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CHARACTERISTICS OF DIFFERENT FOCUSING ANTENNAS IN THE NEAR FIELD REGION

By

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> A Dissertation Submitted to the Faculty of the University of Mississippi in Partial Fulfillment of the Requirements for the Degree of Doctor of Philosophy with a Major in Engineering Science in the School of Engineering

> > THE UNIVERSITY OF MISSISSIPPI

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Abstract

Focusing antennas are of interest in many application including microwave wireless power transmission, remote (non-contact) sensing, and medical applications. Different kinds of antennas such as array antennas, reflector antennas and Fresnel zone plate (FZP) antennas have been used for these applications.

Here, first, a new scheme in designing focused array antennas with desired sidelobe levels (SLLs) in the near field region is presented. The performance of the large focused array antennas is predicted based on the knowledge of the mutual admittances of a smaller array. The effects of various focal distances on the near field pattern of these antennas are investigated. Then, electric field pattern characteristics of the focused Fresnel zone plate lens antennas in the near-field region are presented. The FZP antenna fed by a circular horn is implemented and the effects of various focal lengths on the near field pattern of this antenna are examined. It is shown that the maximum field intensity occurs closer to the antenna aperture than to the focal point and this displacement increases as the focal point moves away from the antenna aperture.

The focusing properties of ultra-wideband (UWB) array antennas are also presented. Large current radiator (LCR) antennas are modeled by replacing the antenna with a set of infinitesimal dipoles producing the same near field of the antenna. LCR antenna arrays are used to provide high concentration of microwave power into a small region. It is shown that the defocusing effect occurs in pulse radiating antennas as well. Invasive weed optimization (IWO), a new optimization algorithm, is also employed to optimize the pulsed array antenna. In the attempt of optimizing the focused arrays, a new scenario for designing thinned array antennas using this optimization method is introduced. It is shown that by using this method, the number of elements in the array can be optimized, which yields a more efficient pattern with less number of elements. By applying this new optimization method to UWB arrays, the peak power delivered to a localized region can be increased.

DEDICATION

This work is dedicated to my Dad, Dr. Reza Karimkashi and my Mom, Ms. Zahra Monsef, for their love and support.

Abbreviations and Acronyms

ABC	Absorbing Boundary Condition
APM	Array Pattern Multiplication
APM/MC	Array Pattern Multiplication including the mutual Coupling
BC	Boundary Condition
DBC	Damping Boundary Condition
FZP	Fresnel Zone Plate
GA	Genetic Algorithm
HPBW	Half Power Beam Width
GO	Geometrical Optics
IBC	Invisible Boundary Condition
IDM	Infinitesimal Dipole Model
IWO	Invasive Weed Optimization
LCR	Large Current Radiator
LHCP	Left-Handed Circular Polarization
MoM	Method of Moment
PSO	Particle Swarm Optimization
RBC	Reflecting Boundary Condition
SD	Standard Deviation
SLL	Side Lobe Levels
TD	Time-Domain
UWB	Ultra-Wideband

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Chapter 1 INTRODUCTION

Antennas are commonly designed and characterized based on their far-field radiation patterns in order to deliver the signal energy to a large distance from the antenna aperture. However, in some applications we need to focus the microwave power at a point close to the antenna aperture. These kinds of antennas, called focused antennas, are of interest in non-contact (remote) sensing, medical applications, and microwave wireless power transmissions.

One of the foremost applications of focused antennas is in non-contact microwave sensing, where we need a focused beam for a precise sensing. In this application, the energy of the focused antenna should be mostly confined on the region we want to sense [1]-[2]. Another application of focused antennas is to achieve microwave-induced hyperthermia for medical applications. It is desired to maximize the power deposition in the near-field of the antenna to heat the cancerous tissue without the heating of healthy tissues adjacent to the tumor [3]-[5].

Another motivation for the development of this kind of antenna comes from the concept of wireless power transmission. This concept has been proposed to transmit continuous power without using transmission lines [6]-[8]. This application can include providing power from earth to space or from space to space to supply the power of orbiting satellites from a power station on earth or an orbiting power satellite, respectively [9]. In addition, it can be used to provide the power from solar power satellite systems in space to terrestrial markets [10]. For this purpose we need a large focusing antenna to transmit the power to a focused point or a desired near field region.

The earliest work on focused antennas was that of Wehner, reported in a 1949 Rand report [11]. Subsequently, Cheng [12], [13] calculated defocus regions, while Bickmore [14] calculated depth of field and measured far-field pattern in the near field of a focused linear slot array antenna. In 1962, Sherman published several calculated patterns of focused antennas [15]. Wheeler [16] discussed the effects of focusing on different patterns. Partially coherent excitation was treated by D'Auria and Solimini [17]. Measurements of focused lenses were made by Bachynski and Bekefi [18] and of focused linear arrays by Fahey *et al.* [19]. Synthesis of the axial field pattern in the radiation near field region was investigated by Graham [20]. The effects of the focused aperture amplitude tapering on axial lobes (forelobes and aftlobes) were investigated by Hansen [21].

Pendry in 2000 [22] showed that a planar slab of negative refractive index material can manipulate the near field in such a way that it achieves perfect imaging, i.e., a perfect reconstruction of the source's near field. He also showed that the near field could be focused with a negative permittivity slab. The experimental verification of negative refraction [23] and sub-wavelength focusing using negative refractive index [24], negative permittivity [25], [26], and negative permeability slabs [27] have demonstrated that near-field lenses are in fact a reality.

Array antennas have been recently considered as focused antennas for different applications [1], [2], and [28]. Conventional and shaped reflector antennas are also used as focusing antennas [7], [29]-[30]. Other alternative focused antennas are the dielectric lens

antennas as well as Fresnel zone plate (FZP) lens antennas consisting of a set of alternative open and opaque annuli arranged on flat or curvilinear surfaces [31]-[35].

Providing a high concentration of electromagnetic energy into small regions can also be done by using an array of pulsed antennas. Precise control of the time delays of the pulses radiated from the individual elements (ultra-wide band antennas) allows the concentration of energy within a region [3], [36]-[41].

Chapter 2 presents the array antenna design procedure to focus the microwave power in the radiation near-field region of the antenna. In this part, a small focused array antenna is implemented using microstrip patch elements to achieve the desired sidelobe levels in the Fresnel region based on Dolph-Chebyshev design. Then, larger arrays are designed by using the knowledge of the mutual admittances between the elements of the smaller array. Chapter 3 presents the focusing properties of Fresnel zone plate (FZP) lens antennas in the near-field region. In Chapter 4, the radiated field properties of a large current radiator (LCR), an ultra-wideband antenna, are investigated. Chapter 5 presents a new optimization algorithm inspired from colonizing weeds for electromagnetics applications. In chapter 6, this optimization method is applied to arrays of LCR pulse antennas in order to increase the peak power delivered to a localized region.

Chapter 2 FOCUSED MICROSTRIP ARRAY ANTENNA

2.1 Introduction

One of the important problems encountered in focused antennas is the need to design these antennas based on the desired properties of their near field radiation patterns. In other words, we need a design method to achieve specific sidelobe levels (SLL), half power beam width (HPBW), depth of focus (the distance between the axial -3dB points about the maximum intensity plane) and so on in the near field and Fresnel regions not only to achieve higher efficiencies and desired near field patterns for specific applications but also to avoid side effects of the near field power.

Among the most serious side effects are high SLLs decreasing the accuracy of the measurements in the non-contact sensing or heating healthy tissues adjacent to the tumor in medical applications. Moreover, in power transmission applications, higher SLLs may cause lower efficiency of the system and also interference to the adjacent communication systems or satellite receiving antennas.

Some attempts have been made to characterize the properties of the focused apertures in the radiation near field region. It was proven theoretically that the desired axial and transversal field patterns can be achieved from the proper aperture phase and amplitude distributions [15], [20]. Moreover, the effects of the focused aperture amplitude tapering on axial lobes (forelobes and aftlobes) were investigated [21].

The main purpose of this chapter is to present a design method for large array microstrip patch antennas to achieve desired SLLs in the radiation near field pattern based on a Dolph-Chebyshev array design. The new method is to include the mutual coupling between array elements in order to have more realistic results from the design of large focusing array antennas. Here, we first describe the design procedure of the focused array antennas. Then, we present simulation and measured data for a 4×4 focused array microstrip patch antenna to verify the efficiency of this design method. In the next section, larger focused array antennas are designed by using the predicted mutual admittances between elements in a smaller array of the same lattice and the same element type and size. A verification of the proposed method is shown by comparing the results from the present method to those obtained from full wave analysis of the array. Finally, the effects of various focal distances on both axial and transverse near field patterns of larger array antennas are investigated.

2.2 Design Procedure

It has been proven theoretically that in the focal plane near the axis of a focused aperture, the electric field will have all the properties of the far field radiation pattern if a quadratic phase taper is adjusted on the aperture of the antenna [15], [20]. Therefore, in order to achieve a spot beam with a desired shape at a focal distance, F, from the antenna aperture, a quadratic phase distribution of

$$U(x, y) = \exp\left(j\frac{k}{2F}(x^{2} + y^{2})\right)$$
(2.1)

and an adequate amplitude tapering can be used, where x, y are the coordinates of the antenna aperture and $k = 2\pi / \lambda$. However, since the wave emanating from the aperture spreads spherically, the maximum intensity of the field along the axial direction doesn't occur at the focal point. In other words, although the focused antenna is designed to have a spot beam at the focal plane, where all the rays add up in phase, a narrower spot beam can be obtained at a plane closer to the aperture. However, it can be shown that the transverse electric field at this plane, near the axis of the aperture, has almost all the properties of the electric field at the focal plane provided these two planes are close to each other. According to Fig. 2.1, if the focal and maximum intensity points are relatively close to each other we can show that:

$$\mathbf{r}_{1} - \mathbf{r}_{2} \cong \mathbf{r}_{1}' - \mathbf{r}_{2}'$$
 (2.2)

where, r_1 and r_2 are the vectors from $(x_1, y_1, 0)$ and $(x_2, y_2, 0)$ points on the aperture to the focal point, F, respectively. Moreover, r'_1 and r'_2 are the vectors from the same points on the aperture to the maximum intensity point, F', respectively. In other words, although the rays don't add up in phase at point F', the phase difference between them is negligible. Therefore, by applying quadratic phase and proper amplitude distributions to microstrip array antennas, the desired radiation near fields at both focal and maximum intensity planes can be obtained. On the other hand, if the focal point and the maximum intensity points are far from each other, (2.2) won't be true anymore. Therefore, the pattern in the maximum intensity plane is different with the one in the focal plane.



Fig. 2.1. Geometry of the focused aperture.

2.3 Focused Array Antenna Design

In order to illustrate the concept and the method, a 4×4 microstrip patch antenna array is designed to operate at the frequency of 10 GHz. A quadratic phased array with a tapered Dolph-Chebyshev amplitude distribution [42] is applied to the elements of the array to obtain -20dB SLL and a focal distance of 20 cm. An RT/duroid substrate of thickness 0.254 mm was used to minimize the radiation from the feeding network. Each inset-fed patch is 10.02 mm long and 11.47 mm wide and the distance between each two elements is 22.5 mm. Fig. 2.2 (a) shows a photo of the antenna prototype. In order to design the array antenna, the feeding network was first designed by using transmission line theory. Then, IE3D [43], a MoM-based electromagnetic simulator, was used for modeling the feeding network to adjust the length and the width of each line to achieve the required phase shifts and amplitude levels. Finally, the entire feeding network and the array are modeled to achieve the near field radiation patterns. The simulated and measured S₁₁ of the array are shown in Fig. 2.2 (b).



(a)



Fig. 2.2. 4×4 Microstrip patch antenna array (a) photo and (b) reflection coefficient.

The University of Mississippi planar near-field set up is used to measure the near field pattern of the antenna at different planes. Fig. 2.3 shows the measured and simulated electric field of the antenna in the axial direction. It can be seen that the measured and simulated results are in good agreement. The maximum intensity of the electric field is at z = 6.2 cm distance while the focal point is at z = 20 cm distance from the antenna aperture.

Contour plots of the measured near field of the antenna at the focal and maximum intensity planes are depicted in Figs. 2.4(a) and 2.4(b), respectively. It can be seen that the beam is narrower at the maximum intensity plane than that at the focal plane. In addition, Fig. 2.5 shows the simulated and measured E-plane and H-plane near-field radiation patterns of the array antenna at the focal plane. Notice the double representation of the x-axis, which is represented in terms of position and the angular direction. Notice also that the angular axis is not uniform. Moreover, the simulated and measured E-plane and H-plane near-field radiation patterns at the maximum intensity plane are shown in Fig. 2.6. Measured and simulated results are in agreement in both planes especially around the main beams and within $\pm 45^{\circ}$ view angular range. It can be seen that as the observation angle moves beyond 45° from the broad side angle, the agreement between the measured and computed patterns degrades. The reason for the disagreement is that an infinite ground plane is assumed in simulations while the measured results for a finite ground plane are given. It should be noted that the probe is located well beyond the antenna physical aperture where the edge diffraction have stronger effects. Comparing these figures one would see that the half power beamwidth (HPBW) at the maximum intensity plane is less than that at the focal plane. Moreover, SLLs are less than -20 dB in both E and H-planes as expected. A Crosspolarization measurement is also performed and cross-polarization values less than -20 dB are obtained in both planes.



Fig. 2.3. The simulated and measured normalized electric field intensity of the 4×4 microstrip patch antenna array versus the axial distance.



(a)



Fig. 2.4. Measured normalized electric field distribution of the 4×4 focused array antenna at the (a) focal plane, z = 20 cm and (b) maximum intensity plane, z = 6.2 cm.



(a)



Fig. 2.5. The simulated and measured normalized electric field distributions of the 4×4 focused array antenna at the focal plane (a) E-plane and (b) H-plane.



(b)

Fig. 2.6. The simulated and measured normalized electric field distributions of the 4×4 focused microstrip array antenna at the maximum intensity plane (a) E-plane and (b) H-plane.

2.4 Large Array Microstrip Patch Antenna

2.4.1 Theory and Formulation

In order to design large finite arrays, the design procedure would be similar to the procedure used in the previous sections. However, full wave analysis for the array would be difficult to achieve because of the required huge computational resources. To avoid that and have an efficient method for the design of large arrays, we implement a technique used in [44] to predict a large array performance from small array measurements including the mutual coupling effects between the array elements. By knowing the mutual admittance of a small array, the mutual admittance matrix of the large array can be constructed while the coupling beyond the small array size is ignored [44], [45].

If an array of *N* identical elements in the xy-plane is considered, the total electric field pattern of such an array at a distance *r* from the antenna is the summation of the electric field patterns of all elements, given by [42], [44]:

$$E(r) = \sum_{i=1}^{N} A_{i} \left[E_{i}(r - r_{i}) \right] e^{-jk(r-r)}$$
(2.3)

where A_i is the effective excitation voltage of *i*-th element including the mutual coupling effect and $E_i(|\mathbf{r} - \mathbf{r}_i|)$ is the electric field pattern of the same element at \mathbf{r} , and \mathbf{r}_i is the position of that element. In other words, the total electric field pattern of an array can be obtained by the summation of the electric field patterns of the elements while the effective excitation coefficients, A_i , are replaced with the excitation voltage coefficients, V_i . The effective excitation voltage of *i*-th element can be defined in terms of the source voltage as:

$$A_{i} = \frac{\boldsymbol{E}_{i-coup}(\boldsymbol{r})}{\boldsymbol{E}_{i}(\boldsymbol{r})} V_{i}$$
(2.4)

where $E_{i-coup}(\mathbf{r})$ is the electric field at the terminal of the *i*-th element including the mutual coupling effects, i.e. the electric field due to the effective excitation voltage. By some simple manipulations, the effective voltage can be obtained as:

$$A_{i} = \sqrt{\frac{P_{i-coup}}{P_{i}}}V_{i} = \sqrt{\frac{\frac{1}{2}\operatorname{Re}\left[V_{i-coup} \cdot I_{i-coup}^{*}\right]}{\frac{1}{2}\operatorname{Re}\left[V_{i} \cdot I_{i}^{*}\right]}}V_{i}$$
(2.5)

where P_i , V_i and I_i are the radiation power, voltage source and current excitation of *i*-th element, respectively. Moreover, P_{i-coup} , V_{i-coup} , and I_{i-coup} are the radiation power, voltage and current at the antenna terminal including the mutual coupling effect. The current I_i is determined through the relation

$$I_i = Y_{ii}V_i \tag{2.6}$$

where Y_{ii} is the input admittance of *i*-th element. The current I_{i-coup} can also be obtained through the relation:

$$I_{i-coup} = \sum_{j=1}^{N} Y_{ij} V_{j}$$
(2.7)

where Y_{ij} is the mutual admittance between *i*-th and *j*-th elements. By obtaining the effective excitation voltage coefficients of the array, the radiation near field pattern of the array can be obtained using equation (3). In brief, the effective excitation voltage coefficients are introduced to take the mutual coupling effects into account while the pattern multiplication method is used.
2.4.2 Design validity

In order to evaluate the validity and efficiency of this method, an 8×8 microstrip patch array is designed and modeled using the presented method. This array antenna, having the same dimensions and element distances of the presented focused antenna, is solved by the full wave MoM software [43], array pattern multiplication (APM), and the new method, array pattern multiplication including the mutual coupling (APM/MC). The mutual admittance matrix of this array is constructed using the presented focused 4×4 microstrip patch antenna array. The mutual admittance matrix of the presented small array, which is of order 4×4 , is achieved by considering the presence of all the patch elements of the array. The mutual coupling between the elements of the same relative distances and positions in the large matrix is kept the same as in the small matrix. It should be noticed that the mutual coupling between elements.

In this design example, each array element is fed separately by a coaxial feed probe to avoid the feed network spurious radiation contributing to the radiation pattern of the antenna. Two substrate layers can be used and the microstrip feed lines are designed on the bottom side of the lower substrate to eliminate the feed network radiation on the antenna side. A Dolph-Chebyshev amplitude tapering and a quadratic phase distribution with the focal distance equal to 40 cm are considered for this 8×8 array antenna. The simulated E-plane and H-plane near field radiation patterns at the maximum intensity plane, 17.5 cm away from the antenna aperture, are computed using the present method and compared with the results obtained from the full wave analysis based on MoM shown in Fig. 2.7. Good agreement between both results is obtained.



Fig. 2.7. Normalized electric field distributions of the 8 x 8 microstrip patch array antenna at the maximum intensity plane (a) E-plane and (b) H-plane.

In the next example, 8×8 array antennas with smaller distances between elements, $\lambda/2$, are considered to allow stronger coupling between elements and evaluate the performance of the presented method. Two antennas, respectively, with SLLs equal to -20 dB and -40 dB with the same focal distance as the previous example are designed. The mutual admittance matrices of these arrays are constructed using a 4×4 microstrip patch antenna with the same elements and the new distances between elements. Figs. 2.8 and 2.9 show the near field radiation patterns of these antennas at Z = 20 cm away from the antenna aperture in both E and H-planes. It can be seen that by including the mutual coupling in the analysis, accurate near field patterns are obtained. The element patterns used in the APM/MC method are needed to be obtained from one active element within a 4 \times 4 array antenna. In other words, one of the patch elements in the 4 \times 4 array is excited in the presence of all other patches. By using this method, the distortion of the current distribution on the patch due to the mutual coupling is taken into account and more accurate results are achieved. It can be concluded that in the case of high mutual coupling between elements the element pattern within the array is needed. Moreover, the effect of the mutual coupling is more critical when lower SLLs are required [46], [47]. It should be clear that the mutual admittance depends only on frequency, element positions, and their input ports relative to their position in the element. Therefore, designing for certain specifications such as sidelobe levels or beam scanning is controlled by the signals to the ports of the antenna elements. If the mutual coupling is affecting the desired sidelobe levels or the beam scanning direction, one might reevaluate these signals to obtain the desired specification for the array using the same mutual admittance matrix.



Fig. 2.8. Normalized electric field distributions of the condensed 8×8 microstrip patch array antenna with -20dB SLL at Z = 0.2 m away from the antenna aperture (a) E-plane and (b) H-plane.



Fig. 2.9. Normalized electric field distributions of the condensed 8×8 microstrip patch array antenna with -40dB SLL at Z = 0.2 m away from the antenna aperture (a) E-plane and (b) H-plane.

2.4.3 Large Focusing Array Antenna

In this section, two focused large array antennas, a 16×16 and a 64×64 microstrip patch array, are considered. Both antennas are designed based on the same element type and size of the presented focused antenna. The 16×16 array has the distance between elements of the condensed array antenna of $\lambda/2$. However, the 64×64 array has the same distances between elements of the 4×4 presented focused antenna, which is $3\lambda/4$. The mutual admittance matrices of these large arrays are constructed using 4×4 microstrip patch antenna arrays with the same dimensions and distances between elements.

For the 16 × 16 array antenna, Dolph-Chebyshev amplitude tapering and various quadratic phase distributions are considered to focus the power at Z = 0.5, 1, 1.5 and 2 meters away from the antenna aperture and all have -20 dB SLLs. The element pattern within the array is considered to obtain the near field pattern of this array. The transverse electric field distributions of this antenna versus the axial distance are shown in Fig. 2.10. All the plots are normalized to the highest maximum intensity value to show a comparison between these values for different focal distances. It can be seen that the maximum intensity of the electric field doesn't occur at the focal point. For the focal point close to the antenna aperture, the maximum intensity is close to the focal point but as the focal point moves away from the antenna aperture, the focal shift increases. In other words, by increasing the focal distance, the maximum intensity distance increases with a much lower rate and cannot go beyond a certain distance. This is because of the spherical spreading of the wave-front away from the source. In other words, the magnitude of the field emanating from the aperture is inversely proportional to the distance, r, and as the focal distance increases, the effect of this spreading factor on the focal shift increases.

The focusing properties of this antenna are summarized in Table 2.1. As the focal point moves away from the antenna aperture the electric field intensity at the maximum intensity point and the depth of focus decreases and the forlobe levels increase. Therefore, choosing a large focal distance, much larger than the antenna aperture, deteriorates the focusing properties of the antenna.

Fig. 2.11 shows the electric field distributions of the antenna at the focal planes for both E and H-planes. It can be seen that the desired sidelobe levels at all the focal distances are achieved. It should be noted that since the near field patterns are plotted at different distances from the antenna aperture, they have different angular view ranges. Therefore, the near field patterns for the smaller focal distances are observed in wider view angular ranges. The electric field distributions at the maximum intensity planes are shown in Fig 2.12. It can be observed that the desired sidelobe levels are achieved for the lowest focal distance. However, by increasing the focal distance the sidelobe levels are increased (See Table 2.1). The sidelobe deterioration is due to the increased distance between the focal plane and maximum intensity plane position.

In the next design example a 64×64 array with Dolph-Chebyshev amplitude tapering and various quadratic phase distributions is considered to focus the power at Z = 3, 5, 7, 9 and 11 meters distances from the antenna aperture and all have -40 dB SLLs. Fig. 2.13 shows the transverse electric field distributions on the axis of the antenna versus the axial distance from the aperture. It can be seen that similar to the previous antenna, for the focal point close to the antenna aperture, the maximum intensity is very close to the focal point, but as the focal point moves away from the antenna aperture, the focal shift increases. The focusing properties of this antenna are summarized in Table 2.2. It can be seen that lower forlobe levels are obtained for this large array antenna compared to the previous one.



Fig. 2.10. Normalized Electric field intensity of the 16×16 microstrip patch array antenna versus the axial distance for different focal distances.



Fig. 2.11. Normalized electric field distributions of the 16×16 microstrip patch array antenna at the focal plane for various focal distances (a) E-plane and (b) H-plane.



(b)

Fig. 2.12. Normalized electric field distributions of the 16×16 microstrip patch array antenna at the maximum intensity plane for various focal distances (a) E-plane and (b) H-plane.

Focal Distance (m)	0.5	1	1.5	2
Electric Field Maximum Intensity (dB)	0	-3.28	-4.66	-5.41
Maximum Intensity Distance (m)	0.35	0.47	0.52	0.55
Depth of Focus (m)	0.341	0.573	0.685	0.75
Forelobe levels (dB)	-5.2	-2.54	-1.42	-0.83
HPBW (m)	0.047	0.065	0.075	0.081
Max sidelobe level (dB)	-20.0	-18.85	-17.82	-17.41

Table 2.1. Focusing properties of 16×16 array antenna for different focal distances.



Fig. 2.13. Normalized Electric field intensity of the 64×64 microstrip patch array antenna versus the axial distance for different focal distances.

The normalized E-plane and H-plane electric field distributions of the antenna at the maximum intensity planes for various focal distances are shown in Fig. 2.14. It can be seen that by moving the focal point away from the antenna aperture, HPBW increases. Moreover, it should be noticed that the SLLs are about -40dB at the maximum intensity planes for various focal distances since the electric field patterns are shown in a small angular range from the axial axis of the antenna. It should be also noticed that by increasing the focal distance, leading to the increase of the focal shift, higher SLLs appear.

Focal Distance (m)	3	5	7	9	11
Electric Field Maximum Intensity (dB)	0	-4.1	-6.7	-8.5	-9.8
Maximum Intensity Distance (m)	2.87	4.57	6.03	7.22	8.23
Depth of Focus (m)	0.90	2.28	4.01	5.84	7.33
Forelobe levels (dB)	-12.3	-8.5	-7.0	-5.9	-5.0
HPBW (m)	0.076	0.121	0.161	0.197	0.228

Table 2.2. Focusing properties of 64×64 array antenna for different focal distances.



Fig. 2.14. Normalized electric field distributions of the 64×64 microstrip patch array antenna at the maximum intensity plane for various focal distances (a) E-plane and (b) H-plane.

2.5 Conclusion

A new scheme in designing focused array antennas with desired SLLs in the near field region was presented. A 4 x 4 array microstrip patch antenna with beam focused in the radiation near field region was examined. Moreover, large focused array antennas based on the knowledge of the mutual admittances of a smaller array was predicted. The effects of various focal distances on the near field pattern of these antennas were investigated. It was shown that the maximum intensity of the electric field was shifted toward the antenna aperture from the focal point where all of the rays contribute in phase because of the quadratic phase distribution on the aperture. In addition, it was shown that if the focal and maximum intensity points are close to each other, desired sidelobe levels at the maximum intensity plane, close to the axis of the antenna, can be achieved.

Chapter 3

FOCUSING PROPERTIES OF FRESNEL ZONE PLATE LENS ANTENNAS

3.1 Introduction

Array antennas [1], [2], [28] and conventional reflector antennas [29], [30] have been widely used to generate quadratic phase distributions and focus the power at nearby points. However, a precise generation of a quadratic phase distribution for large array antennas is complex, costly and limited due to the difficulties in implementing the beam forming networks. On the other hand, the ability to produce a desired aperture distribution using a conventional reflector like parabolic reflector antenna is limited. Although the problem can be solved using shaped reflectors, their implementation is very costly.

Another alternative focused antenna is the Fresnel zone plate (FZP) lens consisting of a set of alternating open and opaque annuli arranged on flat or curvilinear surfaces [48]-[50]. The plane FZP lens has the benefit of being lighter and easier to design and manufacture compared to the array antennas and reflectors. Moreover, FZP lenses are lighter and thinner than traditional

lens antennas especially when a large antenna aperture is needed. In fact, the flatness aspect of the FZP antenna is very advantageous in the manufacturing process.

The focusing behavior of the plane and spherical FZP antennas was first presented in [51]. Later, some axial defocusing characteristics of FZP lens antennas were presented in [32], [33]. In [34] the focusing behaviors of curvilinear and plane FZP lenses were studied. It was shown that the curvilinear FZP lenses don't necessarily have superior focusing abilities compared to the plane FZP lenses. In [35] it was shown that zone plate and hyperbolic lens antennas have similar focusing properties except that the FZP antennas deliver less power to their focal points.

In this chapter, some new focusing characteristics of Soret FZP lens antennas are examined. Although it is well known that phase corrected (Wood-type) zone plate antennas are more efficient than Soret zone plate antennas, the latter one is much simpler to design and fabricate in order to generally represent the focusing properties of FZP lens antennas. Different FZP lens antennas are designed to investigate the effect of focal lengths on both axial and transverse near-field patterns. Simulation and measurement results show the displacement of the maximum intensity of the electric field along the axial direction. Finally, the scanning characteristics of the FZP antenna focused beam are presented.

3.2 Design Procedure

The geometrical optic (GO) method is used in the design of FZP antennas to achieve constructive interference at the focal point. Fig. 3.1 shows the two dimensional configuration of the FZP fed by a circular corrugated horn antenna. The design is performed in the plane (two-dimensional) and then the shape is completed by revolving it around the axis of symmetry.

The FZP is designed as circular concentric zones with radius of r_m for the *m*-th zone. D_m is the diameter and F_1 and F_2 are the focal lengths of the FZP. The values of r_m should be determined such that a ray emanating from one focal point adds up in phase at the other focal point by satisfying

$$L_{1m} + L_{2m} - (F_1 + F_2) = m\lambda/2$$
(3.1)

where *m* is an integer, λ is the wavelength, and L_{1m} and L_{2m} are the distances between the *m*-th ring and the first and second focal points, respectively. This equation can be rewritten as

$$\sqrt{F_1^2 + r_m^2} + \sqrt{F_2^2 + r_m^2} = m\lambda/2 + F_1 + F_2.$$
(3.2)

The exact values of r_m can be obtained by solving this equation.



Fig. 3.1. Two-dimensional configuration of the FZP fed by a circular horn antenna.

3.3 Simulated and Measured Results

FZP antennas with diameter of 0.16 m and various focal lengths, $F_2 = 0.05$ m, 0.075 m, 0.15 m, 0.30 m, 0.45 m, 0.6 m, 0.75 m and 0.9 m, are designed and modeled by a full wave MoM solution [52]. The diameter and the number of zones of each FZP are presented in Table 3.1. It should be mentioned that the antennas with first three focal lengths are focused in the near-field region and the others are focused in the Fresnel region of the antennas. A corrugated circular horn antenna with left-handed circular polarization (LHCP) is used as the antenna feed. Fig. 3.2 shows the V and H cut measured radiation pattern of the horn antenna at the frequency of 32 GHz. A subtended angle of 40 degrees is chosen to have an edge taper value of about -15 dB on FZP rims with the focal length of $F_1 = 0.095$ m.

Focal length (m)	Number of Zones	Diameter (m)
0.05	16	0.168
0.075	14	0.158
0.15	10	0.156
0.3	8	0.155
0.45	8	0.163
0.6	8	0.167
0.75	6	0.171
0.9	6	0.171

Table 3.1. The number of zones and diameter of each Zone plate

In order to verify the simulated results, two FZP antennas with $F_2 = 0.15$ m and 0.45 m are fabricated by etching on an RT/duroid 5880 laminate. Fig. 3.3 illustrates the antenna under test at the University of Mississippi planar near-field set up.

The simulated and measured transverse electric field distributions along the axial direction for the FZP with the focal length of 0.15 m are shown in Fig. 3.4. A good agreement between the simulated and measured results is observed. The maximum intensity of the electric field occurred at z = 0.137 m distance from the antenna aperture. The simulated and measured electric field distribution of the antenna at the focal plane (z = 0.15 m) and the maximum intensity plane (z = 0.137 m) are shown in Figs. 3.5 and 6, respectively. Notice the triple representation of the x-axis in terms of position, angular direction and radial distance. It can be seen that simulated and measured results are in agreement especially around the main beam.



Fig. 3.2. V and H cuts of the measured radiation pattern of the LHCP feed horn antenna at the frequency of 32 GHz.



Fig. 3.3. FZP lens antenna test set-up.



Fig.3.4. Simulated and measured normalized electric field intensity of the FZP antenna with the focal length of 0.15 m along its axial direction.

The electric field distribution of the FZP antenna with the focal length of 0.45 m is also measured in different planes. Fig. 3.7 shows the simulated and measured electric field

distribution of the antenna versus the axial distance. The maximum intensity of the electric field for both the simulated and measured results occur at z = 0.343 m distance from the antenna aperture. Figs. 3.8 and 3.9 show the simulated and measured near field pattern at the focal plane (z = 0.45 m) and the maximum intensity plane (z = 0.343 m), respectively. Comparing the near field distributions at the focal plane and the maximum intensity plane for both FZP antennas, one can observe that a narrower beam is obtained at the maximum intensity plane compared to the focal plane.



Fig. 3.5. Simulated and measured normalized electric field intensity of the FZP antenna with the focal length of 0.15 m at the focal plane (z = 0.15 m).



Fig. 3.6. Simulated and measured normalized electric field intensity of the FZP antenna with the focal length of 0.15 m at the maximum intensity plane (z = 0.137 m).



Fig. 3.7. Simulated and measured normalized electric field intensity of the FZP antenna with the focal length of 0.45 m along its axial direction.



Fig. 3.8. Simulated and measured normalized electric field intensity of the FZP antenna with the focal length of 0.45 m at the focal plane (z = 0.45 m).



Fig. 3.9. Simulated and measured normalized electric field intensity of the FZP antenna with the focal length of 0.45 m at the maximum intensity plane (z = 0.34 m).

In order to evaluate the focusing properties of FZP antenna, the electric field distributions versus the axial distance for FZP antennas with different focal lengths are plotted in Fig. 3.10. It can be seen that the maximum intensity of the electric field is displaced from the focal point toward the antenna aperture. Fig. 3.11 shows this displacement for different focal lengths. The normalized maximum intensity values of the FZP antenna versus the focal length are shown in Fig. 3.12. It is observed that as the focal point moves away from the antenna aperture the focal displacement increases and the maximum intensity value decreases. It should be noted that by increasing the focal length, the maximum intensity length increases with a lower rate and cannot go beyond a certain distance. The reason for this displacement is the spherical spreading of the wave front away from the source. Since the magnitude of the radiating field decays as 1/r, where r is the distance from the source, by increasing the focal length, the effect of this spreading factor on the focal displacement increases. Although, all rays emanating from the source contribute in phase at the focal point, they add up partly in phase with higher intensities at closer points to the aperture causing the occurrence of the maximum intensity point and therefore the focal shift.



Fig. 3.10. Normalized electric field intensity of the FZP antenna versus the axial direction for different focal lengths.



Fig. 3.11. Focal displacement of FZP antenna versus the focal length



Fig. 3.12. Normalized maximum intensity values of the FZP antenna versus the focal length.

The normalized electric field distributions of the antenna for different focal lengths at the maximum intensity planes and focal planes are shown in Figs. 3.13 and 3.14, respectively. It can be observed that by increasing the focal length, the half power beam width (HPBW) at both the focal plane and maximum intensity plane increases. It should be noted that since the near field patterns are plotted at different distances from the antenna aperture, the pattern for the smaller focal lengths are observed in wider range of view angles. The variations of HPBW and depth of focus versus the focal lengths are depicted in Fig. 3.15. The depth of focused is defined as the distance between the axial -3dB points about the maximum intensity plane [15]. It is seen that by increasing the focal length, not only the HPBW but also the depth of focus, increases. The variation of the sidelobe levels (SLL) of the electric field patterns at maximum intensity planes and focal planes versus the focal length are shown in Fig. 3.16. It can be seen that by increasing the focal length in the Fresnel region, SLLs of the electric field patterns at both maximum

intensity and focal planes are increased. Sidelobe levels in the maximum intensity plane increase faster than that in the focal plane since by increasing the focal length, the focal shift increases. Therefore, increasing the focal length of the antenna causes a wider focusing beam and higher SLLs.



Fig. 3.13. Normalized electric field distribution of the FZP at the focal plane for different focal lengths.



Fig. 3.14. Normalized electric field distribution of the FZP at the maximum intensity plane for different focal lengths.



Fig. 3.15. Variations of depth of focus and half power beam width of the electric field pattern at maximum intensity plane versus the focal length of the FZP antenna.



Fig. 3.16. Variations of sidelobe level of the electric field pattern at maximum intensity planes and focal planes versus the focal length of the FZP antenna.

3.4 Beam Scanning by Feed Motion

Scanning the antenna's focused beam is of interest in many applications. Although it is costly and complicated to scan the large antenna beam using phased arrays, it can be done easily in FZP antennas.

The beam scanning is accomplished by displacing the feed horn antenna along its transverse axis. Fig. 3.17 shows the simulated and measured electric field distribution of the FZP with the focal length of 0.45 m at the maximum intensity plane for various feed displacements. These results are summarized in Table 3.2. It can be seen that the beam is steered by approximately 0.015 m for each 0.005 m displacement of the feed. It should be mentioned that the steered beams have approximately the same maximum intensity length, depth of focus and HPBW as the original focused antenna. In other words, although the feed is displaced from the FZP focal point, the scanned beams degradations are negligible.

Feed Displacement (m)	0	0.005	0.010	0.015	0.020
Focused beam displacement (m)	0	-0.0148	-0.0297	-0.0447	-0.0602
SLL (dB)	-14.51	-13.87	-13.68	-13.01	-12.44

Table 3.2. The Steered focused beam properties of the FZA with the focal length of 0.45 m.



(a)



(b)

Fig. 3.17. (a) Simulated and (b) measured steered focused beam of the FZP antenna with the focal length of 0.45 m at the maximum intensity plane for various feed displacements.

3.5 Conclusion

Electric field pattern characteristics of the focused Fresnel zone plate lens antenna in the near-field region were presented. The FZP antenna fed by a circular corrugated horn was implemented and the effects of various focal lengths on the near field pattern of this antenna were examined. It was shown that the maximum intensity occurred closer to the focal point and this displacement was increased as the focal point moved away from the antenna aperture. By increasing the focal length of the antenna, HPBW, depth of focus and side lobe levels were increased. In addition, the focused beam scanning of the FZP antenna was implemented by displacing the feed from the FZP focal point.

Chapter 4

LARGE CURRENT RADIATOR (LCR)

4.1 Introduction

The Large Current Radiator (LCR), which is one of the promising UWB radiators, was proposed by F. Harmuth in 1977 [53]-[55]. This radiator is named Large Current Radiator because it is possible to create a large amplitude current in the radiating element with a relatively small driving voltage. Principally the LCR is a segment of a line conductor through which a current pulse of short duration but large amplitude is driven. Fig 4.1 shows the configuration of an LCR consisting of a generator, a closed loop and a ferrite plate. The LCR is inherently a non-resonating structure and thus permits to radiate electromagnetic (EM) waves with either sinusoidal or non-sinusoidal time variation [56]-[60].



Fig. 4.1. The schematic of an LCR.

4.2 LCR Modeling

Fig. 4.2 shows the dimension of the chosen LCR. In order to model the LCR with the Ferrite plate, only the radiator is modeled while the backward radiation is ignored to model the effect of the ferrite plate. The antenna is modeled in the frequency domain using a MoM based software (FEKO) from 50MHz to 3GHz frequency band and the time-domain radiated pattern is computed by using the inverse Fourier transform. The time-domain pattern, $F(\theta, \varphi, t, r)$, can be shown at certain time instances ($t = t_0$), distances ($r = r_0$) from the aperture, or observation angles ($\theta = \theta_0, \varphi = \varphi_0$).



Fig. 4.2. The configuration of modeled LCR.



Fig. 4.3. Input voltage pulse excited the large current radiator.

The LCR is excited by a voltage pulse with amplitude of 1 V and duration of 5.4 ns as shown in Fig. 4.3. The radiated field by the LCR at a distance of z = 1.2 m from the antenna is shown in Fig. 4.4. The time-domain near-field pattern of the LCR at the distance of z = 1.2 m from the antenna in both E and H-plane are shown in Fig. 4.5.



Fig. 4.4. Time variation of pulse radiated by the large current radiator at broadside observation angles ($\theta = 0$, $\varphi = 0$) and a distance of r = 1.2 m from the antenna.

In the other effort, the LCR is excited by a Gaussian pulse G(t) with different variances.

$$G(t) = \frac{1}{\sqrt{2\pi\sigma^2}} e^{-(t-\mu)^2/2\sigma^2}$$
(4.1)

where σ is the standard deviation and μ is the mean.

The advantage of the Gaussian pulse over the step pulse is that it has a limited bandwidth. The input Gaussian pulses are shown in Fig 4.6 and the radiated pulses at the distance of the z = 1.2 m from the LCR antenna are shown in Fig. 4.7. The near field pattern obtained by exciting the LCR with the Gaussian pulse is shown in Fig. 4.8.



(b)

Fig. 4.5. Time-domain near-field pattern of the LCR at the time instance of t = 1.2/c, and the distance of r = 1.2 m from the antenna aperture in the (a) E-Plane and (b) H-plane.


Fig. 4.6. Input Gaussian pulse (a) time-domain and (b) frequency domain.



(a)



Fig. 4.7. Pulse radiated by the LCR at broadside observation angles ($\theta = 0$, $\varphi = 0$) and a distance of r = 1.2 m from the antenna (a) Time-domain and (b) Frequency domain.



Fig. 4.8. Time-domain near-field pattern of the LCR at the time instance of t = 1.2/c and a distance of r = 1.2 m from the antenna aperture in the (a) E-Plane and (b) H-plane.

(b)

4.3 Mutual Coupling Effects

It has been shown that LCR elements can be employed to achieve antenna arrays with high efficiencies [61]-[63]. This can be done by spacing the LCRs at a distance much larger than the spatial duration of the radiated field pulse. In this case, there is no coupling between elements and the array antenna pattern can be obtained by the summation of each element pattern. In order to show the accuracy of this method, two LCRs spaced at a distance d = 0.3 m are considered. One of the elements is excited by a step pulse and the other one is terminated to a matched load. Fig 4.9 shows the near field pattern of this configuration at both E and H-planes compared to the near field pattern of one LCR element. It can be seen that very similar patterns are obtained for both planes. In other words, the presence of the loaded antenna at a distance larger than the spatial duration of the radiated pulse doesn't affect the radiation pattern of the LCR.

4.4 Conclusion

The radiation near field properties of LCR was examined. An actual LCR excited by both step function and Gaussian pulses and modeled by a MoM solution. In addition, the effect of the mutual coupling between two LCRs spaced at a distance larger than spatial duration of the radiated pulse was presented.







(b)

Fig. 4.9. Comparison between the near field pattern of one LCR and two LCR element array at the time instance of t = 1.2/c and a distance of r = 1.2 m from the antenna aperture in the (a) E-plane and (b) H-plane.

Chapter 5

INVASIVE WEED OPTIMIZATION AND ITS FEATURES IN ELECTROMAGNETICS

5.1 Introduction

Electromagnetic design problems usually involve several parameters that are non-linearly related to the objective functions. In order to solve these problems efficiently, evolutionary optimization algorithms have been considered and successfully applied to electromagnetic problems. Among these optimizers, genetic algorithm (GA) [61] and particle swarm optimization (PSO) [62] have received considerable attention by the electromagnetic community due to their efficiency and simplicity [63]-[66]. In addition, other optimization methods including Ant Colony Optimizer (ACO) [67] and Simulated Annealing (SA) [68] have shown high capability of searching for a global minimum in electromagnetic optimization problems [69]-[73].

Here, a new optimization algorithm, invasive weed optimization (IWO) and some of its new features are introduced by illustrating its application to various electromagnetic problems. This numerical stochastic optimization algorithm, inspired from weed colonization, was first introduced by Mehrabian and Lucus in 2006 [74]. It is shown that this optimizer not only outperforms other optimizers like PSO in certain instances, but also is capable of handling some new electromagnetic optimization problems.

The main purpose of this chapter is to introduce the desirable attributes and new features of the IWO for electromagnetic problems to be applied to the design of the focusing antenna array. Of course, the efficiency of this optimization method compared to the other optimizers depends on the problem and the choice of control parameters. Below, we first represent the proposed IWO algorithm and its desirable features. Then, by conducting several array antenna synthesis problems, including linear and thinned array antennas, the efficiency and specific features of this new algorithm are shown. Finally, the method is employed in designing a U-slot microstrip patch antenna fed by an L-probe to have the desired reflection coefficient for dualband applications.

5.2 IWO

5.2.1 The Inspiration Phenomenon

The IWO, inspired from the phenomenon of colonization of invasive weeds in nature, is based on weed biology and ecology. It has been shown that capturing the properties of the invasive weeds leads to a powerful optimization algorithm. The behavior of weed colonization in a cropping field can be explained as follows: Weeds invade a cropping field (system) by means of dispersal and occupy opportunity spaces between the crops. Each invading weed takes the unused resources in the field and grows to a flowering weed and produces new weeds through its seeds, independently. The number of new weeds produced by each flowering weed depends on the fitness of that flowering weed in the colony. Those weeds that have better adaptation to the environment and take more unused resources grow faster and produce more seeds. The new produced weeds are randomly spread over the field and grow to flowering weeds. This process continues till the maximum number of weeds is reached on the field due to the limited resources. Now, only those weeds with better fitness can survive and produce new weeds. This competitive contest between the weeds causes them to become well adapted and improved over the time.

5.2.2 Algorithm

Before considering the algorithm process, the new key terms used to describe this algorithm should be introduced. Table 5.1 shows these terms. Each individual or agent, a set containing a value of each optimization variable, is called a seed. Each seed grows to a flowering plant in the colony. The meaning of a plant is one individual or agent after evaluating its fitness. Therefore, growing a seed to a plant corresponds to evaluating an agent's fitness.

To simulate the colonizing behavior of weeds the following steps, pictorially shown in Fig. 5.1, are considered:

1. First of all, the *N* parameters (variables) that need to be optimized should be selected. Then, for each of these variables in the *N*-dimensional solution space, a maximum and minimum value should be assigned (Define the solution space).

- 2. A finite number of seeds are randomly dispersed over the defined solution space. In other words, each seed takes a random position in the N-dimensional problem space. Each seed's position is an initial solution, containing N values for the N variables, of the optimization problem (Initialize a population).
- 3. Each initial seed grows to a flowering plant. That is, the fitness function, defined to represent the goodness of the solution, returns a fitness value for each seed. After assigning the fitness value to the corresponding seed, it is called a plant (Evaluate the fitness of each individual).
- 4. Before the flowering plants produce new seeds, they are ranked based on their assigned fitness values. Then, each flowering plant is allowed to produce seeds depending on its ranking in the colony. In other words, the number of seeds produced by each plant depends on the rank of the seed in the colony and increases from the minimum possible seed production, s_{min} , to its maximum, s_{max} . Those seeds that solve the problem better correspond to the plants which are more adapted to the colony and consequently produce more seeds. This step adds an important property to the algorithm by allowing all of the plants to participate in the reproduction contest (Rank the population and reproduce new seeds).

Agent/ Seed	Each individual in the colony containing values of optimization variables
Fitness	A value representing the goodness of the solution for each seed
Plant	one agent/seed after evaluating its fitness
Colony	The entire agents or seeds
Population Size	The number of plants in the colony
Maximum number of plants	The maximum number of plants allowed to produce new seeds in the colony

Table 5.1. Some of the key terms used in the IWO



Fig. 5.1. Flow Chart showing the IWO algorithm.

5. The produced seeds in this step are dispersed over the search space by normally distributed random numbers with mean value equal to the location of the producing plants and varying standard deviations. The standard deviation (SD) at the present time step can be expressed by:

$$\sigma_{iter} = \frac{(iter_{max} - iter)^n}{(iter_{max})^n} (\sigma_{initial} - \sigma_{final}) + \sigma_{final}$$
(5.1)

where *iter*_{max} is the maximum number of iterations, $\sigma_{initial}$ and σ_{final} are defined initial and final standard deviations, respectively, and *n* is the nonlinear modulation index. Fig. 5.2 shows the standard deviation (SD) over the course of a run with 100 iterations and different modulation indexes. It can be seen that the SD is reduced from the initial SD to the final SD with different rates. The algorithm starts with such a high initial SD that the optimizer can explore through the whole solution space. By increasing the number of iterations, the SD value is decreased gradually to search around the local minima or maxima to find the global optimal solution (Dispersion).



Fig. 5.2. Standard deviation over the course of the run.

- 6. After that all seeds have found their positions over the search area, the new seeds grow to the flowering plants and then they are ranked together with their parents. Plants with lower ranking in the colony are eliminated to reach the maximum number of plants in the colony, P_{max} . It is obvious that the number of fitness evaluations, the population size, is more than the maximum number of plants in the colony (Competitive exclusion).
- Surviving plants can produce new seeds based on their ranking in the colony. The process is repeated at step 3 till either the maximum number of iterations is reached or the fitness criterion is met (Repeat).

5.2.3 Selection of Control Parameter Values

Among the parameters affecting the convergence of the algorithm, three parameters, the initial SD, $\sigma_{initial}$, the final SD, σ_{final} , and the nonlinear modulation index, *n*, should be tuned carefully in order to achieve the proper value of the SD in each iteration according to (5.1). A high initial standard deviation should be chosen to allow the algorithm to explore the whole search area, aggressively. It seems that the IWO works well if the initial SD is set around 1 to 5 percent of the dynamic range of each variable. The final SD should be selected carefully to allow the optimizer to find the optimal solution as accurately as possible. A finer local optimum solution can be achieved by decreasing this parameter. However, it should be noticed that tuning the final SD much smaller than the precision criteria of the optimization variables doesn't improve the final SD in each dimension should be selected based on the precision effect of that variable on the objective function. It was shown that the value of nonlinear modulation index has a considerable effect on the performance of IWO [74]. It was suggested that the best choice for *n*

is 3. Besides (5.1), other functions to describe the standard deviation over the optimization process were considered. However, simulation results showed that (5.1) with n = 3 is the best choice.

Maximum and minimum numbers of seeds are the two other important parameters needed to be selected. Based on different examples, it can be concluded that selecting the maximum number of seeds between 3 and 5 leads to a good performance of the optimizer. Moreover, the minimum number of seeds is set to one for all examples.

The maximum number of plants is another parameter that should be chosen in the IWO. Parametric studies show that increasing this parameter does not necessarily increase the performance of the algorithm. It was found that the best performance can be achieved for many problems when the maximum number of plant is set between 10 and 20.

5.3 IWO Features

One important property of the IWO is that it allows all of the agents or plants to participate in the reproduction process. Fitter plants produce more seeds than less fit plants, which tends to improve the convergence of the algorithm. Furthermore, it is possible that some of the plants with the lower fitness carry more useful information compared to the fitter plants. The IWO gives a chance to the less fit plants to reproduce and if the seeds produced by them have good fitness in the colony, they can survive.

Another important feature of IWO is that weeds reproduce without mating. Each weed can produce new seeds, independently. This property adds a new attribute to the algorithm that each agent may have different number of variables during the optimization process. Thus, the number of variables can be chosen as one of the optimization parameters in this algorithm. By optimizing the number of variables, some new electromagnetic design problems can be handled. The effectiveness of this kind of optimization for designing aperiodic thinned array antennas is shown in section 5.5.

Finally, comparing some aspects of the IWO with two common and standard optimizers, GA and PSO, can clarify some features of this new algorithm. Provided that the number of iterations and the population size are considered as common requirements for all evolutionary algorithms, the initial and final standard deviation, nonlinear modulation index, and maximum and minimum number of seeds are the parameter of the IWO need to be tuned. In the GA, crossover and mutation rates and in the PSO, inertial weight, W, cognitive rate, c_1 , social rate, c_2 , and the maximum velocity, V_{max} should be controlled to achieve the desired convergence. It has been shown that the choice of boundary conditions and also the maximum velocity are critical in convergence of the PSO algorithm [75]-[77]. Moreover, in the case of GA, both crossover and mutation rates affect the convergence of the problem [78]. The effect of these tuning parameters on the GA and PSO convergences are difficult to perceive, but by tuning the critical parameters in the IWO, the initial and final SD, a high-level control in the convergence and accuracy of the algorithm is achieved [79]-[80]. In addition, the IWO shows a high stability with different boundary conditions.

5.4 Array Antenna Design Problems

In this section, both the IWO and PSO are applied to the problem of synthesizing the farfield radiation patterns of linear array antennas. The consideration focuses on the optimizing of array antennas to achieve the desired radiation patterns given by user defined functions. For an array antenna with *N* elements, separated by a uniform distance *d*, the normalized array factor is given by

$$AF(\theta) = \frac{1}{AF_{\max}} \sum_{n=1}^{N} I_n e^{j2\pi nd \sin \theta / \lambda}$$
(5.2)

where I_n are amplitude coefficients, θ is the angle from the normal to the array axis, AF_{max} is the maximum value of the magnitude of the array factor, and d is assumed to be $\lambda/2$, where λ is the wavelength.

Comparisons are made between the performances of the IWO and the PSO in achieving desired radiation patterns. In the case of the IWO, restricted and invisible boundary conditions are two possible choices. The restricted boundary condition relocates the particle on the boundary that the particle hits. However, the invisible boundary condition allows a particle to stay outside the solution space while the fitness evaluation of that particle is skipped and a bad fitness value is assigned to that errant particle. The PSO with the Invisible Boundary Condition (IBC), Reflective Boundary Condition (RBC), Absorbing Boundary Condition (ABC), Damping Boundary Condition (DBC), invisible/reflecting boundary condition, and invisible/damping boundary conditions are tested [76]. In addition, the velocity-clipping technique, showing good performance in the PSO, is implemented for different V_{max} values [69], [75].

The same population size and number of iterations are chosen for different algorithms. The population size is fixed to 40 for both algorithms. It should be noted that in the case of the IWO, the population size is fixed by choosing the maximum number for the plant population and also minimum and maximum number of seeds. In the coming examples, the maximum number of plants is fixed to 10 and the number of seeds increases linearly from 1 to 5. In the case of PSO, both the cognitive rate (c_1) and the social rate (c_2) are set to 2.0 and the inertial weight is varied linearly from 0.9 to 0.2 as suggested in [69], [77]. It should be also pointed out that various realizations of the same experiment produced results that are close to each other. All results reported are the average of 50 independent runs of the PSO or IWO algorithms and found to be sufficient.

5.4.1 Optimizing Sidelobe Patterns

In this section, a linear 40-element array is considered to achieve the desired radiation pattern by optimizing the amplitude coefficients. The objective pattern is to obtain sidelobe levels less than a tapered sidelobe mask that decreases linearly from -40dB to -50dB. The beamwidth of the array pattern is 11° and the number of sampling points is 359. It should be pointed out that a "Don't exceed criterion" is utilized in the formulation of the objective function, which means an error will be reported only if the sidelobe levels of the array factor exceed the desired levels. The cost function is defined as follow

$$Cost = \sum_{m=1}^{N_{ob}} \left| AF_d(m) - AF_o(m) \right|$$
(5.3)

where $AF_d(m)$ and $AF_o(m)$ are the desired and obtained array factors at the *m*th observation angle, respectively, and N_{ob} is the number of observation points.

Fig. 5.3 (a) shows the desired and obtained radiation patterns achieved by the IWO. Since the desired envelope is symmetric, we exploit the symmetry of the current distribution. Thus, the number of optimization parameters reduces to half of the array elements. The parameters used for the IWO are summarized in Table 5.2. The performance of the IWO compared to the PSO for different boundary conditions is shown in Fig. 5.3 (b). It can be seen that the performance of the PSO is dramatically changed by choosing different boundary conditions or changing the maximum velocity. Moreover, the algorithm is trapped in a local minimum when the absorbing boundary condition is used while IWO achieves better performance for both invisible and restricted boundary conditions. The PSO algorithm tested by different boundary conditions [76]-[77] doesn't show any better performances. The average numbers of fitness evaluations required per successful run for both the IWO and PSO with different boundary conditions are shown in Table 5.3. It can be seen that the IWO is faster than the PSO to achieve the same optimization goal for this problem.

The performance of the IWO for different standard deviations is shown in Fig. 5.3 (c). Different initial and final standard deviations are tried for IWO with the restricted boundary condition to evaluate the performance of this algorithm. It can be observed that by changing these parameters, the performance of the algorithm is slightly changed. Thus, by applying different boundary conditions or different standard deviation parameters, the IWO shows more stability compared to the PSO. The desired and obtained radiation patterns obtained by the PSO for different boundary conditions are shown in Fig. 5.4. It is seen that the obtained SLLs are above the desired mask for ABC, DBC, and RBC with $V_{max} = 0.3$. It should be mentioned that the 50 independent runs of the same experiment for each curve of the IWO algorithm are closer to each other compared to those for the PSO. These results for the PSO and the IWO are shown in Fig. 5.5 and 5.6, respectively. It can be seen that the curves in Fig. 5.6 are closer to each other compared to those in Fig. 5.5. In other words, the IWO shows more stability compared to the PSO for different runs.

it _{max}	p_{max}	S _{max}	S _{min}	п	initial SD	final SD
1000	10	5	0	3	0.015	0.00005

Table 5.2. IWO parameter values for the linear 40-element array optimization

Table 5.3. Comparison of average number of fitness evaluations required per successful run in

the PSO and the IWO algorithms for the linear 40-element array optimization.

Algorithm	Average number of fitness evaluations
	required per successful run
PSO-ABC	-
PSO-DBC	162328
$PSO-RBC (V_{max} = 1)$	92485
$PSO-RBC (V_{max} = 0.3)$	62677
PSO-RBC ($V_{max} = 0.1$)	42329
PSO-IBC	52496
IWO-Restricted BC	14692
IWO-Invisible BC	18212



(a)



(b)



Fig. 5.3. Amplitude-only synthesis for a linear 40-elemet array (a) radiation pattern obtained using IWO, (b) convergence curves for the IWO and PSO with different boundary conditions. The maximum velocity limit is changed for the PSO, and (c) convergence curves for the IWO with restricted boundary condition and different values of initial and final standard deviations.







(b)



Fig. 5.4. Radiation patterns of the linear 40-elemet array obtained by PSO for (a) absorbing boundary condition, (b) damping boundary condition, (c) reflective boundary condition with V_{max} = 0.3, and (d) invisible boundary condition.



(b)



(c)



(d)



Fig. 5.5. Convergence curves of 50 independent runs for a linear 40-elemet array using the PSO with (a) absorbing boundary condition, (b) damping boundary condition, (c) invisible boundary condition, (d) reflective boundary condition with $V_{max} = 1$, (e) reflective boundary condition with $V_{max} = 0.3$, and (f) boundary condition with $V_{max} = 0.1$.



(a)



(b)



(c)



(d)



(e)



(f)





Fig. 5.6. Convergence curves of 50 independent runs for a linear 40-element array using the PSO with (a) invisible boundary condition with $\sigma_{initial} = 0.015$ and $\sigma_{final} = 0.00005$, (b) restricted boundary condition with $\sigma_{initial} = 0.015$ and $\sigma_{final} = 0.00005$, (c) restricted boundary condition with $\sigma_{initial} = 0.015$ and $\sigma_{final} = 0.000015$, (d) restricted boundary condition with $\sigma_{initial} = 0.015$ and $\sigma_{final} = 0.000015$, (d) restricted boundary condition with $\sigma_{initial} = 0.00005$, (e) restricted boundary condition with $\sigma_{initial} = 0.00005$, (f) restricted boundary condition with $\sigma_{initial} = 0.00005$, (f) restricted boundary condition with $\sigma_{initial} = 0.05$ and $\sigma_{final} = 0.00005$, (f) restricted boundary condition with $\sigma_{initial} = 0.05$ and $\sigma_{final} = 0.00005$, (f) restricted boundary condition with $\sigma_{initial} = 0.05$ and $\sigma_{final} = 0.00005$, (f) restricted boundary condition with $\sigma_{initial} = 0.05$ and $\sigma_{final} = 0.00005$, (f) restricted boundary condition with $\sigma_{initial} = 0.05$ and $\sigma_{final} = 0.00005$, (f) restricted boundary condition with $\sigma_{initial} = 0.05$ and $\sigma_{final} = 0.000015$, and (g) restricted boundary condition with $\sigma_{initial} = 0.05$ and $\sigma_{final} = 0.00015$.

5.4.2 Shaped Beam Synthesis

The amplitude and phase optimization of a linear 50-element array antenna to achieve the desired radiation pattern is considered. Shaping the main beam requires minimizing the absolute difference between the desired and obtained radiation pattern. Meanwhile, the "Don't Exceed" criterion is considered in the sidelobe region. Therefore, the objective is to obtain sidelobe levels less than the mask in the sidelobe regions and a main beam equal to the mask in the main beam region. The same cost function presented in section 5.4.1 is used. The beamwidth of the array pattern is 20 degrees and the number of sampling points is 719.



(a)



Fig. 5.7. Amplitude and phase synthesis for a linear 50-element array (a) Radiation pattern obtained using IWO, (b) Convergence curves for the IWO with different boundary conditions and PSO with reflective boundary condition (RBC) and different maximum velocity limit, and (c) Convergence curves for the IWO with invisible boundary condition and different number of initial and final standard deviations.

The desired and obtained radiation patterns by using the IWO are shown in Fig. 5.7 (a). The same optimization parameters, shown in Table 5.2, are used for this synthesis problem, except for the number of iterations, which is set to 2000. The convergence curves for both the IWO with different boundary conditions and the PSO with reflective boundary condition (RBC) and different maximum velocities are shown in Fig. 5.7(b). It can be seen that the IWO convergence curves for both restricted and invisible boundary conditions converge to the same level. However, In the case of the PSO, by varying the maximum velocity limit, the performance of algorithm dramatically changes. Although in some cases the PSO is faster in convergence compared to the IWO, it traps in local minima. In addition, the 50 various realizations of the same experiment for each curve of the IWO algorithm are closer to each other compared to those for the PSO. Table 5.4 shows the average numbers of fitness evaluations per successful run for both the IWO and PSO. The effect of varying the initial and final standard deviations on the convergence of the IWO with invisible boundary condition is shown in Fig. 5.7(c). It can be seen that by varying the initial SD, the convergence rate of the algorithm is improved and compete with the results obtained by PSO in the first number of iterations. However, neither the initial nor the final SD has any critical effect on the final error level. The IWO appears to be more stable since by applying different boundary conditions or different initial or final standard deviation values, the convergence speed or the level of the cost function doesn't change too much. Therefore, the IWO doesn't need much effort on tuning the parameters. Table 5.4. Comparison of average number of fitness evaluations required per successful run in

	Average number of fitness evaluations	
	required per successful run	
$PSO-RBC (V_{max} = 1)$	130023	
$PSO-RBC (V_{max} = 0.3)$	88026	
$PSO-RBC (V_{max} = 0.1)$	28695	
IWO-Restricted BC	19432	
IWO-Invisible BC	13352	

PSO and IWO algorithms for the linear shaped beam synthesis optimization

5.5 Thinned Array Antenna

In this section, thinned planar array antennas are considered as the next optimization problem to show the effectiveness and some special features of the IWO. By some modifications in the IWO, the number of elements and the position of those elements can be optimized, which results in a new scenario for developing thinned arrays. By applying this scenario, planar thinned arrays with less number of elements and higher efficiencies are obtained.

Thinned arrays, generally produced by removing certain elements from a fully populated half wavelength spaced array, are usually designed to generate low sidelobe levels. Different optimization algorithms including GA, PSO, Simulated Annealing, and Ant Colony have been applied to remove the elements in such a way to have the lowest possible sidelobe levels [79]-[86]. Although, the thinned arrays obtained by using these algorithms produce low sidelobe levels, it has been shown that by considering the aperiodic arrangements, lower sidelobe levels

can be achieved. This can be done by optimizing the inter-element spacing of periodic arrays or already thinned arrays to have lower sidelobe levels [80],[81], [87]-[90].

In this section, by some modifications in the IWO, the number of elements and at the same time their locations, the inter-element spacing, are optimized. It is shown that by using this algorithm, capable of optimizing such a problem, lower sidelobe levels with less number of elements can be achieved. Fewer elements for a given aperture mean reducing the cost and weight of the antenna system. It should be pointed out that the array is uniformly excited (all elements have identical current amplitude and phase). The advantage of uniform amplitude excitation is clear from the point of view of the feed network.

5.5.1 Modified IWO

As was mentioned in section 5.2, in the IWO, each weed (agent) may have a different number of variables during the optimization process. By taking this feature of the algorithm, different number of variables for each agent can be considered during the optimization. This modified IWO works similarly to the routine explained in section 5.2. Some modifications, however, should be made in the algorithm process to take the number of elements as an optimization parameter.

In this modified version of the IWO, each agent, corresponding to an array antenna, has a different number of elements. Thus, the fitness value of each agent is calculated based on the number of elements and the position of each element in that agent. Similar to the general IWO algorithm, each flowering plant produces new seeds based on its ranking in the colony, which means the new arrays appear in the colony. However, the reproduction process is modified to have a different number of elements for each produced array antenna. In the reproduction

process, each element in the array is removed and then reproduces some new elements in that array. The number of the new elements produced by each old element is defined to be a constant value in each iteration. Then, these new elements are dispersed over the aperture by normally distributed random numbers with mean equal to the location of the producing element and varying standard deviations. The standard deviation (SD) is defined similar to (5.1) where it starts from a large value, called the initial SD, and decreases gradually to a small value, called the final SD.

Without any limitations on the reproduced elements, the number of elements increases dramatically. Moreover, the distance between elements should be controlled not to have elements very close to each other. In order to overcome these problems, each new produced element is allowed to be located on the aperture if it is not closer than a predefined value (usually half wavelength or the size of the antenna element of the array) to any of the other elements already located on the aperture.

By choosing a relatively large value for the initial SD, the new elements are dispersed over the aperture and the possibilities of different numbers of elements over the aperture are tested. Then, by decreasing the SD to a small value, the position of each element on the aperture is optimized. Therefore, the number of elements and the location of each element are optimized.

It should be noted that in this modified version of IWO the whole process explained in section 5.2 is carried out. Meanwhile, the modified reproduction process is taken into account for each agent.

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5.5.2 Planar Thinned Array Examples

As the first design problem of thinned arrays, a rectangular planar array with the aperture of $9.5\lambda \times 4.5\lambda$ is considered. The objective is to minimize the maximum SLL in the $\varphi = 0^{\circ}$ and φ = 90° planes. This problem is selected to compare the obtained result with the results in [79] and [88]. In [79] a 20 × 10 element planar array with a half a wavelength distance between uniformly spaced elements was thinned using GA by turning off some elements in that aperture. The optimal solution is a thinned array with 108 turned on elements on the rectangular aperture [79, Fig. 7]. The optimized SLLs are equal to -20.07 dB in $\varphi = 0^{\circ}$ plane and -19.76 dB in $\varphi = 90^{\circ}$ plane [79, Fig. 9]. The fitness value of the optimal solution, defined as the sum of maximum SLLs in both planes, is -39.83 dB. The same problem is considered in [88] by optimizing the inter-element spacing between 108 elements of the obtained thinned array in [79], using a modified real GA to achieve lower SLLs. The optimal solution [88, Fig. 6] shows a lower fitness value, -45.456 dB, and SLLs equal to -29.597 dB and -15.859 dB in $\varphi = 0^{\circ}$ and $\varphi = 90^{\circ}$, respectively [88, Fig. 5].

In order to optimize this problem by using the IWO and based on the described method, the number of elements and their positions are optimized to obtain the lowest SLLs at the desired planes. The normalized array factor of a planar array with *N* elements is given by:

$$AF(\theta) = \frac{1}{AF_{\max}} \sum_{n=1}^{N} I_n \exp(j2\pi (Lx_n \sin\theta\cos\varphi + Ly_n \sin\theta\sin\varphi)/\lambda)$$
(5.4)

where Lx_n and Ly_n are the locations of elements in x and y direction, respectively. This equation assumes that the array lies in the x-y plane. Since the desired pattern is symmetric about the xaxis and y-axis, a quarter of the aperture is considered to reduce the number of optimization
parameters to the quarter of the array elements. The minimum distance between elements is assumed to be half wavelength. It should be noted that the amplitude coefficients, I_n , are assumed to be 1.

Set up of the IWO algorithm for solving this problem is summarized in table 5.5. The final SD is chosen to be a small value to optimize the location of each element with a high precision. An averaging of five runs is considered and found to be sufficient. The best thinned array obtained is presented in Table 5.6. Fig. 5.8(a) shows the radiation patterns of the obtained thinned array in both $\varphi = 0^{\circ}$ and $\varphi = 90^{\circ}$ planes. The fitness value, the sum of maximum SLLs in both planes, is -65.40 dB and the obtained SLLs are -34.72 dB in $\varphi = 0^{\circ}$ plane and -30.68 dB in $\varphi = 90^{\circ}$ plane. The array configuration of the thinned array for the upper right quarter of the aperture is depicted in Fig. 5.8(b) (Compare with [88, Fig. 6] and [79, Fig. 7]). 18 elements (72 elements for the whole aperture) are the optimized number of elements to achieve the lowest SLLs. Comparing these results with those in [79] and [88] it can be concluded that by employing this algorithm, much lower SLLs at both planes are achieved with more than 25% saving on the number of elements. Fig. 5.8 (d) shows the average and single convergence curves of the optimization as the function of the number of iterations.

Table 5.5. IWO parameters for the thinned array optimization problems

it _{max}	p _{max}	S _{max}	S _{min}	п	initial SD	final SD
500	55	5	0	3	0.25λ	0.0005λ

1(0.264,1.720)	2(0.269,0.254)	3(0.354,0.860)	4(0.593,1.314)
5(0.864,0.435)	6(1.093,2.250)	7(1.111,1.324)	8(1.413,0.283)
9(1.576,0.789)	10(1.913,0.253)	11(2.014,1.845)	12(2.249,1.331)
13(2.644,0.301)	14(2.796,0.969)	15(3.192,0.306)	16(3.568,0.742)
17(4.061,0.295)	18(4.708,0.935)		

Table 5.6. The coordinates of the array elements in wavelength: $i(\boldsymbol{x}_i,\,\boldsymbol{y}_i)$



(a)



(b)





Fig. 5.8. Optimized thinned array antenna for reduction of SLL in the $\varphi=0^{\circ}$ and $\varphi=90^{\circ}$ planes, (a) radiation patterns in the $\varphi=0^{\circ}$ and $\varphi=90^{\circ}$, (b) the array configuration of the optimized thinned array, (c) 3D radiation pattern, and (d) convergence curves for five runs and the average of them.

Another optimization problem is the reduction of SLLs in all φ planes. It can be seen from Fig. 5.8(b) that most of the elements are around the x and y axes. Such an array configuration produces high SLLs at some other planes as shown in Fig. 5.8(c). In order to reduce the SLLs in all the φ planes, we decided to define the fitness function as the maximum SLL in $\varphi = 0^{\circ}$, $\varphi = 45^{\circ}$ and $\varphi = 90^{\circ}$ planes to have more elements at the central part of the aperture. This fitness function helps to have less computation and avoid an expensive optimization process. The same optimization parameters shown in Table 5.5 are chosen for this problem. Table 5.7 represents the obtained thinned array configuration. The array radiation pattern cuts in $\varphi = 0^{\circ}$, $\varphi = 45^{\circ}$ and $\varphi = 90^{\circ}$ for the best optimal solution are shown in Fig. 5.9(a). The array configuration, shown in Fig. 5.9(b), consists of 80 elements (for the whole aperture) distributed on the aperture. It is observed that more elements are located at the central part of the aperture as it was expected. Fig. 5.9(c) is the 3D radiation pattern of this array. Though low SLLs are obtained at the three cuts shown in Fig. 5.9(a), SLLs haven't decreased effectively in the other φ planes. The average and single convergence curves of the optimization are shown in Fig 5.9 (d).

Table 5.7. The coordinates of the array elements in wavelength: $i(x_i, y_i)$

1(0.271,0.186)	2(0.289, 0.254)	3(0.292,0.763)	4(0.537,1.250)
5(0.855, 0.258)	6(0.907, 0.785)	7(1.103,2.023)	8(1.184,1.269)
9(1.530, 0.893)	10(1.672,0.257)	11(1.772,2.234)	12(2.174,0.315)
13(2.241,1.112)	14(2.669, 0.399)	15(2.944,1.433)	16(3.188,0.846)
17(3.568,1.535)	18(4.053,0.318)	19(4.701,0.421)	20(4.712,1.691)



(a)



(b)



(c)





Fig. 5.9. Optimized thinned array antenna for reduction of SLL in the $\varphi=0^{\circ}$, $\varphi=45^{\circ}$ and $\varphi=90^{\circ}$ planes, (a) radiation patterns in the $\varphi=0^{\circ}$, $\varphi=45^{\circ}$ and $\varphi=90^{\circ}$, (b) the array configuration of the optimized thinned array, (c) 3D radiation pattern, and (d) convergence curves for five runs and the average of them.

In order to have low SLLs in all planes, the fitness function is defined as the maximum SLLs in all the φ planes. The same optimization parameters are selected for this problem. The result of optimization is a thinned planar array with 92 elements (for the whole aperture) depicted in Table 5.8. Fig. 5.10(a) shows the array configuration on a quarter of the aperture. The radiation pattern of this array is shown in Fig. 5.10(b) where the maximum SLL is -21.2 dB. comparing this results to that of [88], where GA is used to minimize the SLLs for 100 elements sparse array [88, Fig. 7 and Fig. 8], one can see that the IWO results lower SLL (-18.84 dB in [88]) with less number of elements. The convergence curves for five runs and the average of them are shown in Fig. 5.10 (c).

1(0.271,2.119)	2(0.285,0.866)	3(0.385,0.369)	4(0.425,1.457)
5(0.832,1.015)	6(0.940,2.159)	7(1.106,0.316)	8(1.182,1.465)
9(1.511,0.936)	10(1.688,0.273)	11(1.878,1.475)	12(2.118,0.750)
13(2.213,0.251)	14(2.398,2.139)	15(2.770,0.266)	16(2.751,1.061)
17(3.363,0.327)	18(3.367,1.655)	19(3.690,0.979)	20(4.127,0.393)
21(4.218,1.734)	22(4.750,0.350)	23(4.750,1.007)	

Table 5.8. The coordinates of the array elements in wavelength: $i\left(x_{i},\,y_{i}\right)$



(a)



Fig. 5.10. Optimized thinned array antenna for reduction of SLL in all φ planes, (a) The array configuration of the optimized thinned array. (b) 3D radiation pattern. (c) convergence curves for five runs and the average of them.

Another idea to improve the SLLs of the antenna aperture is to remove the symmetry planes and optimize the number of elements and their location for the whole aperture. Fig. 5.11 (a) shows the obtained results where the fitness function is defined as the maximum SLLs in all the φ planes. It can be seen that a thinned array antenna with 81 elements is achieved. The 3D radiation pattern is shown in Fig. 5.11 (b) where the maximum SLL is -22.06 dB. Although the optimization process takes longer for this problem, less number of elements and lower SLLs are achieved. The convergence curves for five runs and the average of them are shown in Fig. 5.11 (c).



(a)



Fig. 5.11. Optimized thinned array antenna for reduction of SLL in all φ planes without using the symmetry planes (a) the array configuration of the optimized thinned array. (b) 3D radiation pattern. (c) convergence curves for five runs and the average of them.

5.6 Dual-Band U-Slot Patch Antenna

In this section, the design of a U-slot patch antenna [91]-[92] to have the desired dualband characteristics is considered. This concept was introduced in [93]-[94] where a U-slot in the patch fed by an L-probe produces notches within the matching band. Fig. 5.12 shows the configuration of the antenna structure. The L-probe feeding technique is used to have a wideband patch antenna [95]-[96] and then the U-slot is cut on the patch to introduce notches, resulting in dual-band operation. The length (L) and the width (W) of the patch and also the position (P_f) and height of L-probe (H_L) are predefined. Then, the optimization of seven other parameters, H, L_L , U_a , U_b , U_d , U_x and U_y is required. Since these parameters are not independent, their ranges should be chosen carefully.



Fig. 5.12. The configuration of the U-slot patch antenna fed by an L-probe with top view (top) and side view (lower).

W	L	<i>H</i> *	H_L	L_L^*	U_a*	U_b *	U_d *	U_x *	U_y *	P_f
44.5	36.4	10.572	7.0	15.941	1.203	1.756	9.333	14.525	21.799	1.0

Table 5.9. Final dimensions of the obtained U-slot patch antenna in mm

* Optimized parameter

The purpose of the optimization is to achieve the desired reflection coefficient within the matching band, 2-4 GHz. The IWO is linked to a MoM program that simulates the antenna reflection coefficient at 20 points within the frequency range. The fitness function is defined as the summation of differences between the relative values of the desired and obtained reflection coefficient at all 20 frequencies. The objective is to have a -12 dB reflection coefficient at 2.4 GHz and 3.3 GHz frequencies and zero at the other frequencies. The reason for choosing -12 dB as the desired reflection coefficient is that decreasing this value results in a very low value in one frequency but relatively high at the other one.

The restricted boundary condition is applied to this optimization problem. The maximum number of plant population is selected to be 10 and the number of seeds is varied linearly from 4 to zero. The nonlinear modulation index is set to 3 over 100 iterations.

The optimized parameters of the U-slot patch antenna are shown in Table 5.9. After simulating the antenna with more number of frequencies within the frequency range, the reflection coefficient shown in Fig. 5.13(a) is achieved. Fig 5.13(b) illustrates the convergence curve of IWO algorithm which is the normalized curve of the fitness function in dB. It can be seen that the desired reflection coefficient within the frequency range of the U-slot antenna is achieved.



Fig. 5.13. U-slot patch antenna fed by an L-probe (a) the obtained reflection coefficient, and (b) the convergence curve.

5.7 Conclusion

A numerical stochastic optimization algorithm based on the weed ecology was introduced for electromagnetic applications. The Invasive Weed Optimization algorithm is capturing the properties of the invasive weeds, which led to a powerful optimization algorithm. By applying the IWO to array antenna synthesis problems, the performance of this algorithm was investigated. It was shown that in certain instances the IWO outperforms the PSO in the convergence rate as well as the final error level. Moreover, the performance of the IWO for different boundary conditions and tuning parameters was evaluated. From the simulation results, it was observed that this algorithm is very stable and efficient against different parameter values. The IWO was also utilized to design aperiodic planar thinned array antennas by optimizing the number of elements and at the same time their positions. It was shown that by using this technique, thinned arrays with less number of elements and lower sidelobe levels, compared to the results already achieved from other methods were obtained. Also, the IWO was applied to the design of a U-slot patch antenna fed by an L-probe to have a dual band performance with the desired reflection coefficient.

Chapter 6

ULTRA-WIDEBAND TWO-DIMENSIONAL FOCUSED ARRAY ANTENNA

6.1 Introduction

Focused array antennas based on UWB impulse waveforms can provide high concentration of electromagnetic energy into small regions by controlling the timing of pulses radiated by each element [38]-[41], [97]-[99]. The performance of the focused array can be improved by increasing the number of elements or element spacing [100]-[104]. However, increasing the number of elements or increasing the spacing between elements produces a larger aperture array. Using thinned arrays, higher performance with less number of elements and the same aperture size can be obtained [98]. In addition, sparse focused array antennas are used to achieve a higher performance for this kind of antenna [104]-[105]. These antennas are of interest in medical applications, radar, homeland security systems and communication systems [3], [36]-[41], [104] [106]-[108].

In this chapter, UWB focused arrays excited by a Gaussian impulse are considered. Each impulse antenna element (LCR) is replaced by a set of infinitesimal dipoles to model the array antenna [109]-[113]. It is assumed that LCR elements are spaced at distances larger than the spatial duration of the pulse to ignore the mutual coupling as well as multiple scattering effects between elements. Different UWB focused arrays are optimized to achieve lower SLLs for the focused beam. The number of elements, polarization, and location of each element are the parameters to be optimized. The near field radiation pattern of each array is compared with a conventional uniformly spaced UWB array antenna. Simulation results show the improvements in the focused beam and sidelobe levels of the optimized array antenna.

6.2 Infinitesimal Dipole Modeling

6.2.1 Methodology

The concept of IDM was first introduced in [109] for modeling the near field of radiating structures in biomedical applications. The idea was formulated in [110]-[113] to find sets of infinitesimal dipoles by using genetic algorithm. Later, the concept was extended by applying quantum particle swarm optimization to the problem and a method of predicting the mutual coupling between antenna elements was proposed [114]. In [115]-[116] IDMs for dielectric resonators in multi layer structures obtained by using the particle swarm optimization method. In [117] a procedure of getting IDMs for wideband antennas was proposed and applied for a wideband stacked dielectric resonator antenna.

A set of infinitesimal dipoles (electric or magnetic), representing an actual antenna, can be obtained by using an optimization algorithm to minimize the difference between the near field of the dipoles and those of the actual antenna. Each dipole is represented by seven parameters: the position of the elements (x, y, z coordinates); the complex dipole moment (the real and imaginary part), and its orientation (θ and ϕ). It should be mentioned that the dipole locations are limited within the volume of the actual antenna. Although electric dipoles, magnetic dipoles, or a combination of them can be used, we use only electric dipoles to represent the antenna.

In this algorithm, the IDM can be achieved by minimizing the error of the near field data defined as

$$e = \sum_{m=1}^{M_0} \sqrt{\frac{1}{3} \sum_{n=1}^{N_o} \left| \frac{\vec{E}_a(r_n, f_m)}{P_a} - \frac{\vec{E}_d(r_n, f_m)}{P_d} \right|^2}$$
(6.1)

where N_o and M_o are the number of observation points and frequency points, respectively. $\vec{E}_a(r_n, f_m)$ and $\vec{E}_d(r_n, f_m)$ are the electric fields on the observation surfaces, obtained by a full wave solution (FEKO) [52] or measurements, and the electric dipoles, respectively. The vector r_n is the position vector of the *n*th sampling point, f_m is the frequency, and *P* is the maximum value of the electric field used for normalization as

$$P_{a,d} = Max \left\{ \vec{E}_{a,d} \left(r_n, f_m \right) \right\}.$$
(6.2)

The subscript a is used to denote the measured or accurate solution obtained for the actual problem, and the subscript d is used to denote the desired solution obtained by IDM model. It should be noted that in (6.1) the near field data is normalized. This is because the IDM is used to

obtain the spatial distribution of the near field but not the exact values depending on the excitation of the antenna. The exact field values due to the excitation of V_0 can be achieved by multiplying all the dipole moments with a factor of $V_0 P_a / P_d$ [117].

6.2.2 LCR and Its Infinitesimal Dipole Model

The LCR introduced in chapter 4 is considered. The near-field is computed on a square observation plane with side lengths of 1000 mm located at a distance 500 mm from the antenna aperture. The near field is computed at 60 frequency points from 50 MHz to 3GHz. The total number of samples is 121 points at each frequency. The modified IWO algorithm with restricted boundary condition is applied. The total number of 30 electric dipoles is obtained after the optimization. The moments, locations and orientations of dipoles are shown in Table 6.1. The electric field components of the near-field due to the dipoles compared to those obtained by Method of Moment (MoM) solution are shown in Fig. 6.1 (a). It is seen that a good agreement is achieved. The IWO convergence curve is shown in Fig. 6.1 (b).

	The dipole				The orientation		
Flamont			The posi	tion of dip	of d	ipole	
Number	11101	liciit			(degrees)		
INUITOCI	Real	Imagin	x	v	7	Α	()
	part	ary part	Λ	T		U	Ψ
1	-9.06	1.19	19.89	-19.99	-21.88	112.63	239.42
2	3.76	-0.22	5.97	-0.13	-27.95	23.97	242.16
3	-4.82	-3.65	9.82	5.32	-25.35	138.35	157.90
4	-7.98	-2.64	4.40	-11.55	-17.66	32.49	110.24
5	-9.97	-6.60	-11.38	-8.62	-10.92	174.85	259.74
6	4.97	-9.46	-7.34	-1.58	-5.10	74.82	275.18
7	7.63	0.84	6.20	7.58	-8.87	55.39	260.80
8	-9.94	1.09	18.56	17.05	-18.67	79.64	320.77
9	-1.02	-6.79	-3.62	-1.37	-10.24	65.30	320.38
10	-5.13	5.75	-19.99	6.73	-29.58	108.75	280.00
11	-6.17	-6.41	-1.58	-7.57	-17.34	170.72	290.79
12	-9.44	6.13	-19.99	-2.03	-15.49	88.84	226.77
13	1.36	7.18	12.83	8.59	-14.53	174.06	296.97
14	5.41	9.57	5.78	16.98	-29.99	43.67	273.48
15	-1.09	-2.25	-14.12	-0.16	-12.34	15.53	231.54
16	1.76	9.16	-10.26	-8.21	-10.30	122.64	3.47
17	-9.03	-3.99	-7.24	8.58	-13.78	76.00	219.89
18	-0.96	4.82	20.00	-0.60	-27.36	98.59	277.40
19	7.76	-4.99	-19.27	-13.93	-8.64	98.38	140.27
20	-0.84	-8.64	-5.78	14.34	-8.39	49.60	335.47
21	6.01	4.48	-5.29	-5.53	-29.17	99.50	245.04
22	-9.61	-7.46	0.21	-4.03	-29.99	61.59	82.82
23	0.98	-2.09	-19.13	3.45	-8.50	31.63	93.30
24	9.62	5.65	11.09	-12.51	-16.29	143.32	100.75
25	-0.38	0.99	8.06	19.72	-18.46	3.44	111.36
26	-6.38	7.93	8.59	16.93	-8.98	26.60	35.00
27	-2.05	-6.40	4.77	-4.17	-21.33	91.46	39.86
28	-6.09	-0.75	14.53	19.61	-13.39	84.00	254.96
29	-4.89	7.60	-14.74	11.01	-5.19	1.49	292.09
30	3.01	3.69	16.87	-5.11	-11.74	26.07	54.03

Table 6.1. The moments, positions and orientations of electric dipoles equivalent to LCR.



Fig. 6.1 The optimized dipoles equivalent to the LCR (a) Comparison of the electric field components of the near fields due to the actual antenna and IDMs over lines passing through the middle of the observation plane. (b) The convergence curve of the IWO.

6.2.3 Time-Domain Results

The electric and magnetic field radiated by an infinitesimal dipole in the near field region can be expressed as [53], [56]

$$\vec{E} = \frac{Z_0 l}{4\pi c} \left[\frac{1}{r} \frac{dI}{dt} \frac{\vec{r} \times (\vec{r} \times \vec{l})}{l \cdot r^2} + \left(\frac{c}{r^2} I + \frac{c^2}{r^3} \int I dt \right) \left(\frac{3(\vec{l} \cdot \vec{r})\vec{r}}{l \cdot r^2} - \frac{\vec{l}}{l} \right) \right]$$

$$\vec{H} = \frac{l}{4\pi c} \left(\frac{1}{r} \frac{dI}{dt} + \frac{c}{r^2} I \right) \frac{(\vec{l} \times \vec{r})}{(l \cdot r)}$$
(6.3)

where *I* is the radiator current, *l* is the dipole length , Z_0 is the wave impedance of the free space, *c* is the velocity of light and *r* is the radius vector to the observation point.

In order to show the efficiency and accuracy of infinitesimal dipole modeling, the same time-domain analysis presented in section 4.2 is applied to the obtained infinitesimal dipole set. In other words, the infinitesimal dipole set is excited by the same Gaussian pulse (σ = 2e-10). The radiated near field patterns of the LCR obtained by using IDMs and MoM solution at a time instance of t = 1.2/c and a distance of r = 1.2 m from the antenna aperture are shown in Fig. 6.2. It can be seen that a good agreement is achieved. Therefore, by replacing the antenna by IDM, a very fast and efficient method for modeling the antenna is achieved. In other words, instead of solving the LCR by using a full wave method, it is replaced by a set of dipoles.



(a)



(b)

Fig. 6.2. Comparison between the near field patterns of an LCR obtained using the MoM and the IDM excited by a Gaussian pulse at a time instance of t = 1.2/c and a distance of r = 1.2 m from the antenna aperture in the (a) E-plane and (b) H-plane.

6.3 Uniformly Spaced Focused Arrays

6.3.1 2 by 2 LCR Array

In order to evaluate the validity and efficiency of this method, a 2 × 2 LCR array is designed and modeled using the IDM method. This array antenna, having the same dimensions and element distances of the presented LCR antenna in section 4.3, is solved by a full wave MoM software, and IDM method. The same Gaussian pulse presented in the previous section is used to excite each LCR element. Since the distance between the LCRs is much higher than the spatial duration of the radiated field pulse, the mutual coupling between the elements is ignored. A comparison between the time-domain near field patterns of the array at a time instance of t =1.2/c at the distance of r = 1.2 m from the antenna, obtained by MoM and IDM is shown in Fig. 6.3. It can be seen that a good agreement is achieved.



(a)



(b)

Fig. 6.3. Comparison between the near field patterns of the 2 by 2 antenna array obtained using the MoM and the IDM at a time instance of t = 1.2/c at the distance of r = 1.2 m from the antenna, in the (a) E-plane and (b) H-plane.

6.3.2 4 by 4 LCR Array

A two-dimensional square array composed of 16 (4 by 4) elements is considered. The array elements are uniformly spaced in the xz-plane with separation of d = 0.3 m between elements (Fig. 6.4). Each LCR element replaced by dipoles is excited by the same Gaussian pulse. Therefore, there is an equally weighted amplitude distribution for the array. The array antenna can focus the radiated pulse by setting a proper delay time at each element. Therefore, the radiating waves arrive simultaneously and add up in phase at the desired focusing point. For a focusing distance *F* from the antenna aperture the time delay of i-th element should be set up as

$$t_i = \frac{\sqrt{F^2 + L_i^2}}{c} \tag{6.4}$$

where c is the speed of light and L_i is the distance of the i-th element form the center of the square aperture. The radiation near field pattern of the array at the focal distance from the aperture at time *t* can be obtained by the summation of the radiated field of each element. Fig. 6.5 shows the three dimensional normalized focused beam pattern at the time instance of t = F/c and the focal plane (F = 1.2 m) of the antenna array. The two dimensional near field pattern at the time instance of t = F/c and the focal plane, *F*, is shown in Fig 6.6. The near field pattern versus the axial length of the antenna array at the same time instance is illustrated in Fig. 6.7. It seems that the maximum intensity occurs at the focal point of the antenna. However, observing the field variations at different time instances, we can see that a higher intensity value occurs at a point closer to the antenna aperture compared to the focal point (Fig. 6.8). The bold line in Fig.

6.8 shows the field at the time instance t = 1.2/c. The variation of peak power values for different time instances are shown in Fig. 6.9.



Fig. 6.4. Geometry of uniformly spaced two-dimensional array antenna.



Fig. 6.5. 3D near field pattern of the 4 by 4 uniformly spaced array antenna at the time instance of t = F/c and the focal plane (F = 1.2 m).



Fig. 6.6. 2D near field pattern of the 4 by 4 uniformly spaced array antenna at the time instance of t = F/c and the focal plane (F = 1.2 m).



Fig. 6.7. The normalized field distribution of the 4 by 4 uniformly spaced array versus the axial distance at the time instance of t = F/c.



Fig. 6.8. The normalized field distribution of the 4 by 4 uniformly spaced array versus the axial distance at different time instances. The bold line shows the field at the time instance t = F/c.



Fig. 6.9. The variation of peak power values for different time instances.

6.3.3 8 by 8 LCR Array

In this section, a larger array antenna composed of 64 (8 by 8) elements is considered. The array elements are separated with d = 0.3 m between elements and a Gaussian pulse excites each element. The antenna is designed to have a focal length of 2.4 m. Fig. 6.10 shows the three-dimensional normalized focused beam pattern at the focal plane of the antenna array. The two dimensional field pattern of the antenna at the focal plane is shown in Fig 6.11. Fig. 6.12 shows the variation of the near field pattern versus the axial length of the antenna at one time instance. Calculating the near field pattern versus the axial length at different time instances, one can see that the maximum intensity occurs at y = 2.35 m from the antenna aperture (Fig. 6.13). The bold line in this figure shows the field at the time instance t = F/c. The variation of the peak power for different instances illustrated in Fig. 6.14, shows the focal shift in the field pattern of the array antenna.



Fig. 6.10. Three-dimensional normalized power pattern of the 8 by 8 array at the time instance t = F/c and focal plane (y = 2.4 m).



Fig. 6.11. Normalized power pattern of the 8 by 8 array antenna at the time instance t = F/c and focal plane (y = 2.4 m).



Fig. 6.12. Variation of the normalized power pattern of the 8 by 8 array antenna with F = 2.4 m versus the axial distance at the time instance of t = F/c.



Fig. 6.13. the variation of the near field pattern versus the axial length at different time instances for the 8 by 8 array with F = 2.4m. The bold line shows the field at the time instance t = F/c.



Fig. 6.14. The variation of the peak power at different time instances for the 8 by 8 array with F = 2.4 m at the focal point.

In the next example, the same 8 by 8 array antenna is considered with a focal length of 4.8 m. Fig. 6.15 shows the variation of the normalized power versus the axial distance at one instance. The variation of the power distribution versus the axial distance for different time instances is shown in Fig. 6.16. The bold line shows the field at the time instance t = F/c. It can be seen that the maximum intensity occurs at y = 4.48 m from the antenna aperture. The variation of the power versus the different time instances for the 8 by 8 array is shown in Fig. 6.17. Table 6.2 depicts the variations of the maximum intensity point for different focal lengths of the 8 by 8 array antenna. It can be seen that similar to narrow band antennas, for the focal point close to the antenna aperture, the maximum intensity is very close to the focal point, but as the focal point moves away from the antenna aperture, the focal shift increases.



Fig. 6.15. The variation of the normalized power pattern of the 8 by 8 array antenna with F = 4.8 m versus the axial distance.



Fig. 6.16. the variation of the near field pattern versus the axial length at different time instances for the 8 by 8 array with F = 4.8 m. The bold line shows the field at the time instance t = F/c.



Fig. 6.17. The variation of the peak power versus the different time instances for the 8 by 8 array with F = 4.8 m at the focal point.

Focal length (m)	1.2	2.4	3.6	4.8	6	7.2	8.4	9.6
Maximum Intensity length (m)	1.18	2.35	3.44	4.48	5.39	6.21	6.96	7.6
Peak Power Difference (dB)	0.005	0.074	0.17	0.30	0.46	0.60	0.86	1.04

Table 6.2. Focusing properties of the 8 by 8 array at different focal lengths

6.4 Array Antenna Optimization

6.4.1 Polarization Optimization of the Array

The invasive weed optimization is employed to optimize the polarization of each of the LCR elements to obtain the minimum SLL for the near field pattern of the 4 by 4 array antenna. The polarization of each LCR can be changed by rotating each element around its axis. The objective function is to minimize the maximum side lobe level of the near field pattern at the focal plane. The convergence curve of the optimization process is shown in Fig. 6.18. Fig. 6.19 shows the 3D radiation pattern of the antenna at the focal plane. The comparison between the near field pattern of the optimized array and the conventional one in the x plane is shown in Fig. 6.20. Similar near field patterns are achieved in the other plane. It can be seen that both the sidelobe levels and half power beam width are decreased. The maximum sidelobe level is decreased from -12.26 to -14.43 dB after the optimization. The angle of rotation for the upper right quarter of the array is shown in Table 6.3.



Fig. 6.18. The convergence curve for the polarization optimization of the 4 by 4 array.



Fig 6.19. 3D radiation pattern of the dual-polarized optimized antenna at the time instance of t = F/c and the focal plane.


Fig. 6.20. Comparison between the near field pattern of the optimized array and the conventional one versus the x axis at the time instance of t = F/c and the focal plane.

Table 6.3	The angle of	f rotation for	the upper i	right quarter (of the optimized	4×4 arrav
1 4010 0.5.	The ungle of	10tution 101	the upper i	15 ¹¹¹ quarter	or the optimized	i array.

	Element # 1	Element # 2	Element # 3	Element # 4
Location (x,z)	(0.45,0.45)	(0.15,0.45)	(0.45,0.15)	(0.15,0.15)
(m)				
Angle of				
rotation	110.4	159.8	158.8	179.7
(Degrees)				

In order to show the efficiency of the polarization optimization, an 8 by 8 array antenna is considered. In this example, the focal length is 2.4 m and the spacing between elements is 0.3 m. The 3D radiation pattern of the optimized array is shown in Fig. 6.21. A comparison of the normalized near field pattern of the single and dual polarized (optimized) antenna in both the x and z planes is shown in Fig. 6.22. The maximum SLL is decreased from -19.45 dB for the conventional design to -22.67 dB for the optimized array.



Fig. 6.21. 3D radiation pattern of the 8 by 8 dual-polarized optimized antenna at the time instance of t = F/c and the focal plane F = 2.4 m.



Fig. 6.22. A comparison of the normalized near field pattern of the single and dual polarized (optimized) antenna at the time instance of the t = F/c in the (a) x-plane and (b) z-planes of the focal plane.

6.4.2 Thinned array Optimization

In this section, the modified invasive weed optimization method is employed to obtain an ultra-wideband thinned focused array. The same aperture dimension of the 8 by 8 uniformly spaced array described in section 6.4.1 is considered. The number of elements, the location and the rotational angle of each element is optimized to obtain the minimum SLL for the near field focused pattern at the focal point of the antenna. It should be noticed that a maximum distance of 0.3 m between elements is assigned during the optimization. Set up of the IWO for the optimization is summarized in Table 6.4. Fig. 6.23 shows the 3D near field pattern of the optimized thinned array antenna at the focal plane (F = 2.4 m). The near field pattern of the thinned array is compared with those of the single and dual polarized uniformly spaced arrays in Fig. 6.24. The maximum SLL of -24.53 dB is achieved for the thinned array. The array configuration of the thinned array for the upper right quarter of the aperture is depicted in Fig. 6.25. The locations and angles of rotation of elements for the upper right quarter of the array are shown in Table 6.5. 32 elements are the optimized number of elements to achieve the lowest SLLs. It can be concluded that lower SLLs and lower half power beam width are achieved with 50% saving on the number of elements. Fig. 6.26 shows the convergence curve of the optimization.

 it_{max} P_{max} S_{max} S_{min} ninitial SDfinal SD1000905030.10.000005

Table 6.4. Set up of the IWO for the thinned array optimization.



Fig. 6.23. 3D near field pattern of the thinned array antenna at the time instance of t = F/c and the focal plane.



Fig. 6.24. A comparison between the near field pattern of the thinned array and single and dual polarized uniformly spaced array at the time instance of t = F/c and in the (a) x- plane and (b) zplane of the focal plane.

-0.5 -0.25 0 0.25 0. Z Transverse Direction (m)

(b)

0.5

0.75

1

1.25

-0.75

-1.25

-1



Fig. 6.25. The upper right quarter of the thinned array.



Fig. 6.26. The convergence curve of the thinned array optimization.

Element	Location (x , z)	Angle of rotation
number	(m)	(Degrees)
1	(0.558, 0.759)	4.82
2	(0.999, 0.943)	11.18
3	(0.273 , 1.034)	8.82
4	(1.042, 0.283)	13.03
5	(0.715 , 1.049)	0.34
6	(0.150 , 0.749)	13.82
7	(1.0401, 0.642)	19.91
8	(0.764, 0.152)	0

Table 6.5. The locations and angles of rotation of elements in the thinned array.

6.5 Conclusion

Some focusing properties of the ultra-wideband time-domain focused array antennas were presented. By replacing each LCR element with a set of dipoles a fast and accurate simulation tool was achieved. Calculating the variation of the power distributions versus the axial length at different time instances showed the focal shift effect in the near-field pattern. It was shown that the maximum intensity occurred closer to the focal point and this displacement was increased as the focal point moved away from the antenna aperture.

To improve the focusing characteristics of the array, the invasive weed optimization algorithm is employed to optimize ultra-wideband focused array antennas. The rotational angle of each UWB element for a uniformly spaced array was optimized to achieve lower sidelobe levels. The number of elements, the location and rotational angle of each element were optimized as the next optimization problem to achieve a thinned array. It was shown that lower SLLs are achieved with 50% saving of the elements.

Chapter 7

CONCLUDING REMARKS AND FUTURE WORK

The theory and implementation of different antennas for focusing applications were presented. Large microstrip patch arrays, Fresnel zone plate lens antennas and ultra-wideband time-domain focused array antennas were designed and modeled to focus the microwave power at a point close to the antenna aperture. All introduced antennas were modeled by full wave solutions to have an accurate prediction of the field variation near to the antenna aperture. It was shown that the maximum intensity of the electric field along the axial direction was displaced from the focal point towards the aperture for all the narrowband antennas. This displacement decreases as the aperture size increases. A very similar behavior was observed in UWB pulse antennas. Moreover, a new optimization algorithm was implemented and applied to the ultrawideband focused array to increase the peak power delivered to a localized region.

The contribution of this work can be summarized in three main points. First, a new concept in designing large array antennas to focus the microwave power in the radiation near-field region was presented. A small focused array antenna using microstrip patch elements to

achieve the desired sidelobe levels in the Fresnel region based on Dolph-Chebyshev design was implemented. Larger arrays were designed by using the knowledge of the mutual admittances between the elements of smaller arrays. Second, some new focusing properties of Fresnel zone plate (FZP) lens antennas in the near-field region were presented. In addition, the focused beam scanning of the FZP lens antennas in the radiation near-field was examined. Finally, some new focusing characteristics of ultra-wideband time-domain array antenna were investigated. In addition, a new ultra-wideband thinned focused array antenna using a new strategy in designing thinned arrays was implemented. It was shown that lower SLLs with less number of elements in the array were achieved.

In addition to the contributions made related to the focused antennas, a new numerical stochastic optimization algorithm was proposed for electromagnetic applications. This algorithm, Invasive Weed Optimization (IWO) was compared to the particle swarm optimization (PSO) algorithm by applying both algorithms to the linear array antenna synthesis, the standard problem used by antenna engineers. It was shown that in certain instances the IWO outperforms the PSO in the convergence rate as well as the final error level. Moreover, The IWO was utilized to design aperiodic planar thinned array antennas by optimizing the number of elements and at the same times their positions. By implementing this new scenario, thinned arrays with less number of elements and lower sidelobes, compared to the results achieved by genetic algorithm (GA) for the same aperture dimensions, were obtained.

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