Accurate Computation of Wide-Band Response of Electromagnetic Systems Utilizing Narrow-Band Information

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***Abstract*—We investigate Cauchy's technique for interpolating a rational function from samples of frequency responses and/or their derivatives. This technique can be used to speed up the numerical computations of parameters including input impedance and RCS of any linear time-invariant electromagnetic system. Here we have applied the technique to find the electromagnetic response of a conducting cylinder over a spectrum of frequency. The numerical results presented are in good agreement with exact computational data. This technique is a true interpolation/extrapolation technique as it provides the same defective result as the original electric field integral equation at a frequency which corresponds to the internal resonance of the closed structure.**

I. Introduction

It is often required to generate the broad-band response of electromagnetic systems. The conventional approach would be to solve the problem of interest at a set of discrete frequencies utilizing the method of moments. If the electromagnetic response has very sharp narrow-band responses then it becomes quite inefficient to utilize the method of moments at a large number of points with small frequency steps.

In this paper, it is proposed to utilize the concept of analytic continuation to extrapolate the broad-band response from narrow-band data. This is accomplished by assuming that the currents on the conductor can be approximated by a rational polynomial. The objective now is to determine the coefficients of the numerator and the denominator polynomial. If, somehow, the coefficients can be determined then broad-band information can be generated. The currents on the electromagnetic structure are computed at a few frequency points and the first four derivatives for the current are also evaluated at the same frequency. These frequency samples need not be equally spaced. Cauchy's technique [1] is employed next to determine the coefficients of the numerator and the denominator polynomial from the information about the current and its derivatives at a few frequency samples.

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Cauchy's technique involves extrapolating a function utilizing the values of the function and its derivatives at unequally spaced points. The Pade approximation, or frequency-domain Prony method, as it is often called, may also be utilized for extrapolation, as has been done in [5] and [7] over a limited range. In signal processing literature this is sometimes referred to as model-based parameter estimation, when the function values are utilized. For measured data it is more advantageous to use the function values only, as the derivative information is difficult to obtain accurately.

This technique is utilized to find the far field of a slit conducting cylinder (TM incidence) over a bandwidth utilizing the information about the current and its derivatives at a few sample points. The second section presents the basic principles and explains how to utilize Cauchy's technique to extrapolate a function by a rational polynomial. In the third section the problem of interest is described. The fourth section presents typical numerical results.

II. Basic Principles

In this paper the currents induced on electromagnetic systems are characterized by a rational transfer function. A rational function $H(s)$ is described as

$$H(s)=\frac{A(s)}{B(s)} \tag{1}$$

where

$$A(s)=\sum_{k=0}^{k=p} a_k s^k \tag{2}$$

$$B(s)=\sum_{k=0}^{k=q} b_k s^k \tag{3}$$

with $a_0=1$. The choice of $a_0=1$ is arbitrary as the ratio $H(s)$ can be scaled by a scalar parameter without changing its value. $H^{(n)}(s_j)$ denotes the nth-order derivative of $H(s)$ at $s=s_j$, and s is the frequency. It is important to note that the frequency points need not be equally spaced. Cauchy's problem is, given $H^{(n)}(s_j)$ for $j=1,2,\cdots,J$ and $n=0,1,2,\cdots,N_j$, to find $A(s)$ and $B(s)$.

A. Cauchy's Technique

We know that both $A(s)$ and $B(s)$ are polynomials but we do not know their orders (p and q) and part of the solution

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procedure is to determine them. It is, however, reasonable to assume that p and q are bounded by some known integers, i.e.,

$$p \leqslant P \tag{4}$$

$$q \leqslant Q. \tag{5}$$

According to the above assumptions, we redefine the polynomials as

$$A_1(s) = \sum_{k=0}^{k=P} a_{1k} s^k \tag{6}$$

$$B_1(s) = \sum_{k=0}^{k=Q} b_{1k} s^k \tag{7}$$

$$a_{10} = 1$$

and try fitting the ratio

$$H_1(s) = \frac{A_1(s)}{B_1(s)} \tag{8}$$

onto the data $H^{(n)}(s_j)$ such that

$$H_1^{(n)}(s_j) = H^{(n)}(s_j) \tag{9}$$

for all j and n, where $H^{(n)}(s_j)$ are the given samples of the function and their derivatives.

B. Uniqueness

For the above technique it can be shown that if the total number of samples is larger than or equal to the total number of known coefficients, $(P+Q+1)$ i.e.,

$$N = \sum_{j=1}^{j=J} (N_j + 1) \geqslant P + Q + 1 \tag{10}$$

then $H_1(s)$ is unique and $H_1(s) = H(s)$.

To show the uniqueness, assume $\tilde{H}(s)$ is another rational function of order (P, Q) satisfying

$$\tilde{H}^{(n)}(s_j) = H^{(n)}(s_j) \tag{11}$$

for all j and n. This means that $E(s) = H_1(s) - \tilde{H}(s)$ satisfies

$$E^{(n)}(s_j) = 0 \tag{12}$$

for all j and n. Equation (12) suggests that $E(s)$ is either zero, or it has zeros at the complex frequency (s) with order larger than or equal to $N_j + 1$. In the latter case, $E(s)$ has zeros of total order no less than $\Sigma_{j=1}^{j=J}(N_j+1) = N$. But the numerator of $E(s) = H_1(s) - \tilde{H}(s)$ cannot have order higher than $P+Q$. Thus, owing to the condition in (10), $E(s)$ must be equal to zero.

C. Linear Problem

From (8) we can infer that $A_1(s) = H_1(s)B_1(s)$ whenever $B_1(s)$ is nonzero. If the denominator $B_1(s)$ is not zero at any s_j, then (9) together with (8) is equivalent to the linear equation

$$A_1^{(n)}(s_j) = \sum_{i=0}^{i=n} C_{n,i} H^{(n-i)}(s_j) B_1^{(i)}(s_j) \tag{13}$$

for all j and n, where $C_{n,i} = n!/[i!(n-i)!]$. The above equation can be easily proved from (8) by mathematical induction.

It is clearly seen that $A(s)$ and $B(s)$ are a solution of the minimum order (p, q) to (13) assuming that the denominator polynomial is always nonzero. Since the ratio $A_1(s)/B_1(s)$ must be unique, the general solution, $A_1(s)$ and $B_1(s)$, to (13) can be written as

$$A_1(s) = A(s)D(s) \tag{14}$$

$$B_1(s) = B(s)D(s) \tag{15}$$

where $D(s)$ can be any polynomial of order no higher than

$$d = \text{minimum}\{P - p, Q - q\}. \tag{16}$$

It is seen that (14) and (15) are valid whether or not $D(s)$ at any s is zero. Since $D(s)$ has $d+1$ independent parameters, the solution of (13) has dimension $d+1$.

D. Matrix Equation

It is very useful for computational purposes to write (13) in matrix form. This is done by substituting (6) and (7) into (13), which results in the following equation:

$$\sum_{k=0}^{k=P} A_{1,j,n,k} a_{1,k} = \sum_{k=0}^{k=Q} B_{1,j,n,k} b_{1,k} \tag{17}$$

where

$$A_{1,j,n,k} = \frac{k!}{(k-n)!} s_j^{(k-n)} u(k-n) \tag{18}$$

$$B_{1,j,n,k} = \sum_{i=0}^{i=n} C_{n,i} H^{(n-i)}(s_j) \frac{k!}{(k-i)!} s_j^{(k-i)} u(k-i) \tag{19}$$

where $j = 1, 2, \cdots, J$ and $n = 0, 1, 2, \cdots, N_j$; $u(k)$ is zero for $k < 0$ and one otherwise (step function). We define

$$A_1 = \begin{bmatrix} & \cdots & \\ & \cdots & \\ A_{1,j,n,0} A_{1,j,n,1} & \cdots & A_{1,j,n,P} \\ & \cdots & \\ & \cdots & \end{bmatrix} \tag{20}$$

$$B_1 = \begin{bmatrix} & \cdots & \\ & \cdots & \\ B_{1,j,n,0} B_{1,j,n,1} & \cdots & B_{1,j,n,Q} \\ & \cdots & \\ & \cdots & \end{bmatrix}. \tag{21}$$

The order of the matrix A_1 is $N \times (P+1)$ and that of the matrix B_1 is $N \times (Q+1)$:

$$a_1 = [a_{10} a_{11} a_{12} \cdots a_{1P}]^T \tag{22}$$

$$b_1 = [b_{10} b_{11} b_{12} \cdots b_{1Q}]^T. \tag{23}$$

Then (17) becomes

$$[A_1, -B_1] \begin{bmatrix} a_1 \\ b_1 \end{bmatrix} = 0. \tag{24}$$

It is clear that the solution space of the above equation has dimension $d+1$, and the rank of the first matrix is $P+Q+2-(d+1)$. Therefore, d can be estimated from the singular values of that matrix.

E. Solutions

After obtaining the value of d, we may replace P by $P-d$ and Q by $Q-d$ in (17)–(24); then the solution to (24) is unique within a scalar, i.e.,

$$a_1 = d_0\left[a_0 a_1 \cdots a_p 0 \cdots 0\right]^T \quad (25)$$

$$b_1 = d_0\left[b_0 b_1 \cdots b_q\right]^T \quad (26)$$

for $P-p>Q-q$.

The other way of finding the a's, b's, p, and q (i.e., $A(s)$ and $B(s)$) is to choose a vector (e.g. the minimum normal vector or the vector corresponding to the smallest singular value) from the solution of (24) to form $A_1(s)$ and $B_1(s)$. This approach is better than the previous approach if the data of the transfer function are imprecise. For experimental data it may be more useful to use the functional values rather than the function and its derivatives. This is the approach taken in the generalized pencil of function method [2]. If we use high-order polynomials $A_1(s)$ and $B_1(s)$ to match the noisy samples, $H^{(n)}(s_j)$, then from (13), (17), or (24), we get

$$A_1(s) = A'(s)D_1'(s) \quad (27)$$

$$B_1(s) = B'(s)D_2'(s) \quad (28)$$

where $A'(s)$ and $B'(s)$ are noise-perturbed versions of $A(s)$ and $B(s)$ respectively, and $D_1'(s)$ and $D_2'(s)$ are noisy versions of $D(s)$. Since $D_1'(s)$ and $D_2'(s)$ both model some amount of noise, both $A'(s)$ and $B'(s)$ are less noise perturbed than the case where $d=0$. It is important to note that the orders P and Q cannot be increased arbitrarily. They must satisfy the condition given by (10).

Thus in this section an elaborate description of the Cauchy's technique and the possible solution techniques have been given. In the next section the implementation of the method of moments to solve for the current at a single frequency is described.

III. Method of Moments

In this section we shall discuss briefly how the method of moments is used to determine the currents on the conducting cylinder and the field scattered by the cylinder. The cylinder is of infinite extent with no variations of the fields or currents along the z axis. The system is illuminated by a TM plane wave. The surface equivalence principle is used to replace the cylinder by surface currents which are considered to be sources of the scattered field. These currents are computed by the method of moments.

The scattered field is produced by surface currents J. In order to have zero field, the scattered field must cancel with the incident field. This is possible if

$$E^s_{\text{ext}}(J)_{\tan} = -E^i_{\tan} \qquad \text{on } S^-. \quad (29)$$

The subscript "tan" indicates the tangential component of the field; "ext" indicates that the scattered fields are produced by the surface currents in an unbounded medium with (ϵ_0, μ_0).

The above E-field equation is solved numerically by the method of moments [3]. We use the pulse expansion and point matching technique to solve the above problem. The expansion function chosen here is the pulse expansion.

From [4] we have the following equation:

$$E_z = -j\omega\mu A_z$$

$$z\cdot\left[-j\omega A_{\text{ext}}(J)\right] = -E^i_z \qquad \text{on } S$$

where S refers to the surface just inside S and

E_z = z-directed incident electric field,
A = magnetic vector potential.

$$A(\rho) = \frac{\mu_i}{4j}\int_{c_j} J(\rho')H_0^{(2)}(k_i|\rho-\rho'|)\,dl'$$

ρ = position vector the field point.
ρ' = position vector the source point.
c_j = contour on which J is present.
$k_i = \omega(\mu_i\epsilon_i)^{1/2}$
$H_0^{(2)}$ = zeroth-order Hankel function of second kind.

We now approximate the surface S by a number of planar zones. In the xy plane the contour is divided into linear segments. We assume the unknown current on each segment to be a constant:

$$J(\rho') = z\sum_{l=1}^{l=N} P_l(\rho')$$

where N is the number of linear segments on S, and I_l is the unknown value of the electric current on the lth segment. P is the pulse function. $P_l(\rho)$ equals unity if ρ is the position vector of a point on the lth segment of S. Thus we get the following equation:

$$\frac{-\omega\mu_0}{4}\sum_{m=1}^{m=N} I_m\int_{c_{j,l}} H_0^{(2)}(k_0|\rho-\rho'|)\,dl' = -E^i_z(\rho) \qquad \text{on } S^-.$$

We satisfy the equation at the middle of each segment of S. The reason for doing this is that we know in the TM case the current flow is in the z direction and thus the current must be singular at the edges. Thus if we use point matching at the center of each segment the fields need not be evaluated at the edges. It has been found that when the match points are at the edges there is anomalous behavior in the current distribution. Thus one should always avoid computing the fields where the currents are known to be singular.

Writing the above equation in matrix form, we get

$$[Z][I]=[V]$$

where Z is the impedance, which is a square matrix of $N\times N$, and the current vector transpose is given as

$$[I]^T = [I_1 I_2 \cdots I_N]$$

where T denotes the transpose. Solving for the equations, we get the expansion coefficients and after that we can easily determine the electric field at any external point. The expressions for $Z_{m,n}$ and V_m are known explicitly as shown below:

$$Z_{mn} = \Delta C_m l_{mn}$$

$$V_m = \Delta C_m e^{jk(x\cos\phi_i + y\sin\phi_i)}$$

where ΔC_m = length of each segment and

$$l_{mn} = \frac{\eta}{4}k\Delta C_n H_0^{(2)}\left[k\sqrt{(x_n-x_m)^2+(y_n-y_m)^2}\right].$$

The above equations are for the impedance matrix and the voltage matrix. The derivatives for the current can be evaluated numerically from the above analytic expressions.

The derivatives for the voltage and impedance matrices are shown below:

$$V_m^{(t)} = \left[j(x\cos\phi_i + y\sin\phi_i) \right]^t \Delta C_m e^{jk(x\cos\phi_i + y\sin\phi_i)}$$

$$Z_{mn} = CZ = C1*C2*(CF1+CF2)$$

where

$$C1*C2 = \frac{\Delta C_m k\eta}{4}$$

$$CF1+CF2 = \int_{\Delta C_n} H_0^{(2)}(|\vec{r} - \vec{r}_m|k)\, dl'$$

$$CZ1 = \frac{dZ_{mn}}{dk} = \frac{CZ}{k} - C1*C2*(CF11+CF12)$$

where

$$CF11+CF12 = \int_{\Delta C_n} H_1^{(2)}(|\vec{r} - \vec{r}_m|k)\, dl' |\vec{r} - \vec{r}_m|$$

$$CZ2 = \frac{CZ1}{k} - \frac{CZ}{k^2} - C1*C2*(CF21+CF22)$$

where

$$CF21+CF22 = \int_{\Delta C_n} H_0^{(2)}(|\vec{r} - \vec{r}_m|k)\, dl' |\vec{r} - \vec{r}_m|^2$$

$$CZ3 = \frac{2CZ2}{k} - \frac{3CZ1}{k^2} + \frac{3CZ}{k^3} + C1*C2*(CF31+CF32)$$

where

$$CF31+CF32 = \int_{\Delta C_n} H_1^{(2)}(|\vec{r} - \vec{r}_m|k)\, dl' |\vec{r} - \vec{r}_m|^3.$$

In the above equation $V_m^{(t)}$ is the tth derivative of the voltage matrix with respect to frequency. In the above equations $CZ1$, $CZ2$, and $CZ3$ are the first, second, and third derivatives of the impedance matrix with respect to k. Solving for the integral equations, we get the expansion coefficients and after that we can easily determine the currents and thus the electric field at any external point by knowing the currents. The scattered fields also can be determined by knowing the currents.

IV. Examples

As a first example consider a slit cylinder as shown in Fig. 1. This problem has been solved utilizing the method of moments [6], and the computed solution over the band $ka = 1$ to $ka = 6$ for a 10° cut in the cylinder is shown in Fig. 1. In the first example the electric field is interpolated between $ka = 2$ to $ka = 4$ by knowing the current of the structure and its first four derivatives at three points, namely at $ka = 2$, 3, and 4. The cylinder is divided into 140 segments; the numerator order is 6 and the denominator order is 7. The method of moments is utilized to find the current distribution on the structure at $ka = 2$, 3, and 4. The first four derivatives of the current at $ka = 2$, 3, and 4 are computed by analytically differentiating the "method of moment impedance matrix" and utilizing the procedure outlined in [7]. Both the exact result and the interpolated results are shown in Fig. 2.

As a second example we consider the same slit cylinder and interpolate the far field between $ka = 2$ to $ka = 5.2$ by obtaining the current of the structure and four derivatives at four points, $ka = 2$, 3, 4, and 5. The cylinder is divided into 140 subsections, the numerator model order is 8, and the denominator order is 9. Both the exact and the interpolated results are shown in Fig. 3. As seen in the figure the three resonance points are obtained exactly.

In the third example we obtain the current and its four derivatives using the method of moments at $ka = 2$, 3, 4, 5, and 5.3. The cylinder is divided into 140 segments, the numerator model order is 10, and the denominator order is 11. In Fig. 4, in which we have overlaid the far field using both the exact and the Cauchy technique, we see that the resonance points are obtained exactly at the same position in both cases. The magnitudes of the resonances are truncated

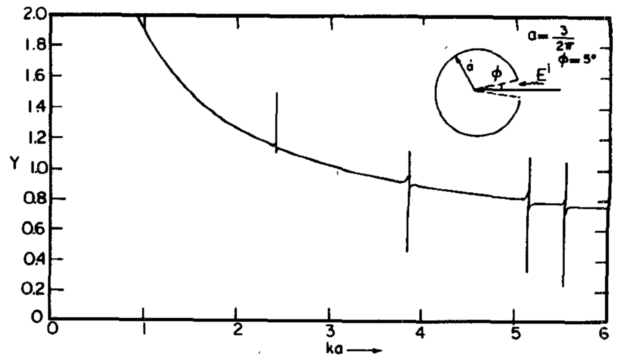

Fig. 1. Scattered far field of a conducting (slit) cylinder using exact method. Vertical axis: V/m.

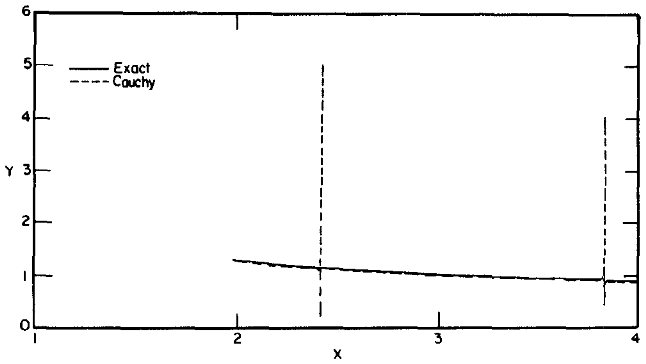

Fig. 2. Scattered far field of a conducting (slit) cylinder using Cauchy's technique (3 points). Vertical axis: V/m.

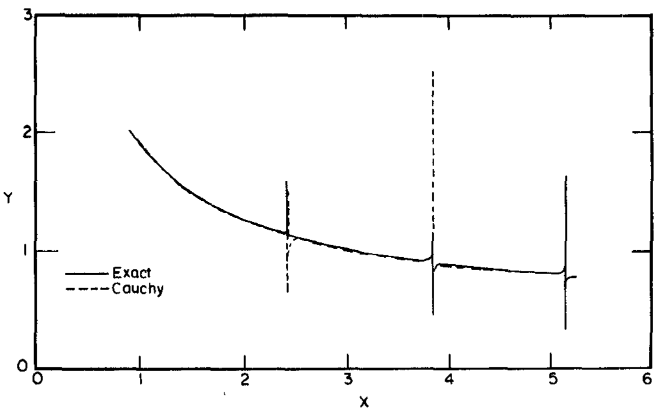

Fig. 3. Scattered far field of a conducting (slit) cylinder using Cauchy's technique (4 points). Vertical axis: V/m.

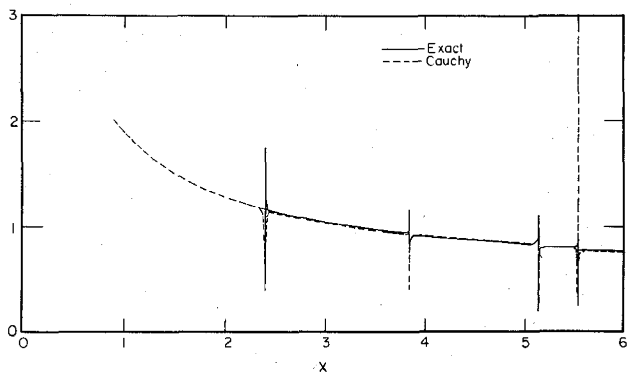

Fig. 4. Scattered far field of a conducting (slit) cylinder using Cauchy's technique (5 points). Vertical axis: V/m.

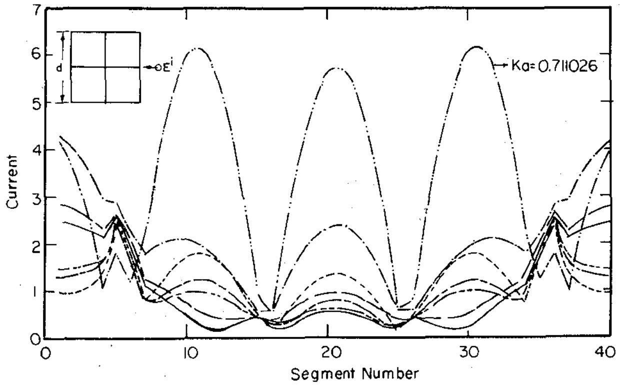

Fig. 5. Currents on a square cylinder using exact method. Vertical axis: A/m.

so that all the resonance points are seen clearly. Thus, by knowing the current and its four derivatives at five frequency points it is possible to extrapolate the fields over a decade from $ka = 0.6$ to $ka = 6.0$.

Next the square cylinder shown in Fig. 5 is considered. Here ten pulse functions have been utilized to expand the current on each face of the square cylinder. The currents are plotted as a function of frequency using the frequency derivative technique. It is seen that the currents responsible for the scattered fields have been masked by the resonant currents. The current breaks down at a value of $ka = 0.711026$, where a is the side of the square. This corresponds to the frequency caused by the internal resonance of the structure.

The theoretical resonance is supposed to occur at $\lambda/\sqrt{2}$, where $\lambda = 1.0$, but because of the approximations made by the method of moments, the resonance occurs at a slightly shifted value. In Fig. 5 the currents are plotted versus the segment number using the exact method. The square cylinder has been divided into 40 segments in the anticlockwise direction with segment number 1 starting from the point where the electric field has been incident, i.e., $x = 0$. Fig. 6 is the same problem as in Fig. 5 except that here the currents have been computed using the frequency derivative method. The curve numbers indicate the currents obtained at different values of ka. For example, the curve with number 4 is the curve obtained at $ka = 0.711026$. It is seen that the frequency derivative technique also shows a similar breakdown. This proves that the principle of analytic continuation

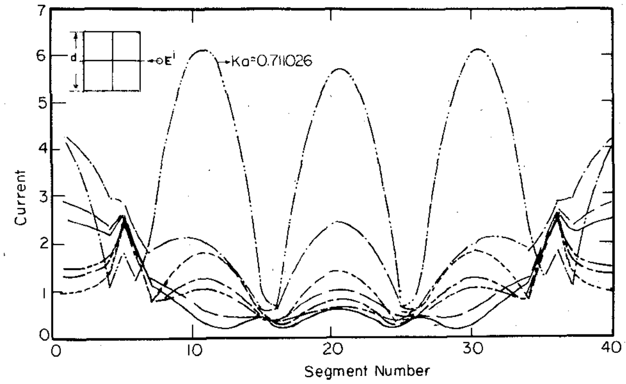

Fig. 6. Currents on a square cylinder using frequency derivative method. Vertical axis: A/m.

has not been violated, as the MOM solution also breaks down at the same value of ka.

V. Conclusions

The electromagnetic response of a conducting cylinder has been found over a spectrum of frequency using the Cauchy technique. The method of moments has been used to compute the current and its derivatives at a few frequency points using the pulse expansion and point matching techniques. Then Cauchy technique is used to extrapolate the currents at other frequencies. The computational time taken by this approach is at each frequency point $O(N^2)$, as opposed to $O(N^3)$ by the conventional method of moments. It has been shown that this is a true interpolation/extrapolation technique in that it provides the same result as the electric field integral equation at a frequency which corresponds to the internal resonance of the structure.

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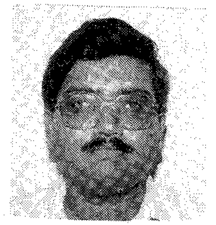

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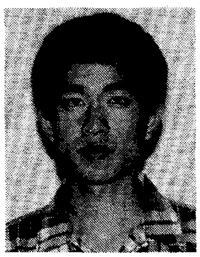

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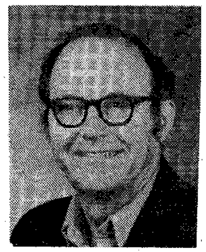
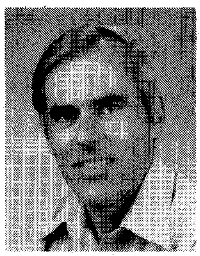
He was a Graduate Teaching Assistant from 1984 to 1985, a Graduate Fellow from 1985 to 1986, a Graduate Research Assistant from 1986 to 1988, and a Postdoctoral Research Associate from 1988 to 1989 at Syracuse University. Since February 1990, he has been a Lecturer with the University of Melbourne, Victoria, Australia. He has contributed over 30 publications dealing with spectral analysis and array processing.



Edmund K. Miller (S'60–M'66–SM'70–F'84) is Leader, Instrumentation Group, Mechanical and Electronic Engineering, Los Alamos National Laboratory. His interest in computational electromagnetics began during his Ph.D. research at the University of Michigan.

Subsequent positions with MBAssociates, 15 years at Lawrence Livermore National Laboratory, the University of Kansas, Rockwell International Science Center, and General Research Corporation have continued his involvement in numerical methods as well as expanding into signal processing and graphical applications. Combining signal processing and modeling as model-based parameter estimation is an area on which he has concentrated over the last few years. He has served as IEEE National Distinguished Lecturer for the Antennas and Propagation Society. He also writes a column on personal computers which is published in several IEEE society newsletters and is past president of the Applied Computational Electromagnetics Society.



Gerald J. Burke (S'68–M'68) was born in San Mateo, CA, on June 28, 1943. He received the B.S. degree in electrical engineering and the M.S. degree from the University of California, Berkeley, in 1965 and 1968, respectively.

Since 1975 he has been employed at the Lawrence Livermore National Laboratory, where he has worked on numerical modeling of antennas in the frequency and time domains and has developed methods for modeling antennas that are buried or penetrate into the ground.