Communication

Efficient Method of Moments for Numerical Analysis of Antennas with Variable Load Impedance

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***Abstract*—A computationally efficient method which is based on method of moments (MoM) for numerical analysis of antennas with variable load impedance is proposed. The proposed method deals with both of the variable load impedance and its current as equivalent voltage source. Current distribution of antennas with the variable load impedance is obtained directly from their unloaded full-admittance matrix and the block impedance matrix corresponding to the variable load impedance. The proposed method is quite computationally efficient because inversion of the full-impedance matrix of the antennas is unnecessary when their load impedance varies. Moreover, the proposed method does not include any approximation and its result shows perfect agreement with that of the full-wave analysis. Numerical results of the proposed method are compared with those of conventional MoM using our in-house code and its performance is demonstrated.**

***Index Terms*—Method of moments, Load impedance, Antennas, Array Antennas.**

I. Introduction

Method of moments (MoM) is well-known as one of the powerful techniques for numerical analysis of antennas or scatterers [1]. The MoM has been applied to numerical analysis of wire antennas[2], [3], planar antennas [4] and dielectric materials[5]. The MoM is computationally efficient because unknown currents to be obtained are only distributed on antennas or scatterers themselves while unknown electric/magnetic fields to be obtained are distributed on entire region including antennas and scatterers for finite element method (FEM) or finite difference time domain (FDTD) method.

Although the MoM is computationally efficient, sophisticated techniques are necessary for numerical analysis of large-scale problems. One of the well-known techniques is an iterative solver such as fast multipole method (FMM) [6], [7], adaptive integral method (AIM) [8], and conjugate-gradient fast fourier transform (CG-FFT) [9]. The iterative solvers update unknown currents iteratively and large-scale problems can be solved without resorting conventional direct solvers. Another technique is so-called macro basis function (MBF) method, such as characteristic basis function method (CBFM) [10], [11] or sub-entire domain (SED) basis functions [12], [13]. The MBF method enables to reduce the size of the original large-scale problems and reduced problems can be solved using conventional direct solvers.

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On the other hand, reconfigurable antennas are widely used because of flexibility in their performance such as directivity, frequency band, and so on. A large reflectarray using single bit phase shifters has been proposed for millimeter-wave imaging systems [14]. A p-i-n diode is loaded with a microstrip patch element and working as an RF switch. Owing to the p-i-n diode, beam scanning capability is available for designed reflectarrays. During design of the reflectarrays, the p-i-n diode is modeled as a series circuit of resistance, inductance and capacitance. Microstrip antennas with tunable reactance devices have been proposed [15]. It has been shown that wide beam scanning angle is available owing to tunable reactance devices. As mentioned here, some of reconfigurable antennas are designed using variable load impedance and their performance must be simulated every time when the load impedance varies. It is well-known that the MoM is useful for numerical analysis of reconfigurable antennas with variable load impedance but a long CPU time is unavoidable for numerical simulation because full-size problems must be solved iteratively even when the sophisticated MoMs such as FMM or CBFM are used. Therefore, how to deal with variable load impedance efficiently during their design using the MoM is quite important in practice.

In this paper, a computationally efficient method for numerical analysis of antennas with variable load impedance is proposed. The proposed method requires inversion of full-impedance matrix without load impedance only one time. Small block matrix equations corresponding to load impedance are solved every time when the load impedance varies. The proposed method is computationally efficient because inversion of full-impedance matrix is unnecessary when the load impedance varies. Small block matrix equations are solved sequentially using the the proposed method without any approximation. As a result, current distribution obtained using the proposed method is accurate.

This paper expands the work presented briefly as a technical report without peer-review [16]. The technical report [16] has just demonstrated capability of the proposed method roughly via numerical simulation of simple dipole/loop antennas. Rigorous formulation of the proposed method reflecting all different port conditions, i.e. port with voltage source, load impedance, and both of them, is unavailable in the technical report. As a result, computational cost of the proposed method has not been discussed rigorously in the technical report. In addition, performance of the proposed method for practical antennas is unavailable in the technical report. On the other hand, this paper gives rigorous formulation of the proposed method

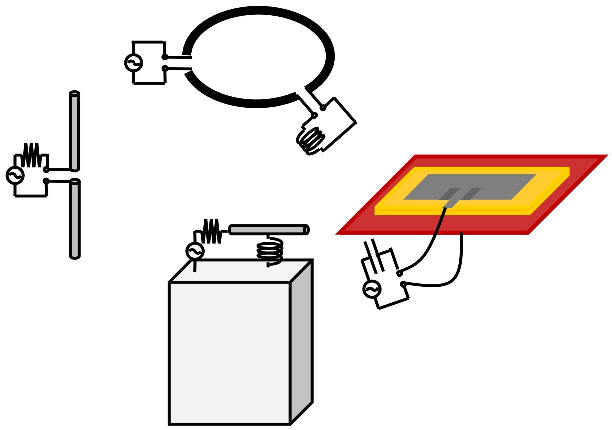

Fig. 1. Antennas with load impedances.

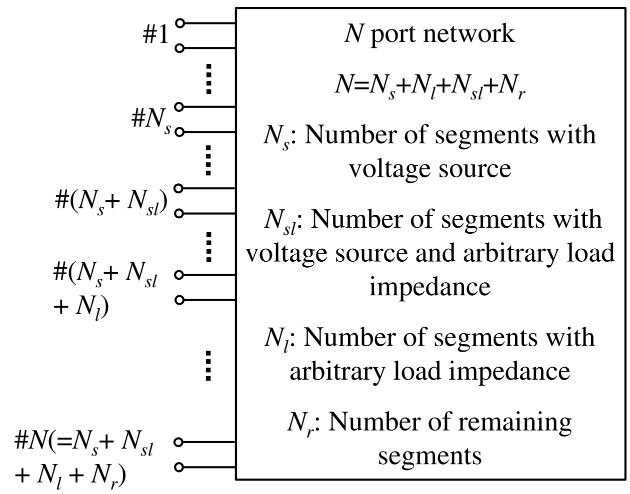

Fig. 2. An N-port network of the antennas shown in Fig. 1.

and block matrix equations corresponding to all different port conditions are clearly formulated. Additional computational cost of the proposed method is estimated quantitatively based on the formulation. Performance of the proposed method is demonstrated via numerical simulation of practical antennas with arbitrary load impedances.

This paper is organized as follows. In section II, detailed algorithm of the proposed method is clearly described and its additional computational cost is estimated. In section III, performance of the proposed method is demonstrated via a couple of practical numerical examples. Finally, conclusion is given in section IV.

II. Proposed Method

Fig. 1 shows antennas with load impedances. According to formulation of the MoM, unknown currents of the antennas can be obtained from an electric field integral equation (EFIE) [1]. Owing to basis/testing functions, the EFIE can be discretized into an $N \times N$ matrix equation where N is the number of unknown current segments. Resultant matrix equation can be solved using direct/iterative solvers and unknown current is obtained. Fig. 2 shows an N-port network model for the antennas shown in Fig. 1. For example, six ports of the antennas shown in Fig. 1 are connected to voltage source, load impedance, or both of them. All remaining ports are short circuited.

A matrix equation to be solved is as follows.

$$(\mathbf{Z}_{N\times N} + \mathbf{Z}^L_{N\times N})\mathbf{I}_N = \mathbf{V}_N \tag{1}$$

Here, $\mathbf{Z}_{N\times N}$ is an $N \times N$ impedance matrix, $\mathbf{Z}^L_{N\times N}$ is an $N \times N$ load impedance matrix, $\mathbf{I}_N$ is an N-dimensional current vector, and $\mathbf{V}_N$ is an N-dimensional voltage vector. It should be noted that $\mathbf{Z}^L_{N\times N}$ is diagonal matrix and its diagonal entries are nonzero for segments with load impedance while are zero for segments without load impedance. Eq. (1) can be transformed as follows.

$$\begin{aligned}\mathbf{Z}_{N\times N}\mathbf{I}_N &= \mathbf{V}_N - \mathbf{Z}^L_{N\times N}\mathbf{I}_N\\ \mathbf{I}_N &= \mathbf{Z}^{-1}_{N\times N}(\mathbf{V}_N - \mathbf{Z}^L_{N\times N}\mathbf{I}_N)\\ &= \mathbf{Y}_{N\times N}(\mathbf{V}_N - \mathbf{Z}^L_{N\times N}\mathbf{I}_N)\\ &(\because \mathbf{Y}_{N\times N} \equiv \mathbf{Z}^{-1}_{N\times N})\end{aligned} \tag{2}$$

Here, (2) is decomposed into block matrices/vectors in order to solve it efficiently for variable load impedance.

$$\begin{aligned}&\mathbf{I}_N = \mathbf{Y}_{N\times N}(\mathbf{V}_N - \mathbf{Z}^L_{N\times N}\mathbf{I}_N)\\ &\begin{bmatrix}\mathbf{I}_{N_s}\\ \mathbf{I}_{N_{sl}}\\ \mathbf{I}_{N_l}\\ \mathbf{I}_{N_r}\end{bmatrix} = \begin{bmatrix}\mathbf{Y}_{N_s\times N_s} & \mathbf{Y}_{N_s\times N_{sl}} & \mathbf{Y}_{N_s\times N_l} & \mathbf{Y}_{N_s\times N_r}\\ \mathbf{Y}_{N_{sl}\times N_s} & \mathbf{Y}_{N_{sl}\times N_{sl}} & \mathbf{Y}_{N_{sl}\times N_l} & \mathbf{Y}_{N_{sl}\times N_r}\\ \mathbf{Y}_{N_l\times N_s} & \mathbf{Y}_{N_l\times N_{sl}} & \mathbf{Y}_{N_l\times N_l} & \mathbf{Y}_{N_l\times N_r}\\ \mathbf{Y}_{N_r\times N_s} & \mathbf{Y}_{N_r\times N_{sl}} & \mathbf{Y}_{N_r\times N_l} & \mathbf{Y}_{N_r\times N_r}\end{bmatrix}\\ &\times\left\{\begin{bmatrix}\mathbf{V}_{N_s}\\ \mathbf{V}_{N_{sl}}\\ 0\\ 0\end{bmatrix} - \begin{bmatrix}0 & 0 & 0 & 0\\ 0 & \mathbf{Z}^L_{N_{sl}\times N_{sl}} & 0 & 0\\ 0 & 0 & \mathbf{Z}^L_{N_l\times N_l} & 0\\ 0 & 0 & 0 & 0\end{bmatrix}\right.\\ &\left.\times\begin{bmatrix}\mathbf{I}_{N_s}\\ \mathbf{I}_{N_{sl}}\\ \mathbf{I}_{N_l}\\ \mathbf{I}_{N_r}\end{bmatrix}\right\}\\ &= \begin{bmatrix}\mathbf{Y}_{N_s\times N_s} & \mathbf{Y}_{N_s\times N_{sl}} & \mathbf{Y}_{N_s\times N_l} & \mathbf{Y}_{N_s\times N_r}\\ \mathbf{Y}_{N_{sl}\times N_s} & \mathbf{Y}_{N_{sl}\times N_{sl}} & \mathbf{Y}_{N_{sl}\times N_l} & \mathbf{Y}_{N_{sl}\times N_r}\\ \mathbf{Y}_{N_l\times N_s} & \mathbf{Y}_{N_l\times N_{sl}} & \mathbf{Y}_{N_l\times N_l} & \mathbf{Y}_{N_l\times N_r}\\ \mathbf{Y}_{N_r\times N_s} & \mathbf{Y}_{N_r\times N_{sl}} & \mathbf{Y}_{N_r\times N_l} & \mathbf{Y}_{N_r\times N_r}\end{bmatrix}\\ &\times\begin{bmatrix}\mathbf{V}_{N_s}\\ \mathbf{V}_{N_{sl}} - \mathbf{Z}^L_{N_{sl}\times N_{sl}}\mathbf{I}_{N_{sl}}\\ -\mathbf{Z}^L_{N_l\times N_l}\mathbf{I}_{N_l}\\ 0\end{bmatrix}\end{aligned} \tag{3}$$

Here, N_s is the number of segments with voltage source, N_{sl} is the number of segments with voltage source and load impedance, N_l is the number of segments with load impedance, and N_r is the number of remaining segments (i.e. $N = N_s + N_l + N_{sl} + N_r$). $\mathbf{Y}_{i\times i}$ is $i \times i$ block admittance matrix where $i = N_s, N_l, N_{sl}, N_r$, $\mathbf{Z}^L_{N_{sl}\times N_{sl}}$ and $\mathbf{Z}^L_{N_l\times N_l}$ are

$N_{sl} \times N_{sl}$, $N_l \times N_l$ block load impedance matrices. $\mathbf{I}_{N_s}$, $\mathbf{I}_{N_{sl}}$, $\mathbf{I}_{N_l}$, and $\mathbf{I}_{N_r}$ are N_s, N_{sl}, N_l, and N_r dimensional block current vector, respectively. $\mathbf{V}_{N_s}$ and $\mathbf{V}_{N_{sl}}$ are N_s and N_{sl} dimensional block voltage vector, respectively. As shown in right hand side of (3), block load impedance matrices and unknown current of segments with load impedance are assumed to be equivalent voltage source.

Eq. (3) can be expressed using following four block matrix equations as follows.

$$\begin{cases} \mathbf{I}_{N_s} = \mathbf{Y}_{N_s\times N_s}\mathbf{V}_{N_s} + \mathbf{Y}_{N_s\times N_{sl}}(\mathbf{V}_{N_{sl}} - \mathbf{Z}^L_{N_{sl}\times N_{sl}}\mathbf{I}_{N_{sl}}) \\ \qquad -\mathbf{Y}_{N_s\times N_l}\mathbf{Z}^L_{N_l\times N_l}\mathbf{I}_{N_l} & (4) \\ \mathbf{I}_{N_{sl}} = \mathbf{Y}_{N_{sl}\times N_s}\mathbf{V}_{N_s} + \mathbf{Y}_{N_{sl}\times N_{sl}}(\mathbf{V}_{N_{sl}} - \mathbf{Z}^L_{N_{sl}\times N_{sl}}\mathbf{I}_{N_{sl}}) \\ \qquad -\mathbf{Y}_{N_{sl}\times N_l}\mathbf{Z}^L_{N_l\times N_l}\mathbf{I}_{N_l} & (5) \\ \mathbf{I}_{N_l} = \mathbf{Y}_{N_l\times N_s}\mathbf{V}_{N_s} + \mathbf{Y}_{N_l\times N_{sl}}(\mathbf{V}_{N_{sl}} - \mathbf{Z}^L_{N_{sl}\times N_{sl}}\mathbf{I}_{N_{sl}}) \\ \qquad -\mathbf{Y}_{N_l\times N_l}\mathbf{Z}^L_{N_l\times N_l}\mathbf{I}_{N_l} & (6) \\ \mathbf{I}_{N_r} = \mathbf{Y}_{N_r\times N_s}\mathbf{V}_{N_s} + \mathbf{Y}_{N_r\times N_{sl}}(\mathbf{V}_{N_{sl}} - \mathbf{Z}^L_{N_{sl}\times N_{sl}}\mathbf{I}_{N_{sl}}) \\ \qquad -\mathbf{Y}_{N_r\times N_l}\mathbf{Z}^L_{N_l\times N_l}\mathbf{I}_{N_l} & (7) \end{cases}$$

Eq. (6) can be transformed as follows.

$$\mathbf{I}_{N_l} = (\mathbf{U}_{N_l\times N_l} + \mathbf{Y}_{N_l\times N_l}\mathbf{Z}^L_{N_l\times N_l})^{-1}\times \\ (\mathbf{Y}_{N_l\times N_s}\mathbf{V}_{N_s} + \mathbf{Y}_{N_l\times N_{sl}}(\mathbf{V}_{N_{sl}} - \mathbf{Z}^L_{N_{sl}\times N_{sl}}\mathbf{I}_{N_{sl}})) \quad (8)$$

Here, $\mathbf{U}_{N_l\times N_l}$ is a $N_l \times N_l$ unit matrix. Eq. (8) is substituted into (5).

$$\mathbf{I}_{N_{sl}} = \mathbf{Y}_{N_{sl}\times N_s}\mathbf{V}_{N_s} + \mathbf{Y}_{N_{sl}\times N_{sl}}(\mathbf{V}_{N_{sl}} - \mathbf{Z}^L_{N_{sl}\times N_{sl}}\mathbf{I}_{N_{sl}}) \\ - \mathbf{Y}_{N_{sl}\times N_l}\mathbf{Z}^L_{N_l\times N_l}(\mathbf{U}_{N_l\times N_l} + \mathbf{Y}_{N_l\times N_l}\mathbf{Z}^L_{N_l\times N_l})^{-1} \\ \times (\mathbf{Y}_{N_l\times N_s}\mathbf{V}_{N_s} + \mathbf{Y}_{N_l\times N_{sl}}(\mathbf{V}_{N_{sl}} - \mathbf{Z}^L_{N_{sl}\times N_{sl}}\mathbf{I}_{N_{sl}})) \quad (9)$$

After transposition of terms in (9), following equation to be solved is obtained.

$$\mathbf{P}_{N_{sl}\times N_{sl}}\mathbf{I}_{N_{sl}} = \mathbf{Q}_{N_{sl}}, \quad (10)$$

where

$$\mathbf{P}_{N_{sl}\times N_{sl}} = \mathbf{U}_{N_{sl}\times N_{sl}} + \mathbf{Y}_{N_{sl}\times N_{sl}}\mathbf{Z}^L_{N_{sl}\times N_{sl}} \\ - \mathbf{Y}_{N_{sl}\times N_l}\mathbf{Z}^L_{N_l\times N_l}(\mathbf{U}_{N_l\times N_l} + \mathbf{Y}_{N_l\times N_l}\mathbf{Z}^L_{N_l\times N_l})^{-1} \\ \times \mathbf{Y}_{N_l\times N_{sl}}\mathbf{Z}^L_{N_{sl}\times N_{sl}} \quad (11)$$

$$\mathbf{Q}_{N_{sl}} = \mathbf{Y}_{N_{sl}\times N_s}\mathbf{V}_{N_s} + \mathbf{Y}_{N_{sl}\times N_{sl}}\mathbf{V}_{N_{sl}} \\ - \mathbf{Y}_{N_{sl}\times N_l}\mathbf{Z}^L_{N_l\times N_l}(\mathbf{U}_{N_l\times N_l} + \mathbf{Y}_{N_l\times N_l}\mathbf{Z}^L_{N_l\times N_l})^{-1} \\ \times (\mathbf{Y}_{N_l\times N_s}\mathbf{V}_{N_s} + \mathbf{Y}_{N_l\times N_{sl}}\mathbf{V}_{N_{sl}}) \quad (12)$$

Here, $\mathbf{U}_{N_{sl}\times N_{sl}}$ is a $N_{sl} \times N_{sl}$ unit matrix. Once $\mathbf{I}_{N_{sl}}$ is obtained from (10), $\mathbf{I}_{N_l}$ is readily obtained from (8). Finally, $\mathbf{I}_{N_{sl}}$ and $\mathbf{I}_{N_l}$ are substituted into Eqs. (4) and (7), $\mathbf{I}_{N_s}$ and $\mathbf{I}_{N_r}$ are obtained. According to Eqs. (4)∼(10), current distribution of the antenna with variable load impedance can be obtained without inverting $(\mathbf{Z}_{N\times N} + \mathbf{Z}^L_{N\times N})$ every time when the load impedance varies (i.e. one time inversion of $\mathbf{Z}_{N\times N}$ is only necessary.).

As shown in this section, the proposed method just solves block matrix equations sequentially and no additional approximations are introduced for its matrix solver. Moreover, no additional approximations to the MoM impedance matrix are introduced to the proposed method. Therefore, current of the antenna obtained using the proposed method perfectly agrees with that of full-wave analysis, i.e. the MoM.

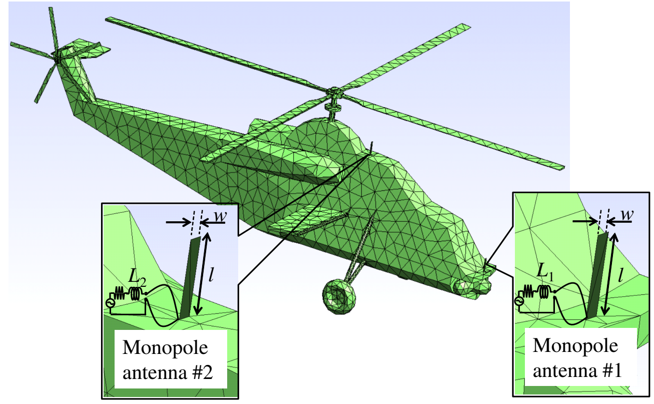

Fig. 3. Two monopole antennas on a helicopter.

Except for inversion of the full-impedance matrix, the most computationally expensive part is inversion of $\mathbf{P}_{N_{sl}\times N_{sl}}$ in (10) or inversion of $(\mathbf{U}_{N_l\times N_l} + \mathbf{Y}_{N_l\times N_l}\mathbf{Z}^L_{N_l\times N_l})$ in (8). The order of CPU time for inversion of these matrices is proportional to N_{sl}^3 and N_l^3, respectively, while that of the full-impedance matrix inversion is proportional to N^3. As shown in numerical examples later, variable load impedances are often loaded with small number of segments. As a result, N_{sl} and N_l are much smaller than N for practical antennas. Therefore, CPU time of the proposed method is negligibly small once the inversion of the unloaded full-impedance matrix is performed.

III. Numerical Simulation

Performance of the proposed method is demonstrated via numerical simulation. Numerical simulation was performed using the MoM with Rao-Wilton-Glisson (RWG) basis function [17]. The MoM with RWG basis function has been implemented by authors as a full-wave analysis method. All numerical simulations in this paper were implemented on an Intel Core i7-4800MQ 2.7 GHz processor with 16 GB RAM.

A. Antennas with variable inductors

Fig. 3 is two monopole antennas on a helicopter. This numerical example is for demonstrating the performance of the proposed method on complicated antennas. Operating frequency is $f = 100$ MHz, length of the monopole antennas is $l = 0.5$ m, and width of the monopole antennas is $w = 0.1$ m. Size of the helicopter is approximately $18.6\times22\times7.1$ m^3. The two monopole antennas are uniformly excited at their ports and variable inductances L_1 and L_2 are loaded with ports for impedance matching. The inductances vary from 50 nH to 140 nH and its interval is 10 nH. The total number of unknowns is $N = 7,537$, $N_s = N_l = 0$, $N_{sl} = 2$, $N_r = 7,535$.

Surface current distributions of two monopole antennas on the helicopter obtained using full-wave analysis and the proposed method are shown in Figs. 4 and 5, respectively.

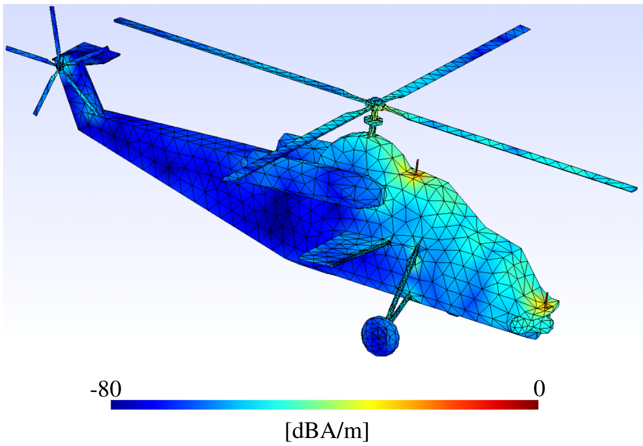

Fig. 4. Surface current distribution of two monopole antennas on a helicopter ($L_1 = L_2 = 100$ nH, Full-wave analysis).

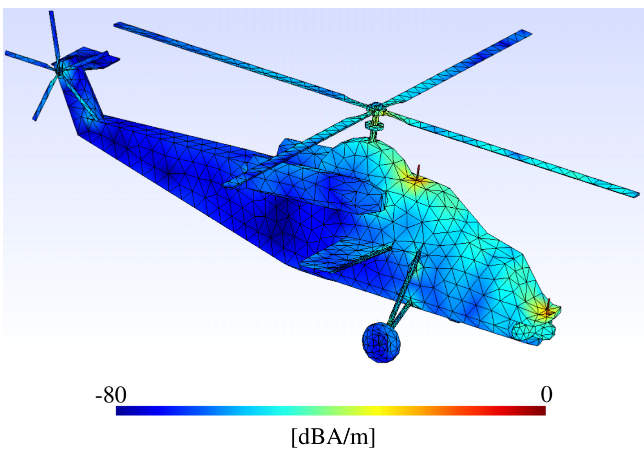

Fig. 5. Surface current distribution of two monopole antennas on a helicopter ($L_1 = L_2 = 100$ nH, Proposed method).

As mentioned earlier, the proposed method does not include approximation, surface current distribution obtained using the proposed method is in perfect agreement with that of the full-wave analysis.

The total CPU time is 136,473 sec. for full-wave analysis while 15,927 sec. for the proposed method. For full-wave analysis, inversion of the $N \times N$ full-impedance matrix is necessary every time when inductance varies. Therefore, inversion of the $N \times N$ full-impedance matrix with load impedance was performed 10 times and resultant CPU time was long. On the other hand, inversion of the $N \times N$ unloaded full-impedance matrix was performed only one time for the proposed method. Resultant $N \times N$ unloaded full-admittance matrix is stored and reused every time when inductance varies. Therefore, total CPU time of the proposed method is quite small because small block matrix equations are only required to be solved every time when inductance varies. It can be concluded that the proposed method is quite efficient for numerical analysis of complicated antennas with variable load impedance.

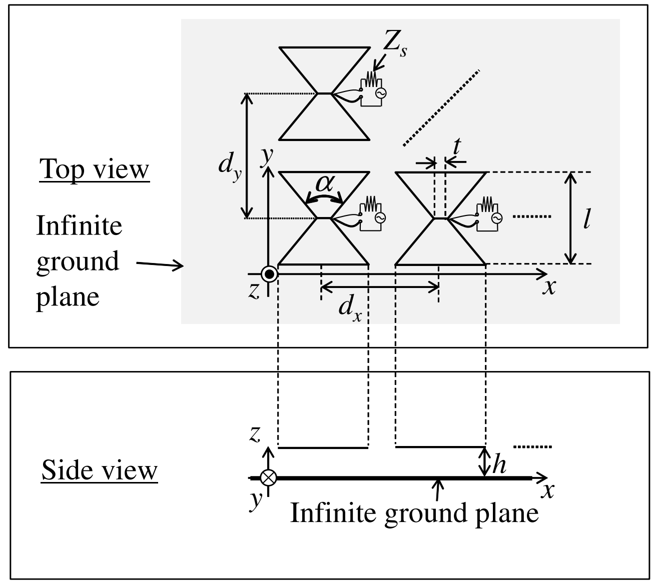

Fig. 6. A planar bowtie array antenna over an infinite ground plane.

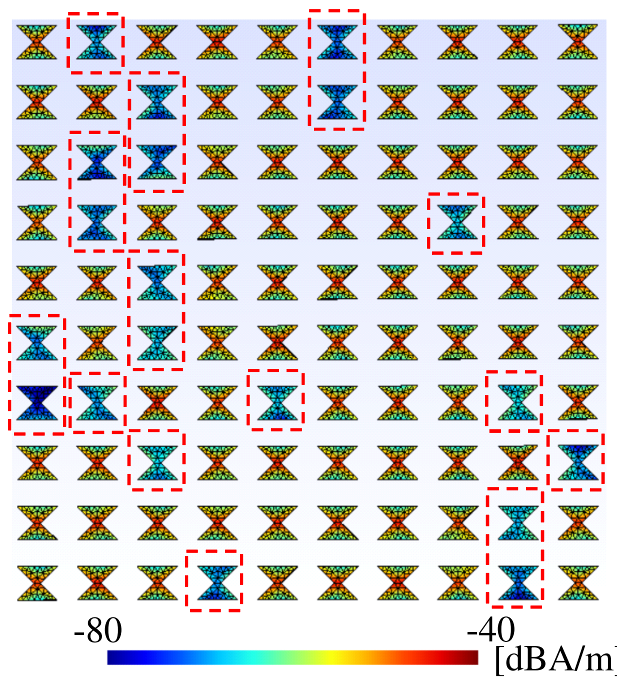

Fig. 7. Surface current distribution of a 10×10 bowtie array antenna over an infinite ground plane (Proposed method).

B. Array antennas with defective elements

Fig. 6 is a planar bowtie array antenna over an infinite ground plane. Multiple array elements are assumed to be defective elements whose feeding ports are open circuited. $Z_s = 100,000\ \Omega$ is loaded with defective elements to be open circuited while $Z_s = 50\ \Omega$ is loaded with operating elements. This numerical example is applicable to source reconstruction of array antennas and their diagnosis. Operating frequency is $f = 2$ GHz, length of an array element is $l = 0.1$ m, width of

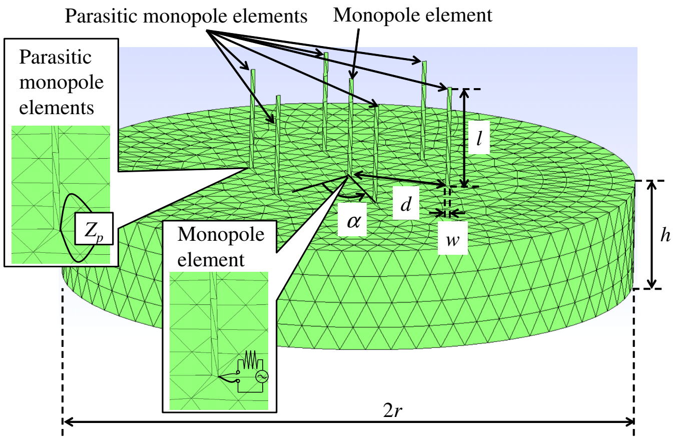

Fig. 8. ESPAR-antenna.

feeding edge is $t = 0.01$ m, opening angle of the array element is $\alpha = 45°$. Height of the array element from the ground plane is $h = 0.0375$ m, array spacing is $d_x = d_y = 0.09$ m, the total number of array elements is $10 \times 10 = 100$. All array elements are excited uniformly. The array element is divided into 63 unknowns and infinite ground plane is modeled as image of the array antenna. The total number of unknowns including image of the array antenna is $N = 12,600$, $N_s = N_l = 0$, $N_{sl} = 200$, $N_r = 12,400$. During the numerical simulation, 20 % of array elements are assigned to be defective elements in a random manner and numerical simulation is performed 100 times.

Fig. 7 shows surface current distribution of a 10×10 bowtie array antenna including 20 % defective elements obtained using the proposed method. Array elements surrounded by dashed red lines are defective elements. It is found that current of defective elements is smaller than that of operating elements because defective elements are open circuited during numerical simulation.

The total CPU time is 76,176 sec. for the proposed method and most of the CPU time is for inversion of the $N \times N$ unloaded full-impedance matrix. Remaining CPU time is mainly for inversion of the $N_{sl} \times N_{sl}$ block impedance matrix $\mathbf{P}_{N_{sl} \times N_{sl}}$ in (10) but is negligibly small because $N_{sl} \ll N$. For example, CPU time for inversion of $\mathbf{P}_{N_{sl} \times N_{sl}}$ was only 0.3 sec. for the array antenna where $N_{sl} = 200$. Therefore, it can be said that the proposed method enables to perform a large number of numerical simulations at the expense of small CPU time. On the other hand, 100 times of full-wave analysis were unable because of long CPU time estimated as 100 days. According to results of numerical simulation, it is demonstrated that the proposed method enables to reflect the effect of load impedance to numerical results, directly.

C. ESPAR Antenna

Fig. 8 shows an electronically steerable passive array radiator (ESPAR) antenna on a ground skirt [18]-[20]. Operating frequency is $f = 300$ MHz, length of monopole elements is $l = 0.25$ m, width of monopole elements is $w = 0.01$ m,

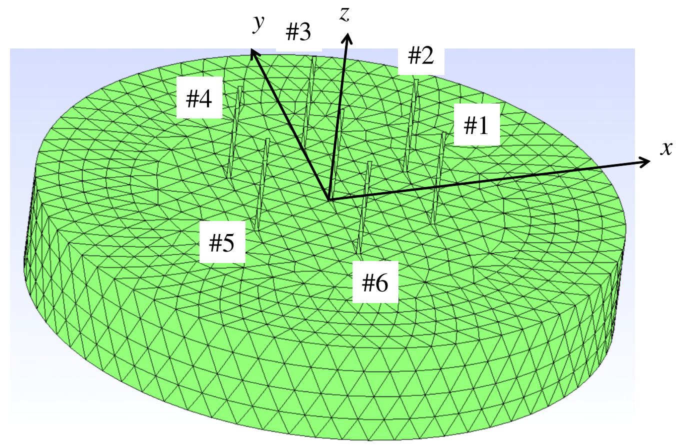

Fig. 9. Element number of parasitic monopole elements of the ESPAR antenna.

TABLE I
VARIABLE LOAD IMPEDANCE OF PARASITIC MONOPOLE ELEMENTS.

Direction of mainbeam	Combination of load impedances
$\phi_0 = 90°$	$L = 10$ nH for #5, #6, #1 $C = 10$ pF for #2, #3, #4
$\phi_0 = 150°$	$L = 10$ nH for #6, #1, #2 $C = 10$ pF for #3, #4, #5
$\phi_0 = 210°$	$L = 10$ nH for #1, #2, #3 $C = 10$ pF for #4, #5, #6
$\phi_0 = 270°$	$L = 10$ nH for #2, #3, #4 $C = 10$ pF for #5, #6, #1
$\phi_0 = 330°$	$L = 10$ nH for #3, #4, #5 $C = 10$ pF for #6, #1, #2
$\phi_0 = 30°$	$L = 10$ nH for #4, #5, #6 $C = 10$ pF for #1, #2, #3

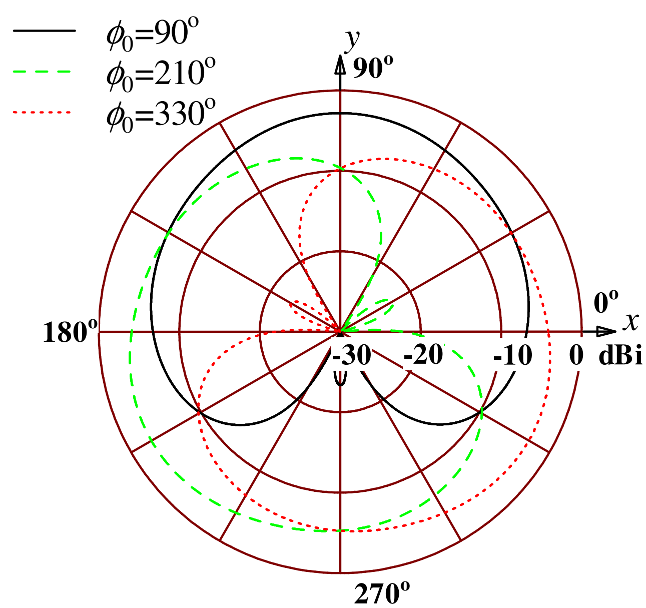

Fig. 10. Actual gain patterns of ESPAR antennas (Proposed method).

spacing between the monopole element and parasitic elements is $d = 0.25$ m, angle between two adjacent parasitic elements is $\alpha = 60°$. Height and radius of the ground skirt are $h = 0.25$ m and $r = 0.7$ m, respectively. A central monopole element is excited at its port by voltage source with $50\ \Omega$ source resistance. The total number of unknowns is $N = 3,902$, $N_s = 0$, $N_l = 6$, $N_{sl} = 1$, $N_r = 3,895$. According to variable load impedance at parasitic elements surrounding the central monopole element, the ESPAR antenna enables to steer its main beam. Table I shows direction of mainbeam and corresponding combination of variable load impedances. Parasitic monopole elements with inductors (i.e. $Z_p = j\omega L$) work as reflectors while those with capacitances (i.e. $Z_p = \frac{1}{j\omega C}$) work as directors.

Fig. 10 shows actual gain patterns of the ESPAR antenna, whose mainbeam direction is $\phi_0 = 90°, 210°, 330°$. It is found that direction of main beam can be scanned as the load impedance varies. The total CPU time for numerical analysis of the ESPAR antennas with six combination of variable load impedances shown in TABLE I is 1,808 sec. for the proposed method while is 9,649 sec. for full-wave analysis. According to results of numerical simulation, it is demonstrated that the proposed method is applicable to design of reconfigurable antennas using variable load impedance.

IV. Conclusion

In this paper, a computationally efficient numerical analysis method for antennas with variable load impedance has been proposed. The proposed method deals with the load impedance and its current as equivalent voltage source. As a result, current distribution of the antennas with variable load impedance can be obtained directly once unloaded full-impedance matrix is inverted. The proposed method does not include approximation and resultant current shows perfect agreement with that of full-wave analysis. Performance of the proposed method has been shown via numerical simulation of practical antennas with variable load impedance.

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References

[1] R. F. Harrington, Field Computation by Moment Methods, Macmillan, New York, 1968.

[2] J. H. Richmond and N. H. Geary, “Mutual impedance of nonplanar-skew sinusoidal dipoles,” IEEE Trans. Antennas Propag., vol. 23, no. 3, pp. 412-414, May 1975.

[3] C.H. Chuang, J.H. Richmond, N. Wang, and P.H. Pathak, “New expressions for mutual impedance of nonplanar-skew sinusoidal monopoles,” IEEE Trans. Antennas Propag., vol. 38, no. 2, pp.275-276, Feb. 1990.

[4] S.M. Rao, D.R. Wilton, and A.W. Glisson, “Electromagnetic scattering by surfaces of arbitrary shape,” IEEE Trans. Antennas Propag., vol. AP-30, no. 3, pp.409-418, May 1982.

[5] D.H. Schaubert, D.R. Wilton, and A.W. Glisson, “ A tetrahedral modeling method for electromagnetic scattering by arbitrarily shaped inhomogeneous dielectric bodies, ” IEEE Trans. Antennas Propag., vol. AP-32, no. 1, pp. 77—85, Jan. 1984.

[6] R. Coifuman, V. Rokhlin, and S. Wandzura, “The fast multipole method for the wave equation: A pedestrian prescription,” IEEE Antennas Propag. Mag., vol.35, no.3, pp.7-12, June 1993.

[7] V. Rokhlin, “Rapid solution of integral equations of scattering theory in two dimensions,” J. Comput. Phys., vol.86, no.2, pp.414-439, Feb. 1990.

[8] E. Bleszynski, M. Bleszynski, and T. Jaroszewicz, “AIM: Adaptive integral method for solving large-scale electromagnetic scattering and radiation problems,” Radio Science, vol.31, no.5, pp.1225-1251, Sept.-Oct. 1996.

[9] T.K. Sarkar, E. Arvas, and S.M. Rao, “Application of FFT and the conjugate gradient method for the solution of electromagnetic radiation from electrically large and small conducting bodies,” IEEE Trans. Antennas Propag., vol.AP-34, no.5, pp.635-640, May 1986.

[10] V. V. S. Prakash and R. Mittra, “Characteristic basis function method: A new technique for efficient solution of method of moments matrix equations,” Microw. Opt. Technol. Lett., vol. 36, no. 2, pp. 95-100, Jan. 2003.

[11] K. Konno, Q. Chen, and R.J. Burkholder, “Numerical analysis of large-scale finite periodic arrays using a macro block-characteristic basis function method,” IEEE Trans. Antennas Propag., vol. 65, no. 10, pp. 5348-5355, Oct. 2017.

[12] W. B. Lu, T. J. Cui, Z. G. Qian, X. X. Yin, and W. Hong, “Accurate analysis of large-scale periodic structures using an efficient sub-entire-domain basis function method,” IEEE Trans. Antennas Propag., vol. 52, no. 11, pp. 3078-3085, Nov. 2004.

[13] P. Du, B.-Z. Wang, and H. Li, “An extended sub-entire domain basis function method for finite periodic structures,” IEEE Antennas Wireless Propag. Lett., vol. 7, pp. 404-407, 2008.

[14] H. Kamoda, T. Iwasaki, J. Tsumochi, T. Kuki, and O. Hashimoto, “60-GHz electronically reconfigurable large reflectarray using single-bit phase shifters,” IEEE Trans. Antennas Propag., vol. 59, no. 7, pp.2524-2531, July 2011.

[15] N. Honma, T. Seki, K. Nishikawa, K. Tsunekawa, and K. Sawaya, “Series-fed beam-scanning antenna employing multi-stage configured microstrip antennas with tunable reactance devices,” IEICE Trans. Commun., vol. E88-B, no. 6, pp.2297-2304, June 2005.

[16] Q. Yuan and T. Ohira, “Equivalent excitation of varactors combined with MoM,” IEICE Tech. Rep., vol. 110, no. 447, AP2010-193, pp. 117-122, March 2011 (In Japanese).

[17] S. M. Rao, D. R. Wilton, and A. W. Glisson, “Electromagnetic scattering by surfaces of arbitrary shape,” IEEE Trans. Antennas Propag., vol. 30, no. 3, pp. 409-418, May 1982.

[18] K. Yang and T. Ohira, “Realization of space—time adaptive filtering by employing electronically steerable passive array radiator antennas,” IEEE Trans. Antennas Propag., vol. 51, no. 7, pp. 1476-1485, July 2003.

[19] J. Cheng, M. Hashiguchi, K. Iigusa, and T. Ohira, “Electronically steerable parasitic array radiator antenna for omni- and sector pattern forming applications to wireless ad hoc networks,” IEE Proc. Microw. Antennas Propag., vol. 150, no. 4, pp. 203-208, Aug. 2003.

[20] R. Schlub, J. Lu, and T. Ohira, “Seven-element ground skirt monopole ESPAR antenna design from a genetic algorithm and the finite element method,” IEEE Trans. Antennas Propag., vol. 51, no. 11, pp. 3033-3039, Nov. 2003.