A Method of Electronic Beam Steering for Circular Switched Parasitic Dipole Arrays

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Abstract- This paper presents a method of electronic beam steering for a smart antenna system based on a switched parasitic array. The antenna system is made up of circular array of half wave dipoles, with four parasitic elements positioned equidistantly, surrounding one active central element. All parasitic elements are assumed to be switched between short-circuit and open-circuit states to change their resonant length. The switching between parasitic elements' states is attained by connecting a switch that assumes *ON/OFF* states at the centre of each parasitic element. This allows some parasitic elements to act as reflectors while short-circuited, whereas other parasitic elements act as directors when open-circuited. Thus, the direction of maximum gain is controlled by open- and short-circuiting the parasitic elements. By appropriately selecting correct combination of short- and open-circuited parasitic elements together with the active element, a set of radiation patterns is formed covering the horizontal plane. Results for the horizontal coverage are presented with the main beam(s) directed at 0^0, 90^0, 180^0, and/or 270^0, as well as for an omnidirectional configuration resemblance. It is expected that increasing the number of parasitic elements will improve directional properties further.

Index Terms—Switched parasitic arrays; Electronic beam steering; Moment methods; Beams.

I. INTRODUCTION

The demand for wireless communication is growing explosively making the existing communication systems inadequate. All impairments of the wireless communication systems need to be addressed as the world is moving in to the future wireless communication. In focusing on improving the performance of wireless communication, access technologies need to be taken into consideration [1]. Amongst several access technologies, radio technologies that include multiple antennas are of great interest [1]. Furthermore, whilst not overlooking the factors like cost, complexity, power consumption and practicality, parasitic arrays is the type of antenna arrays to consider for future wireless communications [2-7].

Electronically steerable parasitic arrays have been studied and discussed in the literature [3-7]. Beam steering using parasitic arrays could be achieved over both planes: in the elevation and azimuth [4]. In the literature, dipoles and monopoles have been mostly considered as the parasitic array elements [4-7]. Switched parasitic arrays could assume different array configurations, for instance, linear, planar and circular arrays [8].

However, circular arrays are considered because of the advantage of azimuthal symmetry, and hence radiation pattern can be varied in both azimuth and elevation planes. Circular switched parasitic dipole arrays with one driven element and all parasitic elements uniformly encircling the active/driven element have been studied in [4-7,9].

Parasitic elements are mounted in a close proximity to the driven element; and therefore radiation field of the active element induces currents in the parasitic elements, which causes the parasitic elements to radiate in turn [10]. This interaction between antenna elements is due to mutual coupling [10]. Mutual coupling is a concept fundamental for the functioning and operation of the parasitic antenna arrays. Parasitic arrays and the concept of mutual coupling are dealt with in depth in [8,11,12]. The most prevalent method for analyzing wire antennas and arrays with effects of mutual coupling is the Method of Moments (MoM) [13].

In the design of parasitic array antennas, it is required that one optimizes both the structural and control parameters [14]. In this paper, it is assumed that the structural parameters of the antenna are fixed, and only the control parameters will be considered.

The control parameters depend on the type of antenna loading used. There are two general parasitic array antennas, namely reactive and switched parasitic antennas [15]. The two parasitic antenna types differ in antenna loading circuitry, whereby switched parasitic antennas uses few state circuitry while reactive antennas use loading circuitry with more states [15].

In this paper a method of electronic beam steering for switched parasitic circular arrays using an idealized radio frequency (RF) switch with two states, *ON/OFF* is presented. This method offers minimal control circuit complexity, low power consumption and lower manufacturing cost for parasitic antennas [9,15].

II. CONTROL FORMULATION

A. Theory

Circular array symmetry may be used in the design of switched parasitic dipole arrays. The most obvious geometry is a simple circular array of N elements, with one active element and ($P=N-1$) equally spaced parasitic element surrounding the active element. The elements are wire dipoles parallel to the z-axis. Their centres are located in a circular manner along the x–y plane and have the same length $\boldsymbol{l}$ and radius $\boldsymbol{a}$ [7]. Fig.1 illustrates the array used for the presented beam steering scheme using parasitic elements.

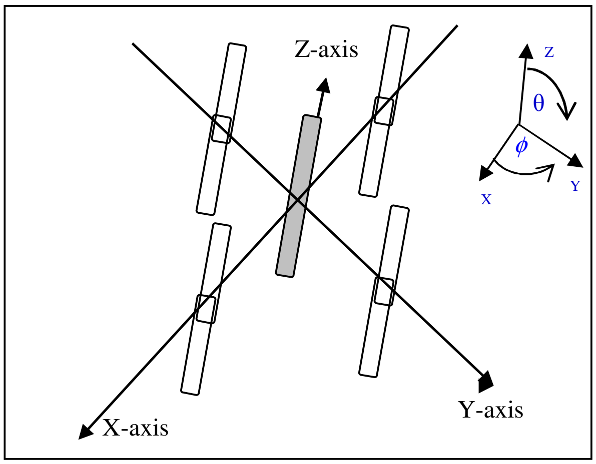

Fig. 1: Antenna geometry of 5 elements array, with one active element encircled by 4 parasitic elements.

The studied antenna consists of a total of 5 elements (N=5): one active element and 4 parasitic elements (P=4). Each of the parasitic elements has a switch connected at the centre. The switch is assumed to have only two states, being *ON* and *OFF*. When the switch is in *ON* position, the corresponding parasitic element is short-circuited and when the switch is *OFF*, the parasitic element is in the open-circuit state. All elements are physically of the same electric length, switching their states between short-and open-circuited results in a variation of the elements' electrical length.

Variation of parasitic elements' electrical length would enable them to be switched between functioning as reflector and director since the resonant length is altered [16-18]. Reflector is the array element that is physically slightly longer than the feed element, while a radiator is the array element that is physically slightly shorter than the feed element in the theory of Yagi-Uda antennas [10,19].

The electric far field of a conversional (all active elements) dipole-based circular array is represented by [19],

$$E_t(\theta,\phi,r) = EF.\underbrace{\left[\sum_{n=1}^{N} I_n e^{j(a\sin\theta\cos(\phi-\phi_n)+\alpha_n)}\right]}_{\text{Array factor}}, \quad (1)$$

where EF represents the element factor. The n^{th} parasitic element encircling the active element at the origin is located at radius $\boldsymbol{a}$ with the phase angle ϕ_n ($n=1,2,\ldots,P$); I_n and α_n are the current amplitude and phase excitation of the n^{th} element of the array. All variables in the expression retain their usual meaning as defined by [19].

However, the array that gives this expression does not consider mutual coupling between the array elements [7, 10], which is critical in parasitic arrays [20]. The array factor is mainly dominated by the dipole currents which depend on the impedance matrix [10,19,21]. The impedance matrix is also dependent on the load of the parasitic elements and the voltage source driving the active element [7].

Equation (1) needs to be modified to be useful in the case of parasitic arrays. The current amplitude I_n and phase α_n of excitation should be replaced by the actual complex current I'_n. The new expression for the electric far field incorporating the effects of mutual interactions is represented by

$$E_t(\theta,\phi,r) = EF.\left[I'_0 + \sum_{n=1}^{N-1} I'_n e^{j(a\sin\theta\cos(\phi-\phi_n))}\right], \quad (2)$$

where I'_0 is the complex current of the driven element encircled by the parasitic elements. I'_n is the complex current of the n^{th} parasitic element with all other variables holding their meaning as in (1). This will enable the array factor to include the effects of mutual coupling amongst the array elements [7].

The equivalent network corresponding to Fig. 1 can be determined and the network equations are given by [19, 21]:

$$\begin{bmatrix} V_1 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} Z_{11} & Z_{12} & Z_{13} & Z_{14} & Z_{15} \\ Z_{21} & Z_{22} & Z_{23} & Z_{24} & Z_{25} \\ Z_{31} & Z_{32} & Z_{33} & Z_{34} & Z_{35} \\ Z_{41} & Z_{42} & Z_{43} & Z_{44} & Z_{45} \\ Z_{51} & Z_{52} & Z_{53} & Z_{54} & Z_{55} \end{bmatrix} \begin{bmatrix} I'_1 \\ I'_2 \\ I'_3 \\ I'_4 \\ I'_5 \end{bmatrix}, \quad (3)$$

where V_1 is the voltage source at port 1(active element). Z_{nn} denotes self impedance of the nth {for n=1,2,..,5} element in the array. Whereas, Z_{nm} denotes mutual impedance of nth element due to mth element {for n & m=1,2,...,5}. I'_n is the complex current along the nth element when the mutual coupling amogst the emelemts has been taken into consideration.

Equation (3) shows an ideal situation where there is no load across the terminals of all parasitic elements and the terminals are short-circuited (voltage source equals zero). Assuming that there is a load connected across the terminals of each parasitic element, there will be a voltage drop across the element terminal. This voltage drop is due to a current flow across the terminal of the parasitic element [10]. Since there are no voltage sources driving the terminals of the passive elements, the voltage drop across passive elements terminals may be given by [8, 10]:

$$\begin{aligned} V_2 &= -Z_L I'_2 \\ V_3 &= -Z_L I'_3 \\ V_4 &= -Z_L I'_4 \\ V_5 &= -Z_L I'_5 \end{aligned} \quad . \tag{4}$$

Substituting (4) into (3), the entries along the main diagonal of (3) can be denoted as,

$$Z_{n_L} = Z_{nn} + Z_L \text{ , for n = \{2,3,4,5\}.} \tag{5}$$

Replacing the diagonal entries in (3) with (5), then a new form of (3) is,

$$\begin{bmatrix} V_1 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} Z_{11} & Z_{12} & Z_{13} & Z_{14} & Z_{15} \\ Z_{21} & Z_{2_L} & Z_{23} & Z_{24} & Z_{25} \\ Z_{31} & Z_{32} & Z_{3_L} & Z_{34} & Z_{35} \\ Z_{41} & Z_{42} & Z_{43} & Z_{4_L} & Z_{45} \\ Z_{51} & Z_{52} & Z_{53} & Z_{54} & Z_{5_L} \end{bmatrix} \begin{bmatrix} I'_1 \\ I'_2 \\ I'_3 \\ I'_4 \\ I'_5 \end{bmatrix} . \tag{6}$$

Expression (6) can be written in a shorter form using matrix notation as

$$\begin{aligned} &\mathbf{V} = [\mathbf{Z} + \mathbf{Z_L}]\mathbf{I}' \\ &\text{or} \\ &\mathbf{I}' = [\mathbf{Z} + \mathbf{Z_L}]^{-1}\mathbf{V} \end{aligned} \quad , \tag{7}$$

where $\mathbf{Z_L}$ is the diagonal matrix containing load impedances across each passive element terminals. **V** is the source voltage matrix; **Z** is the impedance matrix as in (3), and **I'** is the current vector for all currents in the elements [20, 21]. $\mathbf{Z_L}$ is the load matrix, represented as

$$\mathbf{Z_L} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 \\ 0 & Z_{2_L} & 0 & 0 & 0 \\ 0 & 0 & Z_{3_L} & 0 & 0 \\ 0 & 0 & 0 & Z_{4_L} & 0 \\ 0 & 0 & 0 & 0 & Z_{5_L} \end{bmatrix} . \tag{8}$$

Change of any of the diagonal values in (8) results in the change of the impedance matrix. Computation of the impedance matrix is vital since the current vector depends on it as seen in (7). Thus, the impedance matrix influences the radiation characteristics of the antenna. It is analytically possible to calculate the impedance matrix for simpler structures, though it can be more challenging for complex structures [10, 21].

It would be preferable to compute the impedance matrix either numerically or experimentally (measuring the impedance in order to determine the current flowing in each array element) for complex structures [19]. Method of moments (MoM) is one of the numerical methods that could be used to compute the impedance matrix as well as the current distribution for each element in the network [19, 21].

B. Simulation Method

The simulation method presented in this paper is based on Matlab. The method uses Personal Computer Aided Antenna Design (PCAAD), a wire antenna analysis and design software for computing the impedance matrix based on moment methods [22]. Only the initial impedance matrix in (3) is computed using PCAAD. From (6), the impedance matrix can be altered by changing and adding the load matrix (8) to (6).

Each time the load matrix is altered, the impedance matrix is altered; hence all radiation and impedance characteristics of the antenna need to be recomputed. The number of possible combinations of the loads on all parasitic elements is given by the total number of combinations that can be attained while changing the state of one parasitic element at a time. The number of possible combinations is given by the number of states a load can assume to the power of the number of parasitic elements (*P*);

$$Total\_Combinations = States^P \tag{9}$$

In this paper, two state RF switch (*ON*/*OFF*) is assumed; thus, while changing the load state of only one parasitic element at a time, there are

$$16 = 2^4 \quad , \tag{10}$$

combinations. Not all combinations result in useful results, hence manual filtration of the results is done based on the following parameters: gain and input impedance values. Only directions having reasonable parameter values are considered; better (higher) gain and input impedance comparable to the transmission line impedance for impedance match.

III. RESULTS

The initial impedance matrix is altered by short-circuiting and open-circuiting the parasitic element. The four parasitic elements are located along the main axis (*x-y axis*, in the azimuth plane) oriented along the *z-axis* at a radius of 0.5λ from the centre as in Fig. 1. The diameter of all elements is assumed to be 0.001λ and the frequency of operation is 3 GHz.

Structural parameters in this paper are assumed based on the most frequently used structural parameters for circular array of half wave dipoles [8,10,19]. The frequency band used has been chosen to be just higher than the 2.4 GHz band. Use of any frequency band whilst maintaining same ratio and size of the structural parameters (in terms of wavelength) will yield same results.

Fig.2 illustrates ability of circular array symmetry to aid in resembling omnidirectional radiation along the horizontal plane. This radiation pattern results when all four parasitic elements are open-circuited. The method used in this paper indicates that the azimuth antenna gain is very low, of about 2.4 dB in the Omni configuration setup.

Directionality of the antenna is attained when open-circuiting one parasitic element at a time as in shown in Fig.3 and Fig.4. Azimuth gain in these configurations is more than 4.8 dB. Table1 summarizes beam directions, input impedances of the antenna as well as the horizontal gain given by the array factor.

Different combinations result in different beam steering resolution. However, with the type of loading used in this case, some parameters like gain are reduced for smaller steering resolution while still maintaining the same number of array elements. Thus, a beam steering of 90 degree resolution is considered since it gives acceptable results with much higher gain. With 90 degree resolution, the relative pattern along the horizontal plane is symmetric having the same pattern, beamwidth and gain, only direction on main beam differs.

Relative patterns indicate beam steering as well as null formations in other undesired directions. All plots are symmetrical with each plot/figure indicating different desired directions. Only 2-Dimentional relative patterns are presented to indicate beam steering ability along the horizontal plane. Thus plots are viewed as *θ-Cut* at $\theta=90^0$. The geometry used in this paper is the commonly used as in Fig.1; where angle ϕ is measure along the horizontal plane (*x-y* axis), and angle θ is measured from *z*-axis to the ground plane (*x-y* axis).

Table 1: Summery of available directions of the main beam with corresponding gain, input impedance as well as states of the parasitic elements.

Angle of main beam	Set of Parasitic Element State: Short-circuit= "S" Open-circuit="O"				Maximum Gain in Azimuth plane	Input Impedance (Ω)
Omni	O	O	O	O	2.4 dB	82.54+45.7i
0 deg	O	S	S	S	4.83 dB	68.56-2.2i
90 deg	S	O	S	S	4.83 dB	68.56-2.2i
180 deg	S	S	O	S	4.83 dB	68.56-2.2i
270 deg	S	S	S	O	4.83 dB	68.56-2.2i

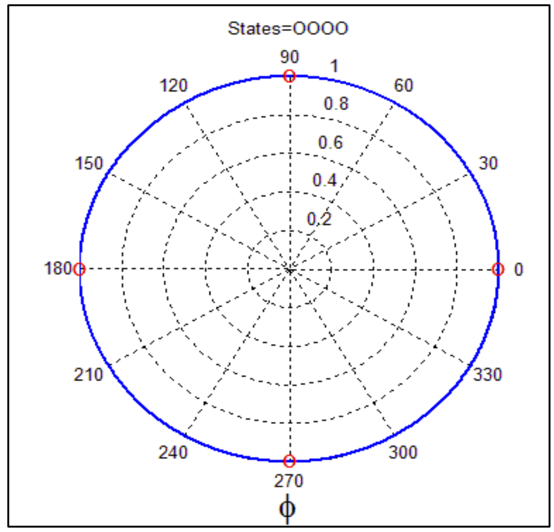

Fig. 2: Omnidirectional configuration resulting when all 4 parasitic elements are open-circuited.

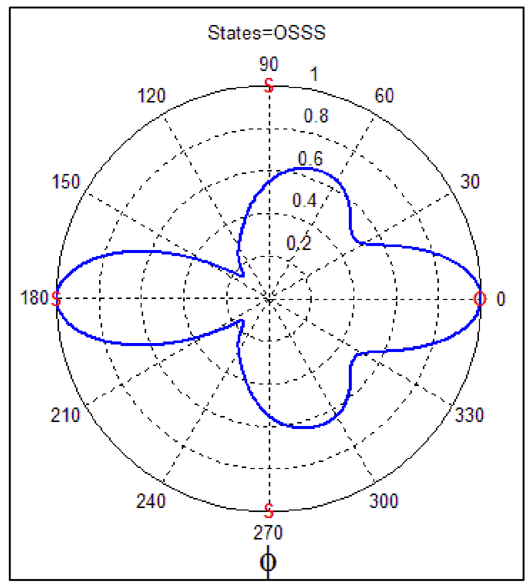

Fig. 3: Main beam direction at 0 deg given by, open-circuiting the load on the positive x-axis, with all other parasitic elements short-circuited.

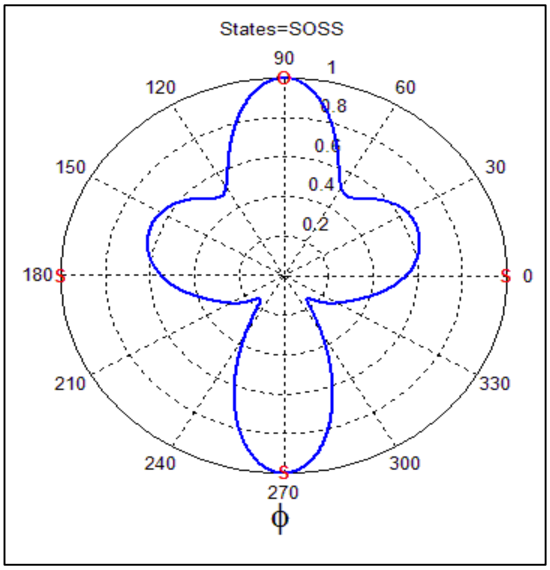

Fig. 4: Main beam direction at 90 deg given by, open-circuiting on the positive y-axis, with all other parasitic elements short-circuited.

IV. Conclusion and Future Work

A. Conclusion

A method of electronic steering of the beam for the parasitic arrays has been achieved. By selecting a correct combination of the parasitic elements' states either short or open-circuit; thereby varying the load impedances of the parasitic elements, nulls and main beams can be placed at desired directions. The main aim of the method is to steer the direction of the main beam in the azimuth plane.

The method presented is using antenna array of 4 parasitic elements encircling an active element covering symmetrically the *x-y* plane with 4 radiation patterns. Hence, the symmetry of the radiation pattern can be exploited for use in symmetrical formations for the array. An omnidirectional configuration is also achieved which makes the antenna array to be have both directional and omnidirectional characteristics. This method is applicable for switched beam smart antenna configuration since all direction where the main beam and nulls can be pointed have already been computed.

B. Future Work

The array configuration used in this paper has less number of parasitic elements. Part of the future work is to increase the number of parasitic elements in other to have high beam steering resolution as well as minimal magnitude of side lobes and achieving higher gain. Ultimately, practical testing and measurements of all parameters considered in the simulation method of this paper will be performed.

V. Reference

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