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ON THE INTEGRAL EQUATIONS OF THIN WIRE ANTENNAS

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ON THE INTEGRAL EQUATIONS OF THIN WIRE ANTENNAS^{*} K. K. MEI^{**} MEMBER, IEEE

<u>Summary</u>— The feasibility of direct numerical calculations of antenna integral equations is investigated. It is shown that integral equation of Hallen's type is the most adequate for such applications. The extension of Hallen's integral equation to describe thin wire antennas of arbitrary geometry is accomplished, and results are presented for dipole, circular loops and equiangular spiral antennas.

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INTRODUCTION

During the past seven years, the advancement of antenna design has been characterized by an exhaustive utilization of antenna geometry. Broad-band antennas are notable examples. In the study of antenna theory, a knowledge of the current distribution is of fundamental importance. Such data may be obtained either by measurement or by solving the antenna integral equation. Integral equations are difficult to solve even for the simplest case of a dipole antenna, however, as a result of the development in modern high speed computers, the range of application of the integral equation method has been greatly enlarged. The purpose of this paper is to present an investigation of the feasibility of direct numerical calculations of antenna integral equations. To simplify the discussion, the trapezoidal rule of integration is assumed throughout, although it is realized that in a practical calculation better integration schemes, such as quadratic rule etc., may need to be used. Typical results of calculations are presented.

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NUMERICAL SOLUTIONS OF DIPOLE ANTENNAS

It is well-known that the axial component of the electric field produced by the current on a cylindrical dipole antennal is given by

$$\frac{\mathrm{d}}{\mathrm{d}z} \int_{\mathbf{L}} \oint \frac{\mathrm{d}J(\mathbf{z}')}{\mathrm{d}z'} G(\mathbf{z}, \mathbf{c}; \mathbf{z}'\mathbf{c}') \,\mathrm{d}\mathbf{c}' \,\mathrm{d}\mathbf{z}' + \mathbf{k}^2 \int_{\mathbf{L}} \oint J(\mathbf{z}') \,G(\mathbf{z}, \mathbf{c}; \mathbf{z}', \mathbf{c}') \,\mathrm{d}\mathbf{c}' \,\mathrm{d}\mathbf{z}'$$
(1)

 $= j\omega \in E(z)$

where the symbol J(z') represents the surface current density, $\oint dc'$ represents the integration around the periphery of the cylinder, and G(z, c; z', c') is the free space Green's function,

$$G(z,c;z',c') = \frac{e^{-jk} \left| \overline{r} - \overline{r'} \right|}{4\pi \left| \overline{r} - \overline{r'} \right|}$$

For simplicity we shall omit the integration $\oint dc'$ in the discussion that follows, i.e., the symbol $\int_{L} dz'$ will represent the surface integral over the cylinder.

When the electric field on the surface of the antenna is considered, (1) reduces to

$$\frac{\mathrm{d}}{\mathrm{d}z} \int_{\mathbf{L}} J'(z') G(z, z') \, \mathrm{d}z' + k^2 \int_{\mathbf{L}} J(z') G(z, z') \, \mathrm{d}z' = -j\omega \, \epsilon \, \mathbf{E}_z^{\mathbf{i}}(z) \tag{2}$$

where $E_z^i(z)$ is the electric field produced by the generator. Eq. (2) is an integrodifferential equation for the current, which may be solved numerically by a combination of the difference equation method and the numerical integration method. The disadvantage of such an approach is that difference equations are generally unstable and critical to the errors in the approximation. An alternative approach is to transform

¹R. King, "The Theory of Linear Antennas," Harvard University Press, Cambridge, Mass.; 1956.

(2) into a pure integral equation. Eq. (2) may be readily transormed into such an equation of one of several familiar forms. The one used by Pocklington² is

$$\int_{\mathbf{L}} J(\mathbf{z}') \left[\frac{\partial^2}{d\mathbf{z}^2} G(\mathbf{z}, \mathbf{z}') + \mathbf{k}^2 G(\mathbf{z}, \mathbf{z}') \right] d\mathbf{z}' = -j\omega \epsilon \mathbf{E}_{\mathbf{z}}^{\mathbf{i}}(\mathbf{z}) ; \qquad (3)$$

integrating both sides of (3), say from 0 to z, gives

$$\int_{\mathbf{L}} J(\mathbf{z}') \left[\frac{\partial}{\partial z} G(\mathbf{z}, \mathbf{z}') + \mathbf{k}^2 \int_0^{\mathbf{z}} G(\xi, \mathbf{z}') d\xi \right] d\mathbf{z}' = -j\omega \epsilon \int_0^{\mathbf{z}} \mathbf{E}_{\mathbf{z}}^{\mathbf{i}}(\xi) d\xi + A \quad . \quad (4)$$

The integral equation used by Hallen¹ is,

$$\int_{L} J(z') G(z, z') dz' = B \cos kz - \frac{jV}{2Z_0} \sin k |z| .$$
 (5)

In these integral equations, the constants of integration A and B are to be determined by the condition that the current vanishes at both ends of the antenna; V and Z, are respectively the voltage applied and the intrinisic impedance of free space.

The numerical solution of an integral equation may be effected by approximating the integration with a finite sum at n different points. The resulting algebraic equations will have the following form: $^{3, 4}$

²S. A. Schelkunoff, "Advanced Antenna Theory," p. 132, John Wiley and Sons, New York; 1952.

³F. B. Hildebrand, "Method of Applied Mathematics," pp. 448-451, Prentice-Hall, New York; 1954.

⁴K. K. Mei and J. G. Van Bladel, "Scattering by perfectly-conducting rectangular cylinders," <u>IEEE Trans.</u>, Vol. AP-11, No. 2, pp. 185-192; March 1963.

$$K_{11} J(z_1) + K_{12} J(z_2) + \dots + K_{1n} J(z_n) = F(z_1)$$

$$K_{21} J(z_1) + K_{22} J(z_2) + \dots + K_{2n} J(z_n) = F(z_2)$$
. (6)

$$K_{n1} J(z_1) + K_{n2} J(z_2) + \dots + K_{nn} J(z_n) = F(z_n)$$

The matrix elements K_{ij} and $F(z_i)$ for Eqs. (3), (4) and (5) are given respectively, as

$$K_{ij} = \int_{\Delta z_{j}} \left(\frac{\partial^{2}}{\partial z^{2}} + k^{2} \right) G(z_{i}, z') dz'$$

$$F(z_{i}) = -j\omega \epsilon E_{z}^{i}(z_{i}) ,$$

$$K_{ij} = \int_{\Delta z_{j}} \left[\frac{\partial}{\partial z} G(z_{i}, z') + k^{2} \int_{0}^{z_{i}} G(\xi, z') d\xi \right] dz'$$

$$F(Z_{i}) = -j\omega \epsilon \int_{0}^{z_{i}} E_{z}^{i}(\xi) d\xi + A ,$$

$$K_{ij} = \int_{\Delta z_{j}} G(z_{i}, z') dz'$$

$$(9)$$

$$F(z_i) = B \cos k z_i - \frac{jV}{2Z_0} \sin k |z_i| \int$$

where Δz_j 's, the subdivisions of the antenna, as shown in Fig. 1 are sufficiently small so that the current in each may be considered constant.

We notice that the integral in (7) does not converge at i = j. Whether the often used approximation for a thin antenna of radius <u>a</u>,

$$\oint_{C} G(z, c; z', c') dc' \approx 2\pi a \left\{ \frac{e^{-jk \left[(z - z')^{2} + a^{2} \right]^{1/2}}}{4\pi \left[(z - z')^{2} + a^{2} \right]^{1/2}} \right\}$$
(10)

can be applied in the divergent integral of (7) is open to question. $^{5-7}$ An inspection of (6), (7) and (10) indicates that such approximation will not lead to the correct solution. This is so because, if approximations (7) and (10) are used, in the limit of small radius <u>a</u> Eq. (6) approaches a diagonal matrix. That is to say, for a very thin antenna, the solution of (6) would then give $J(z) \alpha E_z^i(z)$, which is not compatible with the well founded knowledge of antenna current distributions.

The improper integrals in (8) and (9) at i = j may be integrated by using Cauchy's principal value. In these cases, we may also use the approximation (10). Actual computations based on such an approximation indeed give correct results. This possibly accounts for the fact that approximations (7) and (10) have been successfully used in variational form, ^{8, 9} since the variational formulation introduces an additional integration, which in effect suppresses the divergent nature of the integral.

⁶J. G. Van Bladel, "Some remarks on Green's dyadic for infinite space," IRE Trans., Vol. AP-9, No. 6, pp. 563-566; Nov. 1961.

⁷C. J. Bouwkamp, "Diffraction theory," Report on Progress in Physics, Vol. 17, pp. 35-100; 1954.

⁸C. T. Tai, "A new interpretation of the integral equation formulation of cylindrical antennas," <u>IRE Trans.</u>, Vol. AP-3, pp. 125-127; July 1955.

⁹C. H. Tan, "Input impedance of arc antennas and short helical radiators," IEEE Trans., Vol. AP-12, No. 1, pp. 2-9; Jan. 1964.

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⁵J. Aharoni, "Antennae - An Introduction to Their Theory," pp. 133-135, Clarendon Press, Oxford; 1946.

Of particular importance in the inversion of a large matrix is the problem of round-off errors accumulated through large number of arithmetic operations. In general, the round-off errors depend on the orientations of the hyperplanes represented by each row of the matrix, in the n-dimensional vector space. Qualitatively speaking, the round-off errors will be small if the hyperplanes are essentially perpendicular to one another, and the reverse is true if two or more of them are almost parallel.¹⁰ Inspection of (8) indicates that for small radius \underline{a} the coefficient K_{ij} will be small for i < j, and large for $i \ge j$. Hence, in the limit of a very thin antenna, the matrix elements described by (8) approach those of a triangular matrix. For the same situation, however, the matrix elements described by (9) approach those of a diagonal matrix, which is certainly superior to a triangular one in view of the above consideration on computational errors. We shall, therefore, use integral Eq. (5) as the basis of our calculations.

A few typical results of calculation on dipole antennas are shown in Figs. 2-4. It is of interest to note that calculations based on the model of a slice generator excitation, 1 and those based on the model of a magnetic loop current excitation¹¹ have no noticeable differences in their results.

ARBITRARY THIN WIRE ANTENNAS

The extension of Eqs. (3) and (4) to describe a general curved wire antenna is immediate, provided a curved cylindrical coordinate system is used. Fig. (5) describes such a coordinate system, where

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¹⁰S. H. Crandell, "Engineering Analysis," Sec. 1. 3, pp. 15-18, McGraw-Hill, New York; 1956.

¹¹G. E. Albert and J. L. Synge, "The general problem of antenna radiation and the fundamental integral equation with application to an antenna of revolution," Part 1, <u>Quart. Appl. Math.</u>, Vol. 6, pp. 117-132; July 1948.

<u>s</u> is the arc length measured from the feed gap, and \hat{s} is the unit tangent vector at s. If the radius <u>a</u> of the wire is sufficiently small so that the current density may be considered to be uniform around the periphery of the wire, the corresponding integral Eq. (3) and (4) for a curved wire antenna are respectively, ⁵

$$\int_{\mathbf{L}} J(s') \left[\frac{\partial^2}{\partial s \partial s'} G(s, s') - k^2 G(s, s') \hat{s} \cdot \hat{s}' \right] ds' = j\omega \epsilon \mathbf{E}_s^i(s) \qquad (11)$$

and

$$\int_{\mathbf{L}} J(\mathbf{s}') \left[\frac{\partial}{\partial \mathbf{s}'} G(\mathbf{s}, \mathbf{s}') - \mathbf{k}^2 \int_0^{\mathbf{s}} G(\xi, \mathbf{s}') \hat{\xi} \cdot \hat{s}' d\xi ds' \right] = j\omega \epsilon \int_0^{\mathbf{s}} E_{\mathbf{s}}^i(\xi) d\xi + \mathbf{A}.$$
(12)

The extension of Eq. (5) to describe a general curved wire antenna is not so apparent. The complication arises in that the kernel of the closed-cycle type is essential in the conventional way of deriving integral Eq. (5). Such a kernel has the special property,

$$\frac{\partial}{\partial s} K(s, s') = - \frac{\partial}{\partial s'} K(s, s')$$
 (13)

The structures which give rise to kernels of this type are limited to straight wires, circular arcs, and helical wires.⁹ In the following we shall attempt to generalize Eq. (5) so as to include wire antennas of arbitrary geometry.

In accord with the assumptions of a thin wire antenna, the tangential component of the vector potential and scalar potential on the antenna are given respectively as

$$\mathbf{A}_{\mathbf{s}}(\mathbf{s}) = \int_{\mathbf{L}} J(\mathbf{s}') G(\mathbf{\dot{s}}, \mathbf{\dot{s}}') \mathbf{\dot{s}} \cdot \mathbf{\dot{s}}' d\mathbf{s}'$$
(14)

and

$$\phi(s) = \frac{-1}{j\omega \epsilon} \int_{L} \frac{dJ(s')}{ds'} G(s, s') ds' . \qquad (15)$$

We define a scalar function $\Phi(s)$ as

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$$\Phi(s) = -j\omega\epsilon \int_0^s \phi(\xi) d\xi = \int_0^s \int_L \frac{dJ(s')}{ds'} G(\xi, s') d\xi .$$
 (16)

Integrating (16) by parts and considering J(s) to vanish at both ends, we obtain

$$\Phi(\mathbf{s}) = -\int_0^{\mathbf{s}} \int_{\mathbf{L}} J(\mathbf{s}') \frac{\partial G(\xi; \mathbf{s}')}{\partial \mathbf{s}'} d\xi . \qquad (17)$$

For the <u>s</u> component of the electric field on the antenna to vanish, it is required that

$$E_{s}(s) + E_{s}^{i}(s) = 0$$
 (18)

where $E_s^i(s)$ is the <u>s</u> component of the incident electric field when the antenna is receiving, or it is the impressed field of the source if the antenna is transmitting.

From the well-known equation

we have

$$k^{2} A_{s}(s) - j\omega \epsilon \frac{d\phi(s)}{ds} = -j\omega \epsilon E_{s}^{i}(s)$$
(19)

$$\frac{d^2 \Phi(s)}{ds^2} = -k^2 A_s(s) - j\omega \epsilon E_s^i(s) . \qquad (20)$$

Adding $k^2 \Phi(s)$ to both sides of (20), we obtain

$$\frac{d^{2}\Phi(s)}{ds^{2}} + k^{2}\Phi(s) = k^{2}(\Phi(s) - A_{s}(s)) - j\omega \epsilon E_{s}^{i}(s) . \qquad (21)$$

The solution of (21) is

$$\Phi(s) = C \cos ks + D \sin k |s| + \int_{0}^{s} k (\Phi(\xi) - A_{\xi}(\xi)) \sin k(s - \xi)$$

$$- \frac{j}{Z_{0}} \int_{0}^{s} E_{\xi}^{i} (\xi) \sin k(s - \xi) d\xi.$$
(22)

Since $\Phi(0) = 0$, we see that the constant C must vanish. Now consider the integration

$$F(s) = k \int_{0}^{s} \Phi(\xi) \sin k (s - \xi) d\xi$$
$$= -k \int_{0}^{s} \int_{0}^{\xi} \int_{L}^{J} (s') \frac{\partial G(\eta, s')}{\partial s'} ds' d\eta \sin k (s - \xi) d\xi .$$
(23)

After changing the order of integration in (23), we obtain

 \mathbf{or}

$$F(s) = -k \int_{L} \int_{0}^{s} \int_{\eta}^{s} J(s') \frac{\partial G(\eta, s')}{ds'} \sin k (s - \xi) d\xi d\eta ds'$$

$$= -\int_{L} \int_{0}^{s} J(s') \frac{\partial G(\eta, s')}{ds'} \left[1 - \cos k(s - \eta)\right] d\eta ds' \qquad (24)$$

$$= \Phi(s) + \int_{L} \int_{0}^{s} J(s') \frac{\partial G(\eta, s')}{ds'} \cos k (s - \eta) d\eta ds' .$$

Next we consider the integration

$$H(s) = k \int_{0}^{s} A_{\xi}(\xi) \sin k (s - \xi) d\xi$$

= $k \int_{0}^{s} \int_{L} J(s') G(\xi, s') \hat{\xi} \cdot \hat{s}' \sin k (s - \xi) d\xi ds'.$ (25)

Integration by parts gives

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$$H(s) = \int_{L} J(s') G(\xi, s') \dot{\xi} \cdot \dot{s}' \cos k (s - \xi) \left| \begin{array}{c} \xi = s \\ \xi = 0 \end{array} ds' \\ - \int_{0}^{s} \int_{L} \left[\frac{\partial G(\xi, s')}{\partial \xi} \dot{\xi} \cdot \dot{s}' + G(\xi, s') \frac{\partial (\dot{\xi} \cdot \dot{s}')}{\partial \xi} \right] J(s') \cos k (s - \xi) d\xi \\ = \int_{L} J(s') G(s, s') \dot{s} \cdot \dot{s}' ds' - \int_{L} J(s') G(0, s') \dot{0} \cdot \dot{s}' \cos ks ds' \\ - \int_{0}^{s} \int_{L} \left[\frac{\partial G(\xi, s')}{\partial \xi} \dot{\xi} \cdot \dot{s}' + G(\xi, s') \frac{\partial (\dot{\xi} \cdot \dot{s}')}{\partial \xi} \right] J(s') \cos k(s - \xi) d\xi .$$

$$(26)$$

Substituting (24) and (26) into (22), we obtain the integral equation for the current,

$$\int_{L} J(s') \pi(s, s') ds' = D \sin k |s| + \int_{L} J(s') G(0, s') \stackrel{0}{0} \cdot \hat{s}' \cos k s ds' - \frac{j}{Z_0} \int_0^s E_{\xi}^i(\xi) \sin k(s - \xi) d\xi$$
(27)

where

$$\pi(\mathbf{s}, \mathbf{s}') = \mathbf{G}(\mathbf{s}, \mathbf{s}') \,\,\widehat{\mathbf{s}} \cdot \,\,\widehat{\mathbf{s}}' - \int_0^{\mathbf{s}} \left[\frac{\partial \mathbf{G}(\xi, \mathbf{s}')}{\partial \xi} \,\,\widehat{\boldsymbol{\xi}} \cdot \,\,\widehat{\mathbf{s}}' \right] \\ + \frac{\partial \mathbf{G}(\xi, \mathbf{s}')}{\partial \mathbf{s}'} + \mathbf{G}(\xi, \mathbf{s}') \,\,\frac{\partial(\widehat{\boldsymbol{\xi}} \cdot \,\,\widehat{\mathbf{s}}')}{\partial \xi} \,\, \cos \mathbf{k} \,(\mathbf{s} - \xi) \,\, \mathrm{d}\xi \,\,\,.$$

$$(28)$$

The term $D \sin k |s|$ represents the effect of a slice generator which is redundant when the integral of E_{ξ}^{i} is present. Indeed, if $E_{\xi}^{i}(\xi) = V/2 \,\delta(\xi)$, where $\delta(\xi)$ is the Dirac delta function, we have

$$\frac{-j}{Z_0} \int_0^s E_{\xi}^{i}(\xi) \sin k (s - \xi) d\xi = \frac{-jV}{2Z_0} \sin k |s|$$
(29)

which is consistant with (5).

To show that Eq. (27) reduces to (5) for a dipole antenna, we assume the source to be a slice generator, and notice that in this particular case

$$\frac{\partial G(\xi, s')}{\partial \xi} = - \frac{\partial G(\xi, s')}{\partial s'}$$

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and

$$\hat{\boldsymbol{\xi}} \cdot \hat{\boldsymbol{s}} = 1$$
.

Hence, (27) becomes

$$\int_{L} J(z') G(z, z') dz' = \int_{L} J(z') G(0, z') dz' \cos kz - \frac{jV}{2Z_0} \sin k |z| . \quad (30)$$

Comparing (30) with (5), we have

$$B = \int_{L} J(z') G(0, z') dz'$$

which may be shown to be correct by considering (5) at z = 0. Consequently, the term $\int_{L} J(z') G(0, z') dz'$ should be replaced by a constant, which has to be determined by the condition of the current at the ends of the antenna, otherwise the solution of the integral equation will not be unique. Therefore, the integral equation describing an arbitrary thin wire antenna is

$$\int_{L} J(s') \pi(s, s') ds' = C' \cos ks - \frac{j}{Z_0} \int_0^s E_{\xi}^{i}(\xi) \sin k(s - \xi) d\xi.$$
(31)

The specialization of (31) to a circular loop antenna also agrees with that derived by Adachi. $^{12}\,$

A further check of the integral equation may be effected as following. We differentiate (31) twice with respect to \underline{s} , and make use of the differential relation,

¹²S. Adachi and Y. Mushiake, "Theoretical formulation of circular loop antennas by integral equation method," <u>Sci. Rep.</u>, RITU, B-(Elect. Comm.) Vol. 9, No. 1, Sendi, Japan.

$$\frac{d}{dx} \int_{0}^{f(x)} g(x, x') dx' = \int_{0}^{f(x)} \frac{\partial}{\partial x} g(x, x') dx' + g(x, x') \bigg| \begin{array}{c} f'(x) & (32) \\ x' = x \end{array}$$

and obtain

$$k^{2} \int_{L} \int_{0}^{s} J(s') \left[\frac{\partial G(\xi, s')}{\partial \xi} \frac{\Lambda}{\xi} \cdot \frac{\Lambda}{s'} + \frac{\partial G(\xi, s')}{\partial s'} + G(\xi, s') \frac{\partial (\hat{\xi} \cdot \hat{s}')}{\partial \xi} \right]$$

$$\cos k (s - \xi) d\xi ds' - \frac{\partial}{\partial s} \int_{L} J(s') \frac{\partial G(s, s')}{\partial s'} ds'$$
(33)

$$= -k^2 C' \cos ks + j \frac{k^2}{Z_0} \int_0^s E_{\xi}^i(\xi) \sin k(s - \xi) d\xi - j\omega \epsilon E_{\xi}^i(s) .$$

Multiplying (31) by k^2 and adding the result to (33), results in

$$k^{2} \int_{L} J(s') G(s, s') \hat{s} \cdot \hat{s}' ds' - \frac{\partial}{\partial s} \int_{L} J(s') \frac{\partial G(s, s')}{\partial s'} ds = -j\omega \epsilon E_{s}^{i}(s) \quad (34)$$

which is essentially Eq. (19). Therefore, the integral (31) is shown to be the correct one.

APPLICATIONS

Eq. (31) has been applied to circular loop antennas¹³ and equiangular spiral antennas. The representative results are shown in Figs. 6 and 7. The calculations of current on log-periodic zig-zag antennas are in progress. We shall report the details of calculations in the near future when the numerical results are compiled.

¹³A. Baghdasarian and D. J. Angelakos, "Scattering and radiation from conducting loops," to be published as an ERL report, University of California.



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Fig. 1. Relevant of a dipole antenna and its subdivisions.



10, kL = $\pi/2$.







Fig. 4. Current distribution $I = I_R + jI_i$ on a dipole antenna of parameters $\Omega = 2 \log 2L/a = 10$, $kL = 5\pi/4$.



Fig. 5. A curved cylindrical coordinate system.



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Fig. 7. Current distribution of a 6\.arm equiangular spiral antenna.