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Optimization of Reflection Coefficient of Large Log-Periodic Array by Computing only a Small Part of It

Jian Yang, *Senior Member, IEEE*, and Per-Simon Kildal, *Fellow, IEEE*

***Abstract*—A new method is described in the paper to predict the reflection coefficient of a complete large log-periodic array (with large number of elements) from the computed performance of a small part of the array. The small part is referred to as the partial array. The method involves computing the embedded S-parameters between all the elements of this partial array and then using these S-parameters to predict the performance of the full array. As an example, the method has been applied to optimize the Eleven antenna for 2 - 13 GHz. The result of the optimization is that the reflection coefficient is improved significantly to be below -10 dB over the whole frequency band. The results have been verified by measurements.**

***Index Terms*—Antenna feed, Eleven antenna, log-periodic array antenna, optimization, genetic algorithm.**

I. Introduction

MANY applications require wideband or ultra-wideband antennas, such as in radio astronomy and for ultra-wideband (UWB) communication [1][2]. The radio astronomy applications are mainly the future SKA (SKA = Square Kilometer Array) and the VLBI2010 (VLBI = Very Long Baseline Interferometry) projects. Both these two projects have reflector antennas as candidates for the frequency range of approximately 2 - 13 GHz, whereas the UWB communication systems require small direct radiating antennas covering typically 3 - 10 GHz. Thus the frequency ranges are similar, although the requirements are quite different.

One way to achieve antennas with one or more decade bandwidth is to use log-periodic arrays, being introduced 50 years ago [3]-[5]. Some recent developments of them are described in [6]-[7]. There exist also leaky wave antennas offering similar wideband performance even in the millimeter wave range [8]. We have during the last years been developing a new decade bandwidth log-periodic dual-dipole antenna referred to as the Eleven antenna. It was originally presented in [9], and two hardware realizations of it operating below 2 GHz were described in [10] and [11]. The first successful attempt to make a high frequency model working up to 10 GHz is described in [12], but it suffered from mechanical problem. The Eleven antenna is a multiport antenna, and this can be utilized for generating tracking beams [13]. A combination of an Eleven antenna and a high frequency horn for a dual band feed is described in [14].

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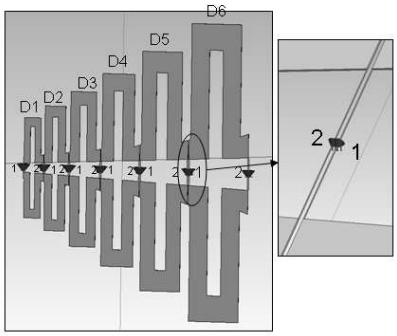
Fig. 1. Example of log-periodic array antenna.

The basic geometry of the Eleven antenna is two parallel folded halfwave dipoles separated by half of its resonant wavelength and located above a ground plane. Then, this basic configuration is scaled log-periodically to a number of folded dipole pairs, and these folded dipoles are cascaded one after another to two log-periodic dipole petals located opposite to each other.

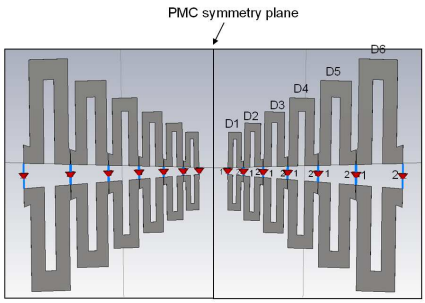
The Eleven antenna has very good features: nearly constant beamwidth with about 11 dBi directivity, fixed phase center location over the whole bandwidth, low profile and simple geometry. Therefore, the Eleven antenna is a good candidate for wideband feeding of the reflector antennas. In the paper, the Eleven antenna is referred to as the Eleven feed when it is used as a feed for reflector antennas.

For low noise systems like SKA and VLBI2010, a critical performance parameter is the input reflection coefficient of the feed, because a strong reflection at the input port of the feed will cause noise mismatch to the LNA and increase the system noise temperature. Therefore, it is very important to minimizing the input reflection coefficient of the Eleven feed. Analytical or quasi-analytical analysis for the Eleven antenna is only available for one-element-pair Eleven antenna [15]. Due to the desired nearly-decade frequency band, the size of the whole Eleven antenna becomes very large in terms of wavelengths at the highest frequency, which makes numerical optimization a very challenging problem because of the large computation time.

In this paper, we introduce a new method that can predict the reflection coefficient of a complete large log-periodic array (with large number of elements) from the S parameters of a small part of the array. This small array is referred to as the partial array. For normal large array antennas, for example, phased array antennas, the concept of predicting



(a)



(b)

Fig. 2. Port definition of cascaded log-periodic array antenna. (a) for a general case; (b) for the Eleven antenna case.

the performance of the whole array from the parameters of a small partial array has been used and shown as a very efficient method [16]. The main contribution of the paper is that we make use of the periodicity of the performance of a cascaded log-periodic array and then provide formulas by which we can establish the performance of the whole array by using the S-parameters of a small partial array. As an example, the method is applied to minimize the input reflection coefficient of a 14-element-pair Eleven feed over 2-13 GHz by using the Genetic Algorithm [17]. The result of the optimization is verified by measurements on an experimental model of the full 14-element-pair array.

II. Theory

The basic principle of the present partial array method can be explained as follows. A log-periodic array is built up by cascading log-periodically scaled elements (scaling factor k) with input ports $1(idipole)$ and output ports $2(idipole)$ one after another, where $idipole$ is the dipole number in the array, see Fig. 1 and Fig. 2(a). In the Eleven antenna case, see Fig. 2(b), there are two identical but opposite located log-periodic arrays. In other words, the two arrays are symmetrical with respect to the symmetry plane of the arrays. Since the excitations on the two symmetrical arrays are identical, we can set the symmetry plane as PMC (Perfect Magnetic Conductor) plane and analyze only one array with the port definition shown in Fig. 2(b). We assume that in a general log-periodic array the first element counted from the input port of the array radiates at the highest frequency, and the last at the lowest frequency of the band.

If the log-periodic array is large, the S-parameters are frequency scaled due to the log-periodicity, except for the edge elements, according to

$$S_{i(l+n)j(m+n)}(f) = S_{i(l)j(m)}(f/k^n) \tag{1}$$

where $S_{i(l)j(m)}$ is the S parameter between the port i of dipole l and port j of dipole m, f the frequency, k the scaling factor for the log-periodic array and n an integer. The port definitions for the example array can be seen in Fig. 2, where D1 represents dipole 1 and so on. A 6-element-pair Eleven antenna has been modeled in CST MS (time-domain finite integration method solver) [18] with two discrete ports on each dipole and with all dipoles located in their correct log-periodic positions, as shown in Fig. 2. The reference impedance of the ports (also called as the balanced or differential ports of folded dipoles) is 200 Ohms, which is the desired input impedance of the array. Fig. 3 shows simulated $S_{1(3)1(3)}(f)$, $S_{1(4)1(4)}(f)$ and scaled $S_{1(3)1(3)}(f/k)$ for the 6-element-pair Eleven antenna. It can be observed that $S_{1(4)1(4)}(f) \approx S_{1(3)1(3)}(f/k)$, which indicates that there is a frequency scaling between the S parameters of the two centrally located dipoles. It should be noted that the output port of one dipole and the input port of the next are separated by a narrow air gap which is not existing when the dipoles are cascaded together but has to be there for modeling the ports in CST. The gap is also modeled periodically and the maximum gap for designing the Eleven feed of 2-13 GHz is 0.26 mm which corresponds to a phase change of 4 degree at 13 GHz.

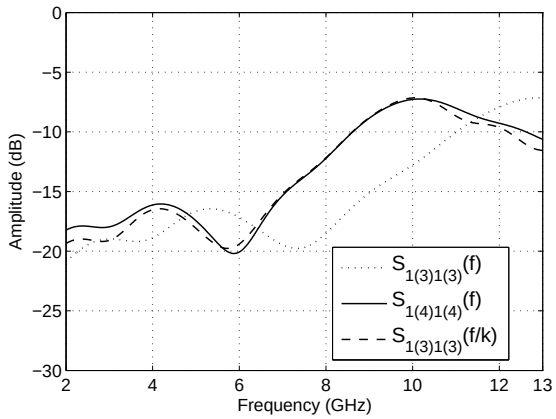
Another important characteristics of the example log-periodic array is that the mutual coupling between two ports separated far apart from each other is much smaller than those between neighboring ports. This can be seen from the curves in Fig. 4 evaluated between ports of the example antenna in Fig. 2. The mutual coupling from element 1 to element 6 is about 20 dB lower than that from element 3 to element 4. This strong reduction of mutual coupling between far-separated elements is certainly a requirement for the proposed partial array method to work, and it is also envisioned that all log-periodic arrays have this property to work properly over very wide frequency bandwidth.

Thus, by using the frequency scaling of S parameters and neglecting mutual couplings between far separated elements, we can obtain the approximate S matrix of the whole log-periodic array by using computed results of a much smaller partial array. Let us now first introduce the following port numbering convention of the whole array: port 1 is the input port of dipole 1, port 2 the output port of dipole N (the last dipole), port 3 the output port of dipole 1, port 4 the input port of dipole 2, and so on. Then, we can represent the log-periodic array with N elements (or N-element pairs for the Eleven antenna) before cascading the elements by the following S matrix of size $2N \times 2N$:

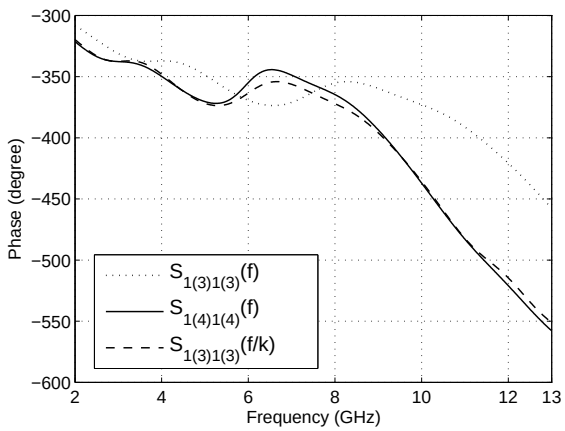
$$\begin{bmatrix} \mathbf{b}_I \\ \mathbf{b}_{II} \end{bmatrix} = \begin{bmatrix} \mathbf{S}_{I,I} & \mathbf{S}_{I,II} \\ \mathbf{S}_{II,I} & \mathbf{S}_{II,II} \end{bmatrix} \begin{bmatrix} \mathbf{a}_I \\ \mathbf{a}_{II} \end{bmatrix} \tag{2}$$

where

$$\mathbf{b}_I = \begin{bmatrix} b_{1(1)} \\ b_{2(N)} \end{bmatrix}, \quad \mathbf{a}_I = \begin{bmatrix} a_{1(1)} \\ a_{2(N)} \end{bmatrix}$$



(a) Amplitude



(b) Phase

Fig. 3. Illustration of frequency scaling of S-parameters by a 6-element Eleven antenna simulated with all mutual couplings included.

$$\mathbf{b}_{II} = \begin{bmatrix} b_{2(1)} \\ b_{1(2)} \\ \vdots \\ b_{1(N)} \end{bmatrix}, \quad \mathbf{a}_{II} = \begin{bmatrix} a_{2(1)} \\ a_{1(2)} \\ \vdots \\ a_{1(N)} \end{bmatrix}$$

$$\mathbf{S}_{I,I} = \begin{bmatrix} S_{1(1)1(1)} & S_{1(1)2(N)} \\ S_{2(N)1(1)} & S_{2(N)2(N)} \end{bmatrix}$$

$$\mathbf{S}_{II,I} = \begin{bmatrix} S_{2(1)1(1)} & S_{2(1)2(N)} \\ S_{1(2)1(1)} & S_{1(2)2(N)} \\ \vdots & \vdots \\ S_{1(N)1(1)} & S_{1(N)2(N)} \end{bmatrix}$$

$$\mathbf{S}_{I,II} = \begin{bmatrix} S_{1(1)2(1)} & S_{1(1)1(2)} & \cdots & S_{1(1)1(N)} \\ S_{2(N)2(1)} & S_{2(N)1(2)} & \cdots & S_{2(N)1(N)} \end{bmatrix}$$

$$\mathbf{S}_{II,II} = \begin{bmatrix} S_{2(1)2(1)} & S_{2(1)1(2)} & \cdots & S_{2(1)1(N)} \\ S_{1(2)2(1)} & S_{1(2)1(2)} & \cdots & S_{1(2)1(N)} \\ \vdots & \vdots & \ddots & \vdots \\ S_{1(N)2(1)} & S_{1(N)1(2)} & \cdots & S_{2(N)2(N)} \end{bmatrix}$$

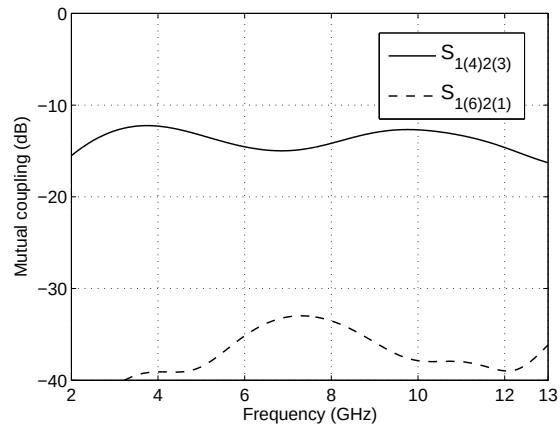


Fig. 4. Mutual coupling between neighbored elements and between far-separated elements in a 6-element Eleven antenna.

By this port numbering convention, we can model the cascading of all elements one after another as:

$$b_{2(l)} = a_{1(l+1)}, b_{1(l+1)} = a_{2(l)} \quad (l = 1, \cdots, N-1)$$

i.e.

$$\mathbf{b}_{II} = \mathbf{D} \cdot \mathbf{a}_{II} \tag{3}$$

where

$$\mathbf{D} = \begin{bmatrix} 0 & 1 & 0 & 0 & \cdots & 0 & 0 \\ 1 & 0 & 0 & 0 & \cdots & 0 & 0 \\ 0 & 0 & 0 & 1 & \cdots & 0 & 0 \\ 0 & 0 & 1 & 0 & \cdots & 0 & 0 \\ \vdots & \vdots & \vdots & \vdots & \ddots & \vdots & \\ 0 & 0 & 0 & 0 & \cdots & 0 & 1 \\ 0 & 0 & 0 & 0 & \cdots & 1 & 0 \end{bmatrix}$$

Substituting (3) into (2), we can obtain the final S matrix of the two port network of the whole log-periodic array after all dipoles have been cascaded as:

$$\mathbf{S} = \mathbf{S}_{I,I} + \mathbf{S}_{I,II}(\mathbf{D} - \mathbf{S}_{II,II})^{-1}\mathbf{S}_{II,I} \tag{4}$$

The solution procedure is then as follows. We calculate the S parameters of the small partial array, for example, the innermost M-element partial array. Then, by using frequency scaling, we can get all S parameters between all the elements of the complete large array that are separated by up to $M-2$ elements, for example,

$$S_{i(l)j(l+M-1)}(f) = S^{PA}_{i(1)j(M)}(f/k^{l-1}), \ (1 \le l \le N-M+1)$$

where S^{PA} is the S parameter of the partial array. For the non-edge elements, we always use the S-parameters for the centrally located elements in the partial array, for example,

$$S_{i(l)j(l+1)}(f) = S^{PA}_{i([M/2])j([M/2]+1)}(f/k^{l-[M/2]}).$$

where $[M/2]$ is the greatest integer less than or equal to $M/2$. For the edge elements, we use the S-parameters of the edge elements in the partial array, for example,

$$S_{i(1)j(2)}(f) = S^{PA}_{i(1)j(2)}(f)$$

$$S_{i(N-1)j(N)}(f) = S^{PA}_{i(M-1)j(M)}(f/k^{N-M}).$$

We also choose to neglect the S parameters between the elements separated by more than $M-2$ elements, so that

$$S_{i(l)j(l+m)}(f) = 0,\ if\ |m| \geq M.$$

Using these approximations we now know all the elements of the array $\mathbf{S}_{I,I}$, $\mathbf{S}_{I,II}$, $\mathbf{S}_{II,I}$ and $\mathbf{S}_{II,II}$, and then the reflection coefficient of the whole large array after cascading can be obtained from the **S** matrix of the two-port network in (4) by defining the termination at the end of the array. For example, we can short-circuit the last port of the log-periodic array as shown in Fig. 1. This is known to work well for our case from our previous results.

One thing is worth to point out: here we use a S matrix of size $2N \times 2N$ to describe the whole log-periodic array before the cascading of all elements and then use the above formulas to obtain the S matrix of size 2×2 for the array after all elements cascaded one after another, instead of using ABCD matrix. The reason for this is that due to the mutual coupling among radiation elements in the array, we can not use the multiplication of all the radiation element's ABCD matrix to get the total ABCD matrix of the whole array after the cascading of all elements one after another.

The criterion for determining the required number of log-periodic elements in the partial array depends on how fast the mutual coupling (S parameter) decays as a function of the number of separating elements. We have not used any exact criterion for deciding the size of the partial array , but instead we have made a decision after testing different sizes of the partial arrays and comparing the result from the partial array method with that from full wave simulation of the whole large array. In this work the partial array was chosen to have 6 elements. Fig. 5 shows the input reflection coefficients of a complete 14-element-pair log-periodic Eleven antenna array, both from simulation using CST on the complete array with a model having 12.8 million meshes and by using the partial array method with the simulated performance of the innermost 6-element partial array with a model having 2.5 million meshes. One can observe that the present method provides a quite accurate result down to -10 dB. The computation time using the present method is about 30 minutes while it is about 2 hours to simulate the whole array in CST. It is quite obvious that the larger the complete array is, the more beneficial the partial array method becomes.

It should be noted that when cascading all elements, we have taken the phase shift due to the small gap between the elements into account. φ_l is the phase shift due to the gap $l_{g,l}$ between dipole *l* and dipole *l+1* in the partial array, and S^{gap} is the S parameter simulated with ports defined with gaps in the partial array. Therefore, the S parameters of the partial array should be corrected as

$$S^{PA}_{i(l)j(m)} = S^{gap}_{i(l)j(m)} \cdot e^{-\mathbf{j}[(i-1)\varphi_l + (j-1)\varphi_m]} \quad (5)$$

where $\varphi_l = 2\pi \cdot l_{g,l}/\lambda$ and the bolded **j** is the sign of an imaginary number.

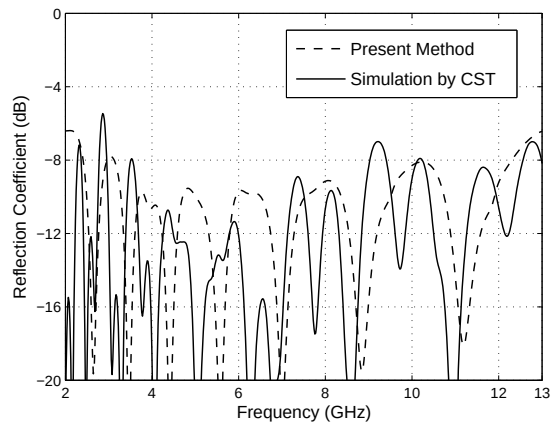


Fig. 5. Example reflection coefficient of a 14-element Eleven antenna by using the present partial array method (dashed) with 6-element partial array and a full wave simulation of the whole 14-element array by CST (solid).

III. Modeling Eleven Antenna in CST

The basic configuration of the Eleven antenna is two parallel folded dipoles separated by half of their resonant wavelength and located above a ground plane. Then, the pair of folded dipoles are extended with log-periodic scaling factor k to a number of pairs, and these folded dipoles are cascaded one after another in each side of the pair, as shown in Fig. 6. Each log-periodic part (or side) of the Eleven antenna is referred to as a petal. A linearly polarized Eleven antenna (Fig. 6) has two petals and a dual polarized Eleven antenna (Fig. 8) has four petals. Consequently, all the results in Fig. 3, Fig. 4 and Fig. 5 have been computed with the presence of the two opposite petals and the ground plane. The Eleven antenna is fed at the differential input ports of the first innermost folded dipole pair with the same amplitude and phase, see port 1 and port 2 in Fig. 6. The origin of the coordinate system for the Eleven antenna is located in the center of the ground plane as shown in Fig. 6 as well. This is already known to be the phase center location [9]. The geometry of the Eleven antenna has two symmetry planes: *x-z* and *y-z* planes. Due to the excitation, the *x-z* plane can be regarded as PEC-type symmetry plane and the *y-z* plane PMC-type symmetry plane (PEC = Perfect Electric Conducting, PMC = Perfect Magnetic Conducting). Therefore, we only need to simulate a quarter of the geometry to obtain the performance of the whole feed. The first innermost pair of folded dipoles have dipole lengths of approximately half wavelength at a frequency which is referred to as the highest geometrical frequency f_{geomax}. The next dipole pair has halfwave lengths at a geometrical frequency f_{geomax}/k, and so on. The last folded dipole pair has then the geometrical frequency f_{geomax}/k^N, where N is the total number of folded dipole pairs in the array. By using a geometrical frequency f_{geo} for each folded dipole pair, the parameters describing the geometry can be expressed in terms of the geometrical wavelength $\lambda_{geo} = c/f_{geo}$, which simplifies the parameter description. The outermost folded dipole pair is short-circuited at its outer port, which is known to work well.

The geometry of the Eleven feed is determined by the

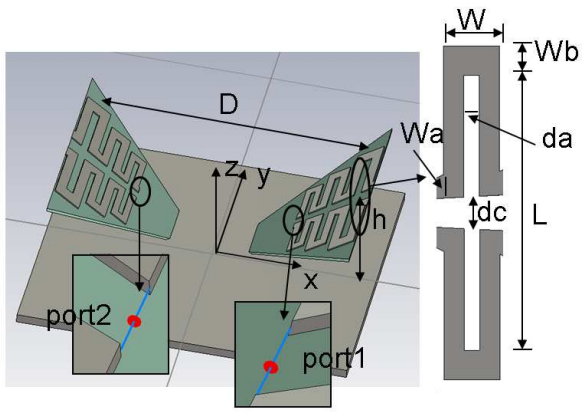


Fig. 6. Modeling of Eleven feed in CST.

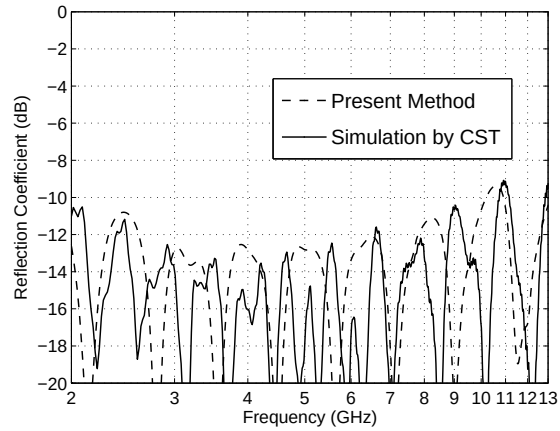


Fig. 7. Reflection coefficient of the optimized 14-element linearly polarized Eleven antenna by using the present partial array method (dashed) and a full wave simulation of the whole array by CST (solid).

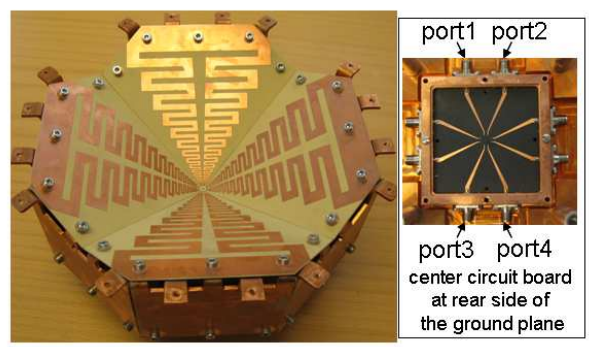


Fig. 8. Hardware of the dual polarized Eleven feed for 2 - 13 GHz. The octagon of metal dewar has a circumradius of 200 mm.

following parameters, as also shown in Fig. 6: the spacing D between the folded dipoles in a pair, the length L of the folded dipole, the height h of folded dipole over the ground plane, the center gap d_c, the folded dipole widths w, w_a and w_b, the arm slot width d_a and the scaling factor k. All dimension parameters are in term of geometrical wavelength and varying log-periodically, with two exception: the thicknesses $t_{metal} = 0.070mm$ of the copper strips which build up each petal with folded dipoles, and the thickness $t_{die} = 0.381mm$ of the dielectric layer that is used to mechanically support these metal strips. The above-mentioned thicknesses of the dielectric and metal are included in the simulations, both for the partial array and the complete array.

IV. Genetic Algorithm Optimization

The goal of the optimization is to minimize the reflection coefficient of the Eleven feed over 2 - 13 GHz under the condition of that the radiation pattern gives a high aperture efficiency when feeding a paraboloid. Therefore, the spacing D between the corresponding folded dipoles in a pair was fixed to $0.5\lambda_{geo}$, which is known to give a good radiation pattern for all heights over the ground plane smaller than $0.5\lambda_{geo}$ [19]. There were also additional mechanical constraints regarding manufacturability on the dimensions of the lines of the innermost dipoles, and that the feed should be cryogenically coolabe. Therefore, it is chosen that the antenna petals are realized on a printed circuit board with Rogers TMM3 substrate ($\epsilon_r = 3.27$ and $tan\delta = 0.002$ at 10 GHz) which has the same thermal expansion coefficient as copper. The present partial array method was applied only to the linearly polarized configuration of the Eleven feed with two antenna petals shown in Fig. 6. The copper and dielectric were included in the simulations, as discussed at the end of Sec. III. Genetic Algorithm [17] was used for the optimization, with the same procedure that was successfully used in [20]. The genes can be expressed as

$$gene = \{k, L, w, w_a, w_b, d_c, d_a, h\}.$$

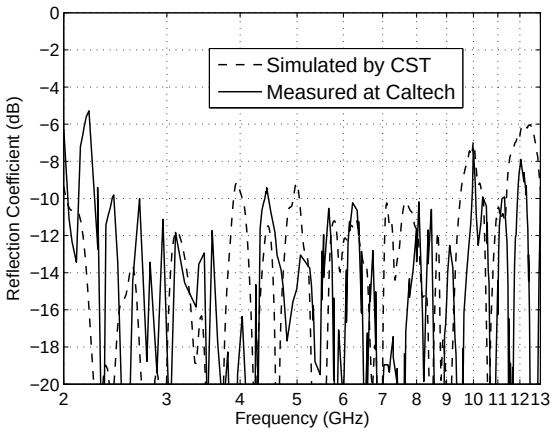
Floating point vector is used to express the chromosome, for example, as

$$\{1.19, 0.176, 0.0365, 0.0229, 0.015, 0.025, 0.0107, 0.172\}$$

The first generation was created randomly with a population of 50 individuals. Then, an elite group of 10 individuals with the best genes was selected and two-point crossover is applied among the group for creating genes for next generation. For the remaining non-elite population, roulette-wheel selection is used to choose the genes, and crossover is used to create the genes for the rest population for the next generation. All genes are checked to see if they satisfy the manufacturability constrains. If the population of the next generation is fewer than 50 due to the mechanical constraints, mutation and random creation are used to create more genes. The whole procedure runs 5 generations to achieve convergence.

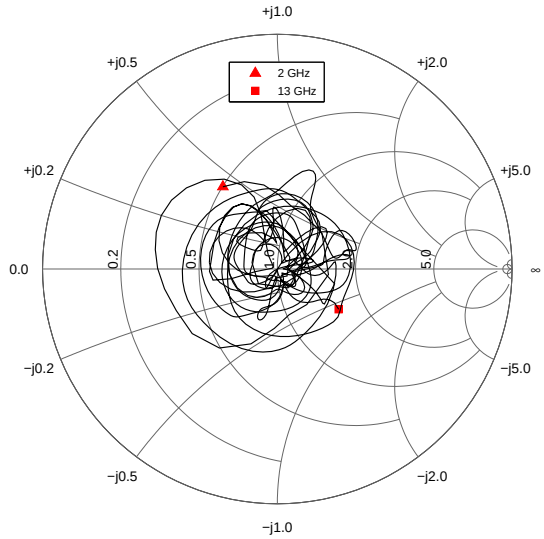
The final minimized reflection coefficient of the whole linearly polarized Eleven feed is shown in Fig. 7, by both the present partial array approach and using CST on the whole optimized array, which is below -9.4 dB over the whole frequency band.

(a) Measurement setup at Caltech

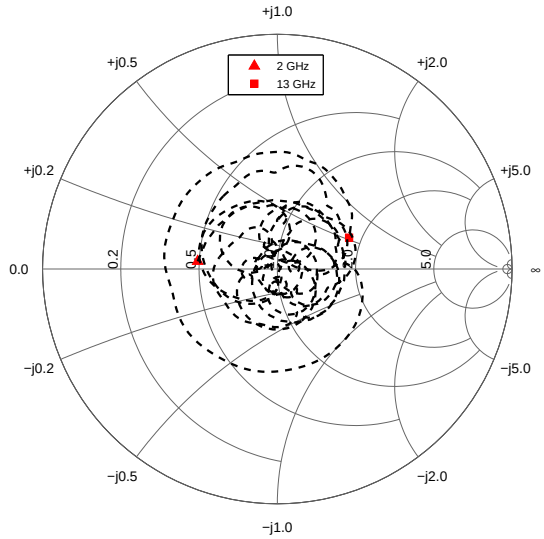


(b) Simulated and measured data in dB

Fig. 9. Simulated and measured reflection coefficient of the dual polarized Eleven antenna.



(a) Measured S_{11}



(b) Simulated S_{11}

Fig. 10. Simulated and measured reflection coefficient in Smith Chart.

V. Experimental Verification

The final results of the optimization were verified against measurements on an experimental model of the Eleven antenna containing the full geometry of 4 petals for dual polarization. The dimensions of each of the 4 petals are identical to one optimized for linear polarization. A feeding network for this Eleven antenna has been designed, which consists of four two-wire lines. these are realized as pairs of two steel wires. Each pair of two wires pass vertically through a hole in the ground plane, forming a two-wire transmission line with characteristic impedance of 200 Ohms, and are connected to a coupled microstrip line on a circuit board at the rear side of the ground plane. The coupled microstrip line has initially differential characteristic impedance (odd-mode characteristic impedance) of 200 Ohms and then is transformed to two normal microstrip lines with characteristic impedance of 50 Ohms. The detailed design of this is out of the scope of this paper, please refer to [21]. Fig. 8 shows the photo of the hardware of the dual polarized Eleven antenna which includes 14 folded dipoles per petal. It also shows the center circuit board at the rear side of the ground plane. Each polarization is excited by 4 coaxial ports (SMA connectors with transitions to the microstrip lines), shown as ports 1 - 4 in Fig. 8. Ports 1 - 4 have port impedance of 50 Ohms and are excited with the same amplitude and phase of 0°, 180°, 0° and 180°, respectively, in order to excite one polarization. This is done in practice by using wideband power dividers and 180 degree hybrids. The measurements were done with calibrations at the four coaxial ports for one polarization with the other four ports for the orthogonal polarization terminated with 50 Ohm loads. Therefore, the effects of the power dividers and hybrids are removed by the calibration in the presented experimental results. Fig. 9 shows the final results for the input reflection coefficient $S_{11,total}$, both simulated by using CST and measured over 2 to 13 GHz. The measurement was done at California Institute of Technology (Caltech) by measuring the full 4-port S-parameters of one polarization of the Eleven antenna (see Fig. 9) and then, the reflection

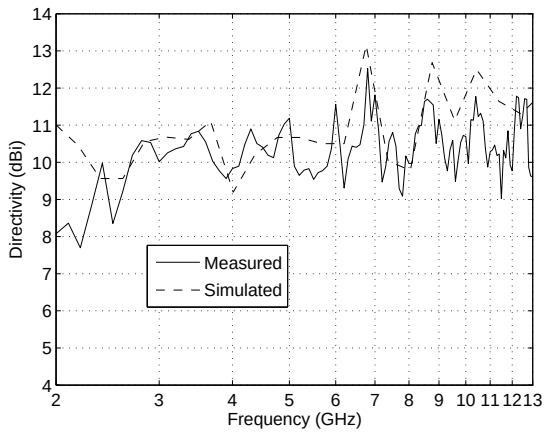


Fig. 11. Simulated and measured directivity of the dual polarized Eleven feed.

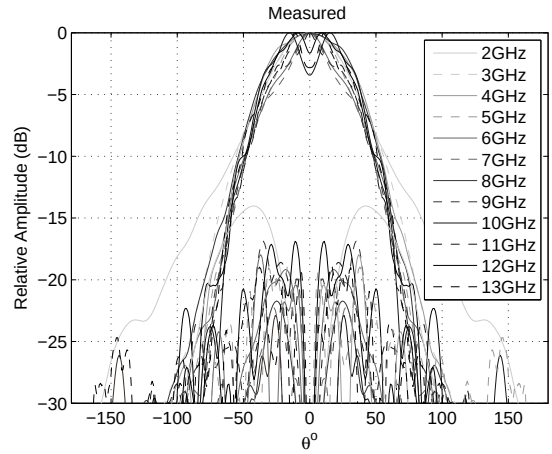


Fig. 12. Measured radiation pattern of BOR1 component.

coefficient $S_{11,total}$ is calculated by using [13]

$$S_{11,total} = S_{11} - S_{12} + S_{13} - S_{14}. \quad (6)$$

Fig. 10 shows the Smith Chart presentation of the measured and simulated $S_{11,total}$. It can be observed that the measured reflection coefficient is below -10 dB at most points from 2.2 to 10 GHz and below -8 dB from 10 to 13 GHz. This is an improvement of 2 dB below 10 GHz and 5 dB from 10 to 13 GHz compared to the Eleven feed published in [12].

It is also needed to quantify the quality of the radiation performance of the optimized Eleven antenna. It is known from [22] that the configuration of the Eleven antenna satisfies the condition for constant radiation characteristics for log-periodic array antennas. Therefore, the optimized Eleven antenna should retain the good radiation performance. Fig. 11 shows the simulated and measured directivity over 2 - 13 GHz, and Fig. 12 shows the co- and cross-polar radiation patterns of the BOR_1 components in the $\varphi = 45°$ plane [23], both measured at DTU (Technical University of Denmark). It can be observed that the beam width of the BOR_1 component is nearly constant over the whole band, which indicates a good radiation performance. For more radiation performance data of this Eleven feed, both the simulated and the measured, please refer to [21].

VI. Conclusion

A new method that can predict input reflection coefficient of a large log-periodic array (with large number of elements) from a small partial array has been presented. The method transforms the computed performance of a small partial array to that of the whole array via S-parameters between the embedded cascaded elements, which is very computationally efficient. The Genetic Algorithm has thereafter been applied for optimization of the performance of the Eleven antenna for 2 to 13 GHz. The reflection coefficient is below -10 dB over 2.2 to 10 GHz and below -8 dB from 10 to 13 GHz, which is an improvement of 2 dB below 10 GHz and 5 dB from 10 to 13 GHz compared to the previous model. At the same time, this Eleven antenna retains the good radiation performance.

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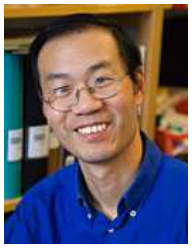

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