept inherently utilizes multiple narrow band elements (patches) with slightly displaced resonant frequencies. Patches are combined so that their frequency characteristics can be similar to a pass-band filter [31].

Another design of the filtering antenna exploits Yagi concept. This antenna exhibits an improved out-band gain suppression to reduce the demands on the front-end pass-band filter. Two parasitic elements in a close distance to the driven element are used to increase the Q-factor of that element. The length and position of other directors and reflectors are optimized to suppress the out-band gain [32].
3 DISSERTATION OBJECTIVES

The previous chapter showed us that most of existing concepts of filtering antennas are based on the integration of filters into the antenna structure. Naturally, the integration of filters increases dimensions of the antenna and makes the design and the fabrication of the antenna more complicated. Radiation of filters and the associated deformation of the radiation pattern of the antenna array are drawbacks of this concept.

The dissertation thesis is therefore focused on shaping the gain of the antenna by a pure synthesis of the antenna array (the excitation power is reflected back to the source by the detuned input impedance of the antenna rather than by a frequency filter).

The antenna array has to provide both the spatial filtering and the spectral filtering. The spatial filtering ensures us that the main lobe direction and the level of the realized gain exhibits negligible variations over the operating band. The spectral filtering ensures us that the value of the realized gain is maximal and constant over the operating band, and is minimal in the out-band of the antenna array.

Moreover, the synthesis is requested to consider additional objectives like impedance matching, side lobe level, main lobe direction, etc.

The main objectives of the dissertation thesis can be formulated as follow:

- Let us develop an approach to the synthesis of a dipole antenna array with the prescribed spectral filtering and spatial filtering. Moreover, the synthesis has to consider additional objectives like impedance matching, side lobe level, main lobe direction, etc.

In order to meet the main objective, partial goals have to be met:

- Let us optimize amplitudes, phases and distances of individual elements of an antenna array by considering an idealized antenna. The idealized array enables us to achieve required gain, impedance matching, side lobe level and main lobe direction in fast and efficient way.

- Let us refine the initial design based on an idealized array by a full-wave simulator.

- Let us experimentally verify the refined design of the selected dipole antenna array.

Since the objectives like the main lobe direction, the level of the realized gain, the side lobe level and the impedance matching are conflicting, a multi-objective optimization has to be used. Exploitation of multi-objective techniques for the described synthesis has not been published yet.
4 OPTIMIZATION OF THE DIPOLE ANTENNA ARRAY

In this chapter, we will focus on the single- and multi-objective optimization of the dipole array. Then, we will compare different ways of evaluating a frequency response of gain.

4.1 Single-objective optimization of the antenna array

In order to make an initial design of a dipole array with the required gain response as efficient and as quick as possible, an idealized antenna array can be considered for evaluating an objective function in a single-objective optimization. For simplicity, we used omnidirectional elements in free space to represent 10 dipoles in an array. The required gain response is synthesized by changing amplitudes and phases of currents exciting omnidirectional radiators; and by varying positions of these radiators. Values of amplitudes, phases and positions (state variables) are iteratively computed by a global optimizer (particle swarm optimization, PSO) to synthesize the required frequency response of the gain. Since the array is one-dimensional, we can optimize the gain in the H plane of dipoles only. In order to minimize CPU-time demands, the global optimizer uses an idealized antenna model with a neglected coupling among elements [33]-[35].

During the synthesis, we evaluate radiation patterns, gain and radiation resistance by a simplified computation implemented in MATLAB. For computing the radiation pattern, we created a script which evaluates radiation function of the antenna array for corresponding state variables. The radiation function of an antenna array is a product of radiation functions of the dipole [36]

\[ |F_d(\Psi)| = \frac{\cos(kl \cos \Psi) - \cos kl}{\sin \Psi}, \quad (4.1) \]

where \( \Psi \) is the angle between the position vector and dipole axis in which we calculate electric field intensity, \( l \) means the length of dipole and \( k \) is wave number.

the ground plane [36]

\[ |F_r(\Psi)| = 2 \cdot \sin(kh \cos \vartheta), \quad (4.2) \]

where \( h \) is the distance of the dipole from the ground plane and \( \vartheta \) is the angle between the position vector and dipole axis and the group radiation function [36]

\[ |F_g(\Psi)| = \sum I_i \cdot e^{(-jkd_i \cdot x \cos \Psi)} \], \quad (4.3) \]

where \( I_i \) is the amplitude of current, \( d \) is the distance between dipoles and \( x \) is the distance of the dipole from the reference point.

So, the final radiation function of the antenna array is [36]

\[ |F(\Psi)| = Fd \cdot Fr \cdot Fg. \quad (4.4) \]
In the next step, we have to create a script returning the integrand for evaluating the radiation resistance. Since the radiation of the antenna array is not rotationally symmetric, we have to fully integrate [36]

\[
R_\Sigma = \frac{30}{\pi} \int_0^{2\pi} \int_0^{\pi/2} |F|^2 \sin \theta d\theta d\phi,
\]

(4.5)

where \( \phi \) is the angle between the position vector and perpendicular axis to the axis dipole in which we calculate electric field intensity.

Finally, the gain can be calculated using [36]

\[
D_{(\phi, \theta)} = 120 |F_{(\phi, \theta)}|^2 / R_{\Sigma m}.
\]

(4.6)

Then, we can find the maximal value of gain in the normal direction to dipole axis related to the dipole array. Once the frequency response of gain is known, we are able to evaluate the objective function, i.e. the squared difference in between the required frequency response of gain and the computed one.

Fig. 4.1 shows the results of the synthesis of the dipole array. We requested the optimizer to suppress the out-band gain for more than 10 dB with respect to the in-band gain in the main lobe direction in the H plane of dipoles. Values of optimal parameters of the idealized dipole array are given in table 4.1.

The radiation pattern in the H-plane shows us that the deflection of the main lobe is smaller than 6.5°. The beam scan in the pass-band of the antenna is smaller than 1°.

**Table 4.1.** Optimal parameters of the idealized antenna array [33].

<table>
<thead>
<tr>
<th>Antenna element</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>8</th>
<th>9</th>
<th>10</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Norm. amplitude A</strong></td>
<td>0.32</td>
<td>0.50</td>
<td>1.00</td>
<td>0.18</td>
<td>0.36</td>
<td>0.16</td>
<td>0.16</td>
<td>0.80</td>
<td>0.41</td>
<td>0.01</td>
</tr>
<tr>
<td><strong>Phase ( \varphi ) [deg]</strong></td>
<td>201</td>
<td>135</td>
<td>179</td>
<td>18.3</td>
<td>204</td>
<td>176</td>
<td>187</td>
<td>330</td>
<td>149</td>
<td>117</td>
</tr>
<tr>
<td><strong>Distance ( d ) [\lambda]</strong></td>
<td>0.00</td>
<td>0.67</td>
<td>0.90</td>
<td>0.36</td>
<td>0.66</td>
<td>0.70</td>
<td>0.52</td>
<td>0.68</td>
<td>0.77</td>
<td>0.34</td>
</tr>
</tbody>
</table>

**Fig. 4.1:** Synthesized frequency response of gain (left) and radiation patterns of the synthesized array in H plane (right) [33].
4.2 Multi-objective optimization of the antenna array

The optimization described in the previous paragraph dealt with a single objective only. We looked for a required frequency response of the gain. If multiple objectives have to be considered during the design, a multi-objective optimization has to be performed. In the described research, we concentrated on the exploitation of a Multi-Objective Self-Organizing Migrating Algorithm (MOSOMA) [37].

In order to obtain frequency responses of investigated quantities describing the dipole array, we optimize amplitudes and phases of excitation currents, and lengths of the individual dipoles. The distance between dipoles is kept uniform being $\lambda/4$. We start the synthesis with an antenna array consisting of 8 separately excited dipoles.

Using MOSOMA, we are able to compute the reflection coefficient at the input port of the antenna, the realized gain, the main lobe direction and the side lobe level for the array to be designed. Computations of antenna parameters are performed in 4NEC (full-wave method of moments) and exported to MATLAB. In MATLAB, objective functions are evaluated and iteration steps of the multi-objective optimization are computed. For simplicity, the flowchart of the optimization routine is depicted in Fig. 4.2.

![Flowchart of the optimization routine](image)

**Fig. 4.2:** The flowchart of the optimization routine.

In order to analyze the array in 4NEC, we have to generate an input file containing input parameters for the 4NEC simulation. The input file *.nec comprises the following parameter:

- **GW** settings of the geometry of individual wires (arms of dipoles);
- **GE** settings of the geometry boundaries (using the ground plane);
- **GN** type of the ground plane;
- **EX** type, amplitude and phase of excitation sources;
- **FR** frequency range and frequency step;
- **RP** radiation pattern render settings.

The automatic generation of the 4NEC input file is a part of the evaluation of the objective function, obviously.

Next, the dipole array is simulated in 4NEC. After finishing the simulation, 4NEC creates an output file *.out which contains results. The resulting data describes
excitation sources, input impedance and admittance, radiation patterns etc. The input impedance of each dipole includes mutual impedances related to coupling to other dipoles.

Then, we can evaluate by the loaded 4NEC simulation results the reflection coefficient and compute the main lobe direction. The gain can be re-computed to the realized gain. Finally, objective functions can be evaluated.

For the multi-objective optimization, we proposed two possible approaches to get a required frequency response of the realized gain. In both the approaches, the calculation of the input impedance is related to a feeding point which is identical with the first dipole. We assume, that dipole elements of an antenna array are connected by transmission lines with the characteristic impedance $Z_0 = 75 \, \Omega$. The distance between the dipoles is $\lambda/4$, so the transmission line transforms the impedance of the dipoles to the input impedance of the dipole antenna array by the following equation [38]:

$$Z_{inp} = \frac{Z_c}{Z_l}$$  \hspace{1cm} (4.7)

where $Z_c$ is the characteristic impedance of the transmission line and $Z_l$ is the impedance of the load.

The reflection coefficient $\rho_l$ is given by the input impedance of the antenna array and the characteristic impedance of the transmission line [38]:

$$\rho_l = \frac{Z_{inp} - Z_c}{Z_{inp} + Z_c}$$  \hspace{1cm} (4.8)

The developed approaches differ in the evaluation of the frequency response of the maximal gain. The first approach uses data from the 4NEC output file for the gain evaluation. From the 4NEC simulation, we can directly obtain values of the gain $G_{dir}$ in the H plane. The gain is transformed to the realized gain by [38]

$$G_{real} = 10 \log \left( (1 - |S11|^2) \cdot G_{dir} \right).$$  \hspace{1cm} (4.9)

where $S11$ means the reflection coefficient at the input.

In the final step, we can find the maximal gain and its corresponding main lobe direction.

The second approach is based on the evaluation of the realized gain by equations given in chapter 4.1. And the obtained value of the gain is then transformed to the realized gain by the equation 4.9.

We considered two objective functions when performing the optimization routine. The first one refers to the reflection coefficient and the second one is related to the realized gain. The main lobe direction is considered in the optimization routine as a constraint. In contrast with the single-objective optimization, we do not require the direction of the main lobe being perpendicular to the plane of the antenna array. The main lobe direction is demanded staying constant in the specific direction over the whole operating band of the antenna array. From the viewpoint of the objective, the main lobe has to deflect minimally from the main lobe direction. In our case, we are willing to accept $\pm 5^\circ$ deflection.
The frequency range of the optimization of the antenna array is set from 2.80 GHz to 3.20 GHz and is divided into twenty steps. The pass-band covers frequency steps from 8 (2.96 GHz) to 13 (3.06 GHz). Then, the stop-bands obviously cover the intervals from steps 1 (2.80 GHz) to 7 (2.94 GHz) and 14 (3.08 GHz) to 20 (3.20 GHz). In the pass-band, magnitude of the reflection coefficient is required being smaller than threshold levels in Table 4.3. In the stop-band, magnitudes of the reflection coefficient are requested being larger. In the stop-band, the realized gain is asked to be suppressed below the threshold levels listed in Table 4.3, and vice versa.

**Table 4.3.** Threshold levels in the fitness function

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Threshold</th>
</tr>
</thead>
<tbody>
<tr>
<td>G_pass = 13;</td>
<td>Gain threshold</td>
</tr>
<tr>
<td>G_stop = -3;</td>
<td>in pass-band</td>
</tr>
<tr>
<td>S11p = -12;</td>
<td>%S11 threshold</td>
</tr>
<tr>
<td>S11s = -5;</td>
<td>in pass-band</td>
</tr>
<tr>
<td>Ang_ref = 180;</td>
<td>Main lobe</td>
</tr>
<tr>
<td></td>
<td>direction reference</td>
</tr>
</tbody>
</table>

Evaluated objective functions related to the frequency responses of the reflection coefficient and the realized gain are depicted in Fig. 4.3. Here, the best 20 results (red color) from 10 000 iteration steps are located near to the origin of the coordinate system and form the so-called Pareto front. From the objective space is obviously that we met the objective related to the reflection coefficient. The objective related to the realized gain was evidently too strict.

![Objective space](image)

**Fig. 4.3:** Objective space: realized gain versus impedance matching. Red dots represent Pareto front of optimal solutions.

We selected all 20 best results of the multi-objective optimization and we objectively determined the best result. For this result, we obtained amplitudes and phases of currents exciting dipole elements, and lengths of dipole elements. Then, we evaluated frequency responses of the reflection coefficient, the realized gain and the main lobe direction applying the first approach (objectives evaluated using 4NEC output data; Fig. 4.4) and the second approach (objectives evaluated analytically considering an idealized array; Fig. 4.5).
Frequency response of the reflection coefficient shows a small phase shift of the resonance frequency. The main lobe direction stays constant over the operating frequency band. If a ground plane is not considered in simulations, the radiation pattern is symmetric according to the horizontal plane. So the main lobe direction can acquire ±180° (Fig. 4.5).

![Image](image1.png)

**Fig. 4.4:** Results of the first computational approach.

![Image](image2.png)

**Fig. 4.5:** Results of the second computational approach.

Using the first approach for the evaluation of the realized gain (the full-wave numerical 4NEC model), the frequency response of the realized gain corresponds to the frequency response of the inverse value of the reflection coefficient. The power of the source enters the antenna and is radiated to the environment.

Using the second approach (the idealized analytical model), the frequency response of the realized gain directly corresponds to the frequency response of the reflection coefficient. Whereas the numerical model of the antenna considers a mutual coupling of antenna elements, the idealized analytical model neglects the coupling.

The comparison of investigated approaches shows that the antenna array has to be optimized with both the approaches separately, and obtained results have to be confronted (Fig. 4.4 and Fig. 4.5). CPU-time demands of the second approach (the idealized analytical model) are approximately eight times a higher compared to the first approach (the numerical 4NEC model).

The results presented in this chapter were obtained after the first run of the multi-objective optimization with specific settings of parameters. In the following, the synthesis is refined to reach better selective performance of the antenna array.
5 SERIAL FEED VERSUS PARALLEL FEED OF THE ANTENNA ARRAY

In this chapter, we will study two different methods of feeding the dipole array:

1. **Parallel feed.** An antenna array consists of dipoles completed by the ground plane. Each dipole is fed separately (an amplitude and a phase of the excitation current of each dipole is set independently).

2. **Serial feed.** An antenna array consists of dipoles above the ground plane, which are connected by simplified transmission lines.

Both the described structures, which are simulated in the full-wave moment solver 4NEC, are completed by the reflector to limit the radiation into a single half-space and increase the gain of the array.

We start the synthesis with an array consisting of 8 dipoles. Then, we increase the number of dipoles to 16 elements and 32 elements. We expect that the rate of the gain selectivity is determined by the number of radiated elements in the array.

The frequency range of optimization stays the same as listed in chapter 4.2. Goals of the optimization routine for both approaches are summarized in Table 5.1. Thresholds are marked *pass* in the pass-band and *stop* in the stop-band. Further, we demand the main lobe direction deflection ±5° from the reference angle over the whole operating band. The preliminary results of the synthesis of the dipole array were published in [39].

<table>
<thead>
<tr>
<th>Parameter</th>
<th>$G_{\text{pass}}$</th>
<th>$G_{\text{stop}}$</th>
<th>$S11_{\text{pass}}$</th>
<th>$S11_{\text{stop}}$</th>
<th>$SLL_{\text{pass}}$</th>
<th>$SLL_{\text{stop}}$</th>
<th>$\text{Ang}_{\text{ref}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>With TL</td>
<td>&gt; 13 dB</td>
<td>&lt; 5 dB</td>
<td>&lt; -12 dB</td>
<td>&gt; -5 dB</td>
<td>&gt; 8 dB</td>
<td>&lt; 2 dB</td>
<td>±5°</td>
</tr>
<tr>
<td>Without TL</td>
<td>&gt; 16 dB</td>
<td>&lt; 8 dB</td>
<td>-</td>
<td>-</td>
<td>&gt; 8 dB</td>
<td>&lt; 2 dB</td>
<td>±5°</td>
</tr>
</tbody>
</table>

5.1 Synthesis of antenna array with parallel feeding

In the case of the dipole array with parallel feeding, we do not include the transmission line into the 4NEC simulation. The feeding network of an antenna array will be designed separately. So we can optimize only a required frequency response of the gain and the side lobe level. In order to achieve required properties of the antenna array, we change amplitudes and phases of excitation currents and lengths of radiating elements. The distance between each element is uniform and is set at $\lambda/4$. The distance of the dipole array above the ground plane is uniform and is assumed to be $\lambda/4$.

The optimization routine includes also the condition requiring the stability of the main lobe direction (the direction of the main lobe is required to stay constant over the operating bandwidth). Respectively, the main lobe deflection is asked to be minimal.

In case of the dipole array with parallel feeding without including the transmission line, we compare the gain for various numbers of radiating elements only (see Fig. 5.1).
Here, the tendency of the increasing gain and improving out-band rejection depending on the number of dipoles is clear.

Fig. 5.1: Frequency response of gain (a), main lobe direction (b) and side lobe level (c) of antenna array with parallel feeding.

The comparison of the main lobe direction of the dipole antenna array is depicted in Fig. 5.1. In case of 8 dipoles, there is a slight deflection of the main lobe direction in the operating band. In the remaining cases, the direction of the main lobe stays constant through the entire operating band. Better side lobe levels in the pass-band for antenna arrays with more dipoles can be seen. In order of the 8 dipoles antenna array, the side lobe level achieves best values at lower frequencies.

In the last configuration of the dipole array, we used only 30 dipoles. The number of dipoles had to be reduced from 32 to 30 due to the limited number of feeding ports in the 4NEC program.

5.2 Synthesis of antenna array with serial feeding

If we include a simplified transmission line into the synthesis of the antenna array, we are able to obtain a required frequency response of the realized gain and the reflection coefficient. The optimization of the antenna array is then performed by changing the amplitude and phase of the excitation current, lengths of radiating elements and distances between each element. The distance of the dipole antenna array from the ground plane is the same as in the previous chapter and is assumed to be $\lambda/4$. 
The synthesis also includes the optimization of the side lobe level. The condition requiring the stability of the main lobe direction is again added into the optimization routine.

![Graphs showing frequency response](image)

**Fig. 5.2:** Frequency response of reflection coefficient (a), realized gain (b), main lobe direction (c) and side lobe level (d) of antenna array with serial feeding.

The comparison of impedance matching of the dipole array with included transmission lines is shown in Fig. 5.2. Obviously, the impedance matching achieves good values around the resonant frequency. But impedance matching −12 dB in the operating band has not been satisfied in most configurations. This goal was satisfied for the configuration with 16 dipoles only after 9 optimization runs.

Frequency responses of the realized gain show that the value of the gain depends on the number of dipoles clearly. The increase of selectivity of the antenna array with an increasing number of dipoles is also evident. Fig. 5.2 demonstrates that the increasing number of radiated elements increases the stability of the main lobe direction and side lobe levels in the operating band.

For most configurations of dipole arrays, the gain evidently increases with the increasing number of optimization runs. An improvement of the impedance matching in relation to the number of optimization runs is also clear.
6 REALISTIC DIPOLE ARRAYS IN FULL-WAVE SIMULATOR

In this chapter, we redesign the dipole array with serial feeding to obtain a planar structure. The planar array is simulated in CST Microwave Studio and compared with results of the synthesis described in the previous chapter.

6.1 Comparison of dipole arrays with parallel feeding

We imported the dipole array with a parallel feeding into the CST Microwave Studio to validate results of the synthesis performed in MATLAB and simulated in 4NEC. In CST, we created numerical models of all configurations of dipole arrays. Dimensions of the dipole array are identical with optimal results given in chapter 5.1.

All discrete ports used in CST simulations were referenced to the characteristic impedance 73 Ω. Boundary conditions of the bounding box were changed so that the electric potential $E_r = 0$ is assigned one side of the bounding box. The respective side is situated in parallel to the dipoles in the array. That way, an infinite ground plane representing similar conditions in 4NEC was created. The dipole array was situated in the distance $\lambda/4$ from the infinite ground plane. The other sides of the bounding box were set to “open add space”.

![Graphs of frequency responses](image)

**Fig. 6.1:** Frequency responses of gain (a), main lobe direction (b) and side lobe level (c) of array with parallel feeding consisting of 30 dipoles.
The comparison of dipole arrays with the parallel feeding simulated in CST and 4NEC is shown in Figure 6.1 for 8 dipoles configuration. The gain is obviously decreased for units of decibels in CST simulations (a differential solver) compared to 4NEC ones (an integral solver). The tendencies of main lobe direction and side lobe level achieved in CST follow values obtained from 4NEC.

The comparison of synthesis results of the dipole array with parallel feeding performed by CST and 4NEC is listed in Table 6.1. Here, maximal values of the directivity gain and the side lobe level are presented. Values of the main lobe direction are given in range over the entire pass-band.

<table>
<thead>
<tr>
<th>Solver</th>
<th>CST</th>
<th>4NEC</th>
</tr>
</thead>
<tbody>
<tr>
<td>8</td>
<td>11.2</td>
<td>0.15</td>
</tr>
</tbody>
</table>

### 6.2 Comparison of dipole arrays with serial feeding

In order to manufacture the synthesized dipole array, we decided to design a planar version of the dipole array with serial feeding. For the planar design, we used a thin substrate CuClad 217 with a low value of relative permittivity ($\varepsilon_r = 2.17$, $h = 0.508$ mm, $\tan \delta = 0.0009$).

In the first step, we designed a transformer [microstrip transmission line] $\rightarrow$ [balanced stripline] (balun). We chose an ultra-wideband balun with a compact size (a low height $y_f$, especially). The balun is depicted in Figure 6.2. Due to the extension of the ground plane by $y_g$, we are able to improve the impedance matching of the antenna array. Dimensional specifications of the balun are listed in Table 6.2. [40].

![Fig. 6.2: Ultra wideband balun [40].](image)

The characteristic impedance of the microstrip line at the input of the balun is 50.0 $\Omega$. The characteristic impedance of the balanced stripline at the opposite side of the balun is then 60.6 $\Omega$. The real part of the input impedance of the dipole antenna in the resonance is about 73.0 $\Omega$. So, we need to insert an impedance transformer between the balun and the dipole array. In our case, a $\lambda/4$ impedance transformer was used and the characteristic impedance of the transformer was set to 66.8 $\Omega$. 
Table 6.2. Dimensions of the balun.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>( w ) [mm]</th>
<th>( r_1 ) [mm]</th>
<th>( y_x ) [mm]</th>
<th>( y_y ) [mm]</th>
<th>( w_y ) [mm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Value</td>
<td>146.00</td>
<td>72.25</td>
<td>2.38</td>
<td>47.85</td>
<td>1.50</td>
</tr>
</tbody>
</table>

The balun with the impedance transformer achieves a sufficient performance (see Fig. 6.3). The reflection coefficient is below \(-10\) dB in the whole frequency band of operation. The transmission coefficient approaches 0 dB over the entire frequency range.

![Graph of S-parameters](image)

Fig. 6.3: S-parameters of the balun in end-to-end connection.

Having an appropriately designed balun, we can finish the design of the planar dipole array with serial feeding (Fig. 6.4). For the planar dipole array, dimensions listed in chapter 5.1 were used.

![Top view of the planar array of 8 dipoles with serial feeding](image)

Fig. 6.4: Top view of the planar array of 8 dipoles with serial feeding.

Lengths of dipoles and distances between them stay unchanged. Widths of the dipoles agree with the dipole diameters used in chapter 5.1. The left-hand arm of each individual dipole is placed on the opposite side of the CuClad 217 substrate than the right-hand one.

We placed a ground plane in the distance \( \lambda/4 \) from the dipole array. Boundary conditions were set the same way as in the previous chapter. So, we again get an infinite ground plane under the antenna array.

Figures 6.5 shows simulation results of the 8 dipoles array with serial feeding. Simulations were performed in 4NEC (wire structure optimized in MATLAB) and CST (planar structure). We compare frequency responses of reflection coefficient, realized gain, main lobe direction and side lobe level. Results show a disagreement between 4NEC models (wire structures) and CST ones (planar structures). This disagreement is caused by the transformation of the wire dipole array to the planar one.
Fig. 6.5: Frequency responses of reflection coefficient (a), realized gain (b), main lobe direction (c) and side lobe level (d) of array with serial feeding consisting of 8 dipoles.

The planar dipole arrays with serial feeding were further optimized by particle swarm optimization implemented in CST. Requested frequency responses of reflection coefficient and realized gain were objectives of this single-objective optimization (objectives were weighted and summed). The input parameters were swept ±15% from the optimal parameters obtained in chapter 5.2.

Figure 6.5 shows the frequency response of the reflection coefficient of the dipole array. Deterioration of the reflection coefficient after transforming the wire structure (MATLAB) to the planar structure (CST) is clear. But the bandwidth was increased. After the additional optimization of the planar dipole array (CST opt.), we got better results compared to MATLAB optimizations.

By the optimization of the planar dipole array in CST, we got better realized gain (compared to the wire array optimized in MATLAB). A slight deflection of the main lobe direction in the operating band after transforming the wired structure to the planar structure is clear. A good agreement between MATLAB results of the side lobe level and CST ones was achieved.

This chapter shows that antennas resulting from a very efficient multi-objective synthesis of an array of wire dipoles can be simply redesigned to an array of planar dipoles. A simple single-objective tuning of the planar design can ensure us that parameters of original antennas can be preserved.
7 EXPERIMENTAL VERIFICATIONS

The planar array of 8 dipoles with the serial feeding was fabricated in the prototyping laboratory of the Department of Radio Electronics, Brno University of Technology. We fabricated the dipole array from the substrate CuClad 217 ($\varepsilon_r = 2.17$, $h = 0.508$ mm, $\tan \delta = 0.0009$). The planar structure of the antenna was equipped with an SMA connector. We completed the dipole array by the ground plane of finite dimensions $160$ mm $\times$ $260$ mm from the tinplate. The distance between the ground plane and the dipole array was $\lambda/4$. As standoffs, we used polystyrene bricks (see Fig. 7.1).

![Manufactured planar array of 8 dipoles with serial feeding and the ground plane: the fabricated prototype.](image)

Fig. 7.1: Manufactured planar array of 8 dipoles with serial feeding and the ground plane: the fabricated prototype.

Fig. 7.2a shows the comparison of the simulated reflection coefficient and the measured one at the input of the dipole array. Obviously, resonance frequency of the measured antenna is slightly lower compared to the simulated one. Correspondingly, the maximum value of the side lobe level and main lobe direction are also shifted to lower frequencies. The frequency shift is dominantly caused by parasitics which were not considered in the numerical model of the planar array.

The fabricated dipole array achieved a worse value of the reflection coefficient in minimum. On the other hand, the bandwidth of the array was slightly increased. Nevertheless, both the simulated dipole array and the measured one covered the entire pass-band.

The comparison of the simulated and measured realized gain is depicted in Fig. 7.2b. Clearly, the measured realized gain is higher that the simulated one.

When depicting Fig. 7.2b, we measured the realized gain at ten frequencies in the range from $2.8$ GHz to $3.2$ GHz. Then, we interpolated values of the measured gain by a polynomial (the blue one). The measured dipole array exhibited a better selectivity, but in a wider frequency range.

The variations of the main lobe direction of the measured antenna in the pass-band were smaller compared to the simulated antenna (see Fig. 7.2c). Except of the slight frequency shift, frequency responses of the side lobe level are very similar (see Fig. 7.2d).
We can conclude that experiments approved the functionality of the developed method of the synthesis of dipole arrays.
8 CONCLUSION

The dissertation thesis was focused on synthesizing the frequency response of gain of an antenna array. Attention was turned to the synthesis of a conventional dipole array with the different number of radiating elements. The developed synthesis procedure was based on reflecting the excitation power by the detuned input impedance of the antenna rather than by implementing a frequency filter at the antenna input.

First, we demonstrated functionality of a single-objective optimizer in order to shape the frequency response of gain of the dipole array. We synthesized an idealized dipole array with the neglected coupling between elements. The synthesis was performed by MATLAB, but CPU-time demands of the procedure were high.

In order to enrich objectives by the frequency response of the reflection coefficient, the side lobe level and the main lobe direction, we had to convert the single-objective optimization to a multi-objective one. For the multi-objective synthesis, we used a global stochastic algorithm called MOSOMA.

Finally, we compared properties of dipole arrays with parallel feeding (each dipole was fed separately) and serial feeding (neighboring dipoles were connected by a simplified transmission line). All configurations of dipole arrays were completed by the ground plane to limit the radiation to a single half-space and increase the gain.

The obtained results showed us that the selective behavior of the dipole array with filtering performance strongly depends on the number of radiating elements.

In order to verify obtained results, the synthesized dipole arrays were modeled in CST (a differential time-domain simulator). For the dipole array with parallel feeding, a good agreement was achieved. For the dipole array with serial feeding, the wire structure was transformed to the planar one, and the array was completed by a balun. In order to improve the performance, we optimized planar arrays by the particle swarm optimization in CST. Then, we obtained better results compared to dipole arrays.

We selected a planar array of 8 dipoles with serial feeding to be fabricated and measured. The dipole array was completed by a reflector with finite dimensions. The reflector slightly improved performance of the dipole array.

Comparison of the simulated and measured planar array showed a slight shift of characteristics to lower frequencies. The shift could be caused by parasitics which were not considered in the numerical model. Nevertheless, we achieved a good agreement between simulated and measured results.

The research described in the thesis was limited to the synthesis of gain in the H plane of an array consisting of $1 \times n$ dipoles. When extending the research to the synthesis of antenna arrays consisting of $n \times n$ radiating elements, the frequency response of gain can be synthesized both in the H plane and the E plane.

The design and the optimization of a slot antenna array as a complementary structure to the dipole array can be considered as a potential extension of the described research. The slot antenna array might be fed by a substrate integrated waveguide (SIW) to suppress a parasitic radiation of the feeding network. That way, parameters of the slot array can be improved. The described structure can be used to verify the methodology of the synthesis of antenna arrays with the prescribed frequency response of gain.
REFERENCES


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ABSTRACT

In the thesis, we present a method of the synthesis of a dipole antenna array with prescribed spectral and spatial filtering capabilities. Thanks to the spatial filtering capabilities, the main lobe direction and the value of gain vary negligibly over the operating band. Thanks to the spectral filtering capabilities, the value of gain is maximal in the operating band and minimal out of the operating band. In order to synthesize a dipole array with prescribed filtering capabilities, amplitudes, phases and dimensions of antenna elements are optimized. The initial optimization is speeded up by considering an idealized antenna array when evaluating objective functions. Since the optimization comprises requirements on the main lobe direction, the value of gain and impedance matching, a multi-objective optimization is used. The optimized antenna array is analyzed by a full-wave simulator to verify results of the synthesis. Finally, the synthesized dipole array is manufactured and its performance is experimentally verified.