Interpolation Algorithm for Fast Evaluation of EM Coupling between Wires

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Abstract – Efficient and accurate evaluation of the EM field radiated by a current flowing along a wire is essential to solve the electromagnetic coupling between arbitrary oriented wires. In this paper, a numerically efficient algorithm for the evaluation of coupling is presented. The currents along the wires are expressed in terms of local basis functions. The coupling between each two expansion functions of different wires, using an exact kernel and the Galerkin Method of Moments, requires an integration over the mantle of the radiating element and an integration over the mantle of the receiving element. The computational cost for this 2×2D integration is reduced by an interpolation technique. In order to reduce the number of evaluation points and to control accuracy, the interpolation technique is applied to a function that represents the difference between the electric field radiated by a wire element and the analytically known point dipole field. The proposed algorithm is implemented using already available standard routines.

Keywords: Exact and reduced kernel, electromagnetic scattering, Method of Moments, integral equation, fast algorithms, and antennas.

I. INTRODUCTION

Consider the analysis of the electromagnetic coupling between arbitrary oriented wires along which the currents are expanded by means of basis functions. Applying the Galerkin Method of Moments (MoM) [1], the computation of electromagnetic coupling between different wires is based on the evaluation of the electric field radiated by the current of a single wire expansion function over the surface of the observation element (e.g. a segment or two adjacent segments). This field induces a current on all other wires, appointing therefore the mutual interaction. Efficient and accurate field evaluation is therefore essential to solve electromagnetic coupling between wires especially when wires are close to each other.

Focusing on the analysis of scattering from an arbitrary oriented wire, an Electric Field Integral Equation (EFIE) known as Pocklington's equation is solved by applying the Galerkin MoM. The induced current and the scattered fields are interrelated by a so called "kernel". Choosing the observation point on the central axis of the wire results in the "reduced kernel" formulation, while choosing the observation point on the mantle surface results in the "exact kernel" [2, 3]. In particular the exact kernel formulation provides more accurate results than the reduced one for the analysis of thick wires ($ka \approx 0.1$, being a the wire radius) [4, 5]. As a consequence, in order to be consistent with the exact kernel formulation, the evaluation of coupling matrices requires the calculation of the field on the mantle of wires yielding the computation of 2D integrals.

Even though efficient techniques are employed for the computation of impedance matrix elements of a single wire [4, 5], a computational burden is present when coupling matrices have to be evaluated.

A similar computational bottleneck has been encountered in the analysis of electromagnetic scattering from 3D objects [6] where the time spent on calculation of matrix elements is much more than solving the system matrix. Furthermore when computing time-harmonic scattering from 1D/2D large PEC objects, the computational complexity is in the evaluation of fields produced by a given current distribution [7, 8]. In all these cases, tabulation and interpolation techniques have been investigated in order to accelerate the evaluation of the problem solution and to guarantee the required accuracy.

In a similar fashion this paper presents a novel, efficient and accurate scheme well suited to accelerate the computation of coupling matrix elements. The electric field radiated by a current flowing along a wire element can be adequately sampled on a non-uniform grid defined in a way that is consistent with the field behavior [9] and subsequently interpolated [10]. In order to further accelerate the generation of coupling matrix elements and to control the accuracy, an interpolation technique is applied to a smoothed function obtained by subtracting a properly chosen analytical term from the actual electric field contribution.

II. FORMULATION

In this section, our attention is firstly focused on the analysis of the current induced along a single wire by an impressed voltage and/or by an incident plane wave.

Secondly, an expression for the electric field radiated by such induced current distribution is derived.

A. Single Wire

In free space consider a single PEC wire antenna along the positive z axis fed by a voltage delta gap or/and illuminated by an incident field. A cylindrical coordinate system (ρ, φ, z) is introduced. The induced current along the mantle satisfies the frequency domain electric field integral equation (EFIE), well-known as Pocklington's equation [2],

$$\left(\partial_z^2 + k^2\right) \int_0^L K_E(z - z', \omega) I(z', \omega) dz' = -j\omega\epsilon \left(V(\omega)\delta(z - z_g) + E_z^i(z\mathbf{i}_z, \omega)\right) \quad (1)$$

where L is the wire length and $K_E(z - z', \omega)$ denotes the exact kernel [4, 5]. Equation (1) is discretized by applying the Galerkin Method of Moments [1]. Therefore, a set of N rooftop basis functions $\psi_n(z)$ is defined with a uniform mesh-size $\Delta = L/(N + 1)$ such that the current distribution is approximated as,

$$I(z) \approx \sum_{n=1}^{N} I_n \psi_n(z). \quad (2)$$

Next, by choosing the same type of rooftop functions $\psi_j(z),\ j = 1, \ldots, N$ as testing functions, an inner product is defined and applied on both sides of equation (1). Thus, equation (1) is discretized as,

$$\underline{\underline{\mathbf{Z}}}\ \underline{\mathbf{I}} = \underline{\mathbf{F}}_e \quad (3)$$

where $\underline{\underline{\mathbf{Z}}}$ denotes a $N \times N$ symmetric Toeplitz matrix, and the N-dimensional vectors $\underline{\mathbf{I}}$, $\underline{\mathbf{F}}_e$ represent the unknown current distribution vector and the weighted forcing excitation vector due to external sources, respectively.

The general expression of the electric field as a function of the magnetic vector potential is here recalled,

$$\mathbf{E} = \frac{1}{j\omega\epsilon} \left(k^2 \mathbf{A} + \nabla (\nabla \cdot \mathbf{A})\right) \quad (4)$$

$$\mathbf{A}(\mathbf{r}) = \mathbf{i}_z \int_{z'=0}^{L} \frac{I(z')}{2\pi a} \int_{\varphi'=0}^{2\pi} \frac{\exp(-jkR)}{4\pi R} a d\varphi' dz', \quad (5)$$

and then employed in order to find the electric field radiated by a straight wire. In equation (5), $R = |\mathbf{r} - \mathbf{r}'|$ denotes the distance between the source point $\mathbf{r}'$ on the wire mantle and the observation point $\mathbf{r}$ where the field is computed. In particular, since the current is expanded in terms of N rooftop basis functions $\psi_n(z)$, the radiated electric field can be expressed as a sum of N separate electric field contributions,

$$\mathbf{E}(\mathbf{r}) = \sum_{n=1}^{N} I_n \mathbf{E}_n(\mathbf{r}). \quad (6)$$

The elementary field $\mathbf{E}_n$ represents the field radiated by the current of the n-th basis function (i.e., two adjacent segments). This field is explicitly formulated as,

$$\mathbf{E}_n(\mathbf{r}) = \frac{1}{j\omega\epsilon} \frac{1}{8\pi^2} \int_{z'=(n-1)\Delta}^{(n+1)\Delta} \psi_n(z') \int_{\varphi'=0}^{2\pi} \frac{\exp(-jkR)}{R^3} \times \left\{ -\left[(jkR)^2 + jkR + 1\right] \mathbf{i}_z + \left[(jkR)^2 + 3jkR + 3\right] \frac{(\mathbf{r} - \mathbf{r}')}{R} \frac{(z - z')}{R} \right\} d\varphi' dz'. \quad (7)$$

Note the double integration that follows from considering the electric current flow on the surface of the wire.

B. Mutual Coupling

As a simplification, without loss of generality, the mutual coupling between two wires in the configuration depicted in Fig. 1 is here discussed.

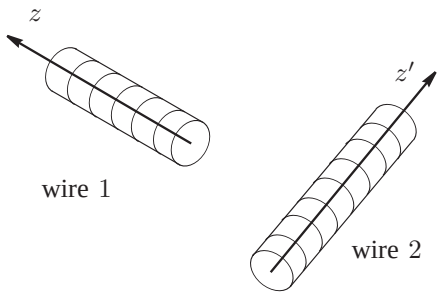

Fig. 1. Wires geometry.

Consider two perfectly conducting wires, wire 1 and wire 2, with their own local coordinate systems. External sources are present (i.e., voltage delta gap or/and incident field). The currents are approximated by means of N_1 and N_2 rooftop basis functions defined on $N_1 + 1$ and $N_2 + 1$ segments, respectively. It is noted that a wire element is formed by two adjacent wire segments which represent the support of an expansion function. Henceforth, the n-th basis function of wire 2 and the m-th testing function of wire 1 will be referred to as n-th source (or radiating) and m-th observation element. In this environment, an induced current of the n-th basis function ψ_n of wire 2, radiates an elementary electric field $\mathbf{E}_n(\mathbf{r})$ as in (7). This field impinges on wire 1 and induces a current along each of its segments. To describe the mutual coupling, the total incident field on the m-th observation element of wire 1 is written as the sum of two parts,

$$\mathbf{E}_m^{(1)}(\mathbf{r}) = \mathbf{E}_{ext}^{(1)}(\mathbf{r}) + \sum_{n=1}^{N_2} \mathbf{E}_n^{(1,2)}(\mathbf{r}). \quad (8)$$

The first term on the right-hand side of equation (8) denotes the field due to external sources. The summation represents the field induced on wire 1 by the current

flowing along all the source elements of wire 2. It is noted that wire 1 and wire 2 can be interchanged without losing validity of the approach.

By applying the Galerkin MoM procedure to this problem, the system matrix equation $\underline{\underline{\mathbf{Z}}}\underline{\mathbf{I}} = \underline{\mathbf{F}}_e$, is elegantly extended as follows

$$\begin{bmatrix} \underline{\underline{\mathbf{Z}}}^1 & -\underline{\underline{\mathbf{C}}}^{1,2} \\ -\underline{\underline{\mathbf{C}}}^{2,1} & \underline{\underline{\mathbf{Z}}}^2 \end{bmatrix} \begin{bmatrix} \underline{\mathbf{I}}^1 \\ \underline{\mathbf{I}}^2 \end{bmatrix} = \begin{bmatrix} \underline{\mathbf{F}}_e^1 \\ \underline{\mathbf{F}}_e^2 \end{bmatrix} \tag{9}$$

in which each forcing vector on the right-hand side represents the weighted field of external origin. The diagonal "self matrices" $\underline{\underline{\mathbf{Z}}}^1$, $\underline{\underline{\mathbf{Z}}}^2$, represent the interaction between elements of the same wire, while, the off-diagonal "coupling matrices" $\underline{\underline{\mathbf{C}}}^{1,2}$, $\underline{\underline{\mathbf{C}}}^{2,1}$, describe the interaction between elements of different wires.

In a similar fashion equation (9) can be expanded to a generic number of P arbitrary oriented wires by using,

$$\begin{bmatrix} \underline{\underline{\mathbf{Z}}}^1 & -\underline{\underline{\mathbf{C}}}^{1,2} & \dots & -\underline{\underline{\mathbf{C}}}^{1,P} \\ -\underline{\underline{\mathbf{C}}}^{2,1} & \underline{\underline{\mathbf{Z}}}^2 & \dots & -\underline{\underline{\mathbf{C}}}^{2,P} \\ \vdots & \vdots & \vdots & \vdots \\ -\underline{\underline{\mathbf{C}}}^{P,1} & -\underline{\underline{\mathbf{C}}}^{P,2} & \dots & -\underline{\underline{\mathbf{Z}}}^P \end{bmatrix} \begin{bmatrix} \underline{\mathbf{I}}^1 \\ \underline{\mathbf{I}}^2 \\ \vdots \\ \underline{\mathbf{I}}^P \end{bmatrix} = \begin{bmatrix} \underline{\mathbf{F}}_e^1 \\ \underline{\mathbf{F}}_e^2 \\ \vdots \\ \underline{\mathbf{F}}_e^P \end{bmatrix}. \tag{10}$$

III. EFFICIENT EVALUATION OF MATRIX ELEMENTS

Elements of self matrices and the known excitation vectors are efficiently evaluated as explained in [4]. Our attention is focused on the computation of coupling matrix elements. Each one of them involves the evaluation of two times a double integral. Consider the (m, n)-th element of matrix $\underline{\underline{\mathbf{C}}}^{1,2}$. It represents the mutual interaction between the field radiated by a current flowing along element n of wire 2 and the m-th element of wire 1. A first double integration has to be carried out to evaluate the field $\mathbf{E}_n^{(2)}$ radiated by a current distribution along the n-th source element of wire 2, (see equation (7)). An additional double integral has to be computed to determine the mutual interaction between this field and the induced current distribution along the m-th observation element of wire 1. This kind of straightforward calculation is a time-consuming process since many integrals are involved. To simplify the notation, the n-th radiating element is placed in the center of a cylindrical coordinate system (ρ', φ', z') as shown in Fig. 2.

Noting that the electric field $\mathbf{E}_n^{(2)}$ is rotationally symmetric, the observation region of the n-th source element can be defined in a plane (ρ', z') with $\rho' > 0$ by $\rho_{\min}$, $\rho_{\max}$, $z_{\min}$, $z_{\max}$, as depicted in Fig. 2. This region contains the projection of the mantle of the m-th observation element of wire 1.

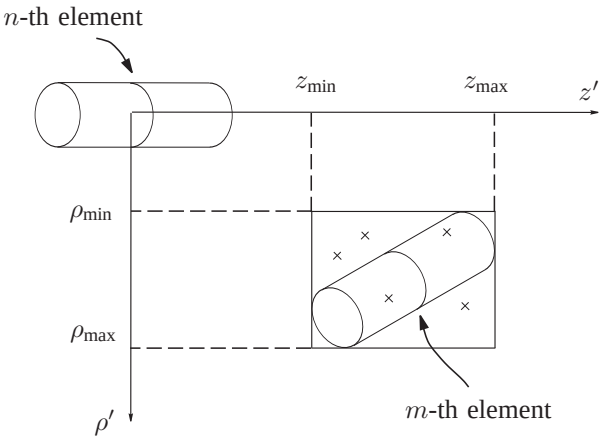

Fig. 2. Interpolation area defined to evaluate the electromagnetic interaction between two wire elements.

A. Interpolation technique

Instead of carrying out the straightforward double integration in equation (7), the field $\mathbf{E}_n^{(2)}$ is computed on a discrete grid of points within the observation region and is subsequently interpolated. For this purpose, we have investigated tabulation and interpolation techniques for the evaluation of the radiated electric field in order to accelerate the computation of coupling matrix elements, [6]. Even though uniform and random sampling algorithms have been explored, the most efficient technique in choosing points is by sampling the radiated electric field in a way that follows the behavior of the field. The proposed algorithm is developed based on the following three ideas:

1) An interpolation technique is applied to reduce the number of points where the radiated electric field is computed.
2) By making use of standard routines a set of points on a non-uniform grid is generated and an interpolation function is then defined. A numerical adaptive multidimensional integration routine [9] has been modified in order to generate a set of points used in a subsequent interpolation step where a piecewise polynomial surface is defined as interpolant function following the method proposed by Renka and Cline [10] (routines E01SAF, E01SBF). In this way, the integration routine will choose the distribution of points according to the behavior of the field $\mathbf{E}_n^{(2)}$ which is to be interpolated.
3) By subtracting the analytically known point dipole field from the elementary electric field $\mathbf{E}_n^{(2)}$ and by applying the interpolation technique to this difference, the efficiency of the method is improved for a fixed accuracy.

In order to accelerate the generation of coupling matrices, further efficiency is expected by observing that the function to be interpolated is relatively smooth. This term

should be quick to evaluate and resemble the (far) field of the n-th radiating element. A function difference $\mathbf{D}_n^{(2)}$ is defined as the difference of the field $\mathbf{E}_n^{(2)}$ and the electric field $\mathbf{E}^{(p)}$ radiated by a point dipole placed in the origin of the coordinate system. Function $\mathbf{E}^{(p)}$ behaves asymptotically (i.e., for $R \to \infty$) as $\mathbf{E}_n^{(2)}$ and is singular when the distance R vanishes,

$$\mathbf{E}^{(p)}(\mathbf{r}) = \frac{1}{j\omega\epsilon}\frac{\Delta_1}{4\pi}\frac{\exp(-jkR)}{R^3}\left\{-\left[(jkR)^2 + jkR + 1\right]\mathbf{i}_z + \left[(jkR)^2 + 3jkR + 3\right]\frac{\mathbf{r}z}{R^2}\right\}. \quad (11)$$

Thanks to these properties, the resulting function $\mathbf{D}_n^{(2)} = \mathbf{E}_n^{(2)} - \mathbf{E}^{(p)}$ has a behavior considerably smoother than $\mathbf{E}_n^{(2)}$ and is therefore interpolated in a numerically easier way with a higher accuracy. The flow chart in Fig. 3 shows the fundamental steps of the proposed algorithm. Our numerical scheme begins by setting a desired accuracy ε with which the elementary radiated electric field of equation (7) has to be evaluated. While the adaptive routine [9] integrates numerically the function difference $\mathbf{D}_n^{(2)}$, choosing points (ρ_i', z_i') in the observation region $[\rho_{\min},\ \rho_{\max}] \times [z_{\min},\ z_{\max}]$ following the behavior of this function, the implemented scheme gathers the first N_a points $\{(\rho_i', z_i')\}_{i=1}^{N_a}$. Next, on this non-uniform set of N_a points an interpolated function $\widetilde{\mathbf{D}}_n^{(2)}$ is defined by using a NAG routine [10]. To examine the obtained accuracy of $\widetilde{\mathbf{D}}_n^{(2)}$ compared to $\mathbf{D}_n^{(2)}$ and normalized to the incident field $\mathbf{E}_n^{(2)}$, a relative error $\widetilde{\varepsilon}$ has been defined as,

$$\widetilde{\varepsilon} = \frac{|\mathbf{D}_n^{(2)} - \widetilde{\mathbf{D}}_n^{(2)}|}{|\mathbf{E}_n^{(2)}|}. \quad (12)$$

The proposed algorithm calculates the error $\widetilde{\varepsilon}$ in N_e points and terminates if $\widetilde{\varepsilon} \le \varepsilon$. If the error condition is not met, extra N_a points are added via the integration routine to the previously defined set. An interpolant function is determined on this new grid of $N_a + N_a$ points and the error $\widetilde{\varepsilon}$ is subsequently calculated. Until the error condition is met, the algorithm keeps adding N_a points. Finally, the approximated value of the radiated field is computed as follows,

$$\widetilde{\mathbf{E}}_n^{(2)} = \widetilde{\mathbf{D}}_n^{(2)} + \mathbf{E}^{(p)}. \quad (13)$$

It is worth mentioning that the computational efficiency of the proposed algorithm is strictly related to the number N_a and on the termination condition (i.e., the choice of the N_e points). Based on our numerical experience, we suggest N_a to be in the order of ten and $1 \le N_e \le 4$. The N_e points are chosen in anticipation of the subsequent N_a points by the implemented algorithm.

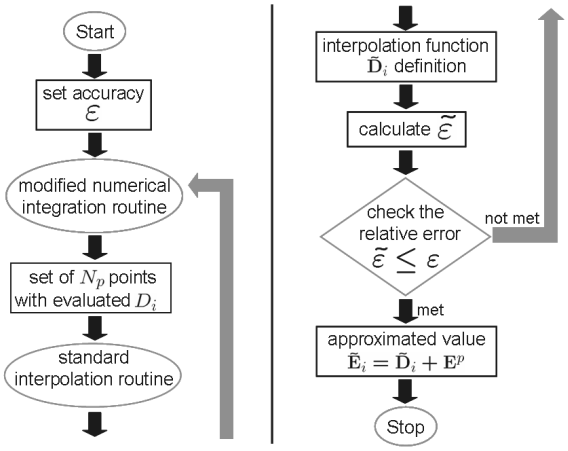

Fig. 3. Flow chart of the proposed numerical interpolation algorithm.

IV. NUMERICAL RESULTS

To validate the method, three representative examples containing mutually coupled wires are given. All simulations were performed on an Intel Xeon platform running at 2 GHz with 3 GB of RAM. First, the case of two parallel thin wires of length $L = \lambda/2$ and radius $a = \lambda/1000$, placed at a distance d, is considered. Both wires are subdivided in $N+1$ segments and N rooftop functions are defined. A desired accuracy of $\varepsilon = 10^{-3}$ is used. The computational time required for evaluating the system matrix $\underline{\underline{\mathbf{Z}}}$ by the straightforward double 2D integration and by the interpolation method is compared by varying the distance d and the number of expansion functions N. As can be observed in Table 1, the interpolation method greatly reduces the CPU time needed to calculate the impedance matrix even in case of a coarse discretization, (e.g. $N = 10$). When the two wires are close to each other (e.g. $d \le 0.1\lambda$) the computational time reduction can be appreciated only by refining the segmentation. As a matter of fact at small distances the function difference $\mathbf{D}_n$ may not be as smooth as when the distances are larger since the source field $\mathbf{E}_n$ differs from the point dipole field $\mathbf{E}^{(p)}$. Moreover, the maximum relative error ε_c on coupling elements is defined as,

$$\varepsilon_c = \max\left\{\frac{|C_{m,n} - \widetilde{C}_{m,n}|}{|C_{m,n}|}\right\}_{m,n=1}^{N} \quad (14)$$

and evaluated as the distance d and the number of expansion functions N vary (see Table 1). Coupling value $C_{m,n}$ represents the value computed with the integration method while $\widetilde{C}_{m,n}$ is computed with the proposed interpolation method.

The number of field evaluations using the straight forward 2D integration and using the interpolation method is also analyzed. Fig. 4 shows how the number of evaluations N_p required for the computation of

Table 1. Computational cost analysis and maximum relative error on coupling elements of two parallel thin wires $ka \approx 0.006$ by varying the distance d and the number of expansion functions N. Desired accuracy $\varepsilon = 10^{-3}$.

d [λ]	CPU Time ratio (2 × 2D Int) : (Interp)		
	$N = 68$	$N = 34$	$N = 10$
1	10.12 : 1	10.09 : 1	7.00 : 1
0.5	10.14 : 1	9.34 : 1	5.00 : 1
0.3	8.72 : 1	7.45 : 1	3.28 : 1
0.1	5.31 : 1	1.32 : 1	0.046 : 1

d [λ]	$\varepsilon_c = \max\left\{\frac{\lvert C_{m,n} - \widetilde{C}_{m,n}\rvert}{\lvert C_{m,n}\rvert}\right\}_{m,n=1}^{N}$		
	$N = 68$	$N = 34$	$N = 10$
1	$4.958 \cdot 10^{-13}$	$2.000 \cdot 10^{-11}$	$1.952 \cdot 10^{-9}$
0.5	$4.058 \cdot 10^{-12}$	$1.673 \cdot 10^{-10}$	$1.221 \cdot 10^{-8}$
0.3	$7.767 \cdot 10^{-11}$	$3.094 \cdot 10^{-9}$	$1.662 \cdot 10^{-6}$
0.1	$2.114 \cdot 10^{-6}$	$4.290 \cdot 10^{-5}$	$9.628 \cdot 10^{-5}$

coupling matrix $\underline{\underline{\mathbf{C}}}^{1,2}$ varies as a function of the desired accuracy ε for the configuration depicted above of two parallel thin wires at a distance $d/\lambda = 0.5$ and discretized with 35 segments. Fig. 4 shows that the total number of evaluations for the interpolation method is usually far less than for the integration, resulting in a considerable reduction of computation time. Fig. 4 also shows that increasing the accuracy in computing the electric field corresponds to an increase in the number of field evaluations N_p. This increment is much smaller when the double integration is carried out than when the interpolation is applied.

Second, we compare the computational time together with the relative error ε_c on coupling elements $C_{m,n}$ for the case of two parallel mutually coupled thick wires. In this case $L = \lambda/2$, the radius $a = \lambda/60$, and the wires are equally discretized with $N + 1$ segments. A desired accuracy of $\varepsilon = 10^{-3}$ is defined. Table 2 again shows how the proposed method enhances efficiency in the computation of coupling matrix elements even for thick wires.

Finally, two arbitrary oriented thin wires with $L_1 = \lambda/2$, $L_2 = 0.2236\lambda$ and radius $a_1 = a_2 = \lambda/1000$ are analyzed. The first wire has end points $\mathbf{r}_1 = (0.3, 0.3, 0.5)\lambda$ and $\mathbf{r}_2 = (0.3, 0.3, 1.0)\lambda$, while the second wire has end points $\mathbf{r}_3 = (0.3, 0.7, 0.5)\lambda$ and $\mathbf{r}_4 = (0.3, 0.8, 0.7)\lambda$. Again the CPU time comparison is carried out for different segmentations with a desired $\varepsilon = 10^{-3}$. Note that minimum and maximum distances between the two wires are $d_{\min} = 0.4\lambda$, $d_{\max} = 0.64\lambda$, respectively. Results for this case are shown in Table 3.

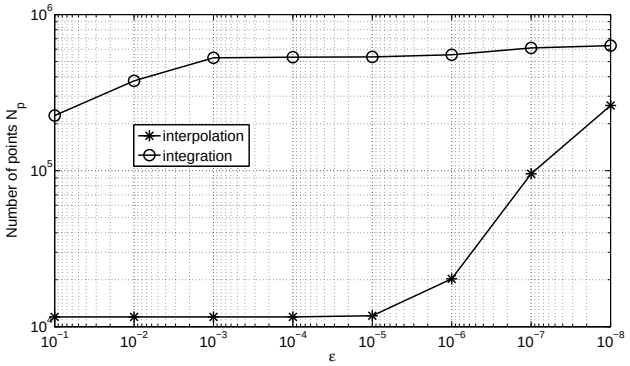

Fig. 4. Number of evaluations N_p versus the desired accuracy ε for two parallel thin wires of length $L = \lambda/2$ and radius $a = \lambda/1000$ placed at a distance $d/\lambda = 0.5$ and discretized with 35 segments.

Table 2. Computational cost analysis and maximum relative error on coupling elements of two parallel thick wires $ka \approx 0.105$ by varying the distance d and the number of expansion functions N. Desired accuracy $\varepsilon = 10^{-3}$.

d [λ]	CPU Time ratio (2 × 2D Int) : (Interp)	
	$N = 50$	$N = 20$
1	10.13 : 1	9.29 : 1
0.5	10.11 : 1	8.49 : 1
0.3	8.37 : 1	8.04 : 1
0.1	0.12 : 1	0.18 : 1

d [λ]	$\varepsilon_c = \max\left\{\frac{\lvert C_{m,n} - \widetilde{C}_{m,n}\rvert}{\lvert C_{m,n}\rvert}\right\}_{m,n=1}^{N}$	
	$N = 50$	$N = 20$
1	$2.726 \cdot 10^{-9}$	$1.086 \cdot 10^{-9}$
0.5	$1.141 \cdot 10^{-8}$	$1.383 \cdot 10^{-9}$
0.3	$1.043 \cdot 10^{-7}$	$1.920 \cdot 10^{-8}$
0.1	$2.396 \cdot 10^{-5}$	$5.201 \cdot 10^{-5}$

Table 3. Computational cost analysis of two arbitrarily oriented thin wires $ka \approx 0.006$, by varying the number of expansion functions N. Desired accuracy $\varepsilon = 10^{-3}$.

N	CPU Time ratio (2 × 2D Int) : (Interp)
68	10.83 : 1
34	10.83 : 1
10	7.07 : 1

V. CONCLUSION

A numerically efficient interpolation algorithm for the calculation of the coupling matrix elements has been

presented. Within a prescribed accuracy, coupling matrix elements can be computed more efficiently than by straightforward double integrations. This method is developed based on three main ideas elucidated in Sec. III-A.

Numerical examples show that the proposed algorithm results in a reduction of the CPU time of typically a factor 7. When the distance between the source and the observation element is small compared to the wavelength (e.g. $d \leq 0.1\lambda$) the smoothness of the function difference deteriorates and the advantages of the algorithm are lost. This is mainly due to the different behavior of the point dipole field and the field radiated by a source element in the near-field region. In this case the straightforward double integration may be more efficient than the interpolation technique.

When designing antennas to be installed on conducting surfaces, optimization routines can be used to properly select the best configuration. In this case, the proposed basic concept can be used to gain efficiency in the optimization process, drastically reducing the filling time of the wire-surface coupling matrices. Research towards the feasibility of this application is pursued.

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