netics in a Complex World: Challenges and Perspectives, University of Sannio, Benevento, Italy, February 20-21, 2003.

7. L. Sevgi, "Complex Electromagnetic Problems and Numerical Simulation Approaches," IEEE EMC Symposium, May 2003, Istanbul, Turkey.

8. David Morgan, *A Handbook for EMC Testing and Measurement*, London, Peter Peregrinus Ltd. (IEE Series), 1994.

9. N. İnce (ed.), *Modeling and Simulation Environment for Satellite and Terrestrial Communication Networks*, Kluwer Academic Publishers, 2001.

10. L. Sevgi, "On the Science, Scientific Process, and Scientific Filter," *IEEE Antennas and Propagation Magazine*, **44**, 2, April 2002, pp. 122.

Limitations of the Thévenin and Norton Equivalent Circuits for a Receiving Antenna

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The investigation carried out in this paper was stimulated by a recent paper published by Love, in which the appropriateness of the use of the Thévenin and Norton equivalent circuits for a receiving antenna was questioned [1]. A review of the available literature led to the conclusion that the limitations inherent in the Thévenin and Norton equivalent circuits had not been adequately examined, and this led to the investigation that is reported on in this paper. The Thévenin and Norton equivalent circuits are useful in the reduction of the equivalent circuit for a transmitting-receiving antenna system to simpler networks that facilitate the evaluation of the received power. One finds in the literature that the calculated power dissipation within these equivalent circuits is often equated to the reradiated and scattered power from the receiving antenna [2]. Such calculations are not correct, because power dissipation in the network from which the Thévenin and Norton equivalent circuits were obtained cannot be made using the Thévenin and Norton equivalent circuits. However, as we will show, the Thévenin and Norton equivalent circuits can be used to find a reradiated electromagnetic field that is a part of the total field scattered by a receiving antenna. As part of the derivation of this new result, we develop a derivation of the Thévenin and Norton equivalent circuits from the basic principles of uniqueness and superposition applied to electromagnetic fields.

Consider the representation of a transmitting and receiving antenna system as shown in Figure 1. The receiving antenna is terminated in a load impedance Z_L at its terminal plane, T_1, and the transmitting antenna has a generator with EMF V_g in series with an impedance, Z_g, connected at its terminal plane, T_2. Silver has shown that an equivalent circuit of the form shown in Figure 2 exists for any two antennas that are coupled to their respective source and load terminations by transmission lines or waveguides supporting single propagating modes, such that equivalent terminal voltages and currents can be defined on the terminal planes [3]. Power conservation for this transmitting-receiving antenna system can be established by integration of the real part of the Poynting-vector flux over the metal bounding surfaces of the two antennas, plus the terminal planes of the two antennas, and the surface of an

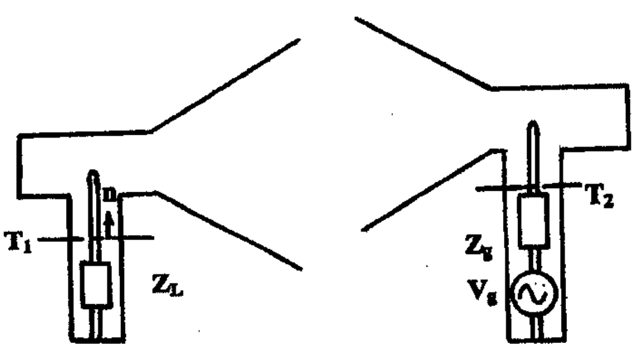

Figure 1. An arbitrary transmitting-receiving antenna system.

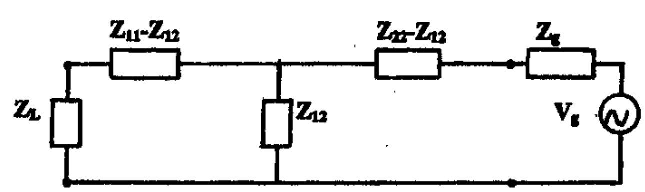

Figure 2. The equivalent network for the transmitting-receiving antenna system in Figure 1.

enclosing sphere with very large radius. This gives the power conservation statement

$$P_{in} = P_L + P_{LR} + P_{LT} + P_R, \quad (1)$$

where P_{in} is the transmitting-antenna input power at the terminal plane T_2, P_L is the power across the terminal plane T_1 into the receiving-antenna load impedance, P_{LR} is the power loss in the bounding metal surface of the receiving antenna, P_{LT} is the power loss in the bounding metal surface of the transmitting antenna, and P_R is the power radiated away to infinity. The latter consists of the radiated power from the transmitting antenna, the scattered power from the receiving antenna, and a power interaction term. The total field at infinity is the sum of the transmitting-antenna radiated field, $\mathbf{E}_t, \mathbf{H}_t$, and the receiving-antenna scattered field, $\mathbf{E}_s, \mathbf{H}_s$. Due to interference between these two fields, the total power radiated away includes a non-zero interaction term, i.e.,

$$\begin{aligned} P_R &= \frac{1}{2}\mathrm{Re}\oint_S (\mathbf{E}_t + \mathbf{E}_s)\times(\mathbf{H}_t + \mathbf{H}_s)^* \cdot \mathbf{a}_r dS \\ &= \frac{1}{2}\mathrm{Re}\oint_S \left[\mathbf{E}_t \times \mathbf{H}_t^* + \mathbf{E}_s \times \mathbf{H}_s^* + \left(\mathbf{E}_t \times \mathbf{H}_s^* + \mathbf{E}_s \times \mathbf{H}_t^*\right)\right] \cdot \mathbf{a}_r dS, \end{aligned} \quad (2)$$

where the last term is the interaction term. The power dissipated in the impedance elements Z_{11}, Z_{12}, Z_{22}, and Z_L in the equivalent circuit shown in Figure 2 may be solved for and equated to P_{in}. The power dissipated in the impedance elements in the equivalent circuit corresponds to the power terms on the right-hand side of Equation (1). However, there does not seem to be any easy way to relate the individual power terms in Equation (1) to the power dissipated in the impedance elements, with the exception of the power in the load impedance and the input power at the transmitting antenna. Although the equivalent circuit of a transmitting-receiving antenna system is important from a conceptual point of view, in practice it is not used very often. There are other more useful ways to calculate the received power, using concepts such as the effective complex vector length of an antenna, the effective receiving aperture or cross section, along with the gain function and impedance and polarization mismatch factors [4].

We now turn to a discussion of the Thévenin and Norton equivalent circuits for the receiving antenna, which are shown in Figure 3. In this figure, the Thévenin impedance, Z_T, the open-circuit voltage, V_{oc}, and the short-circuit current, I_{sc}, are given by

$$Z_T = Z_{11} - \frac{Z_{12}^2}{Z_{22} + Z_g}, \quad (3)$$

$$V_{oc} = \frac{Z_{12}}{Z_{22} + Z_g} V_g, \quad (4)$$

$$I_{sc} = \frac{V_{oc}}{Z_T}. \quad (5)$$

The power delivered to the receiving-antenna load impedance may be calculated using either the Thévenin or the Norton equivalent circuit, and is given by

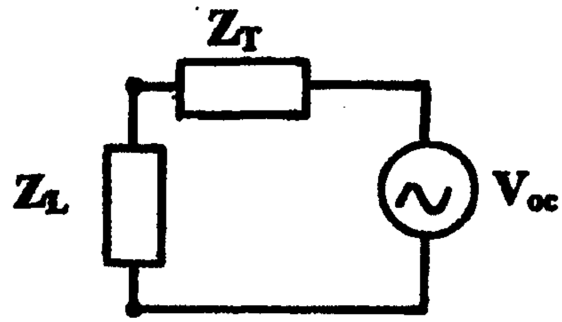

Figure 3a. The Thévenin equivalent circuit for a receiving antenna.

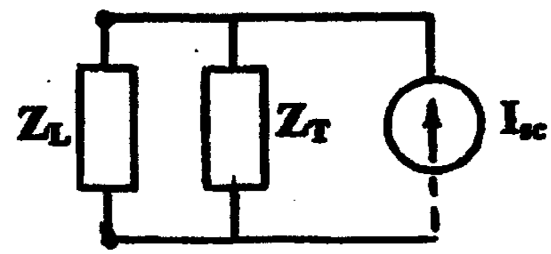

Figure 3b. The Norton equivalent circuit for a receiving antenna.

$$P_L = \frac{1}{2}\left|\frac{V_{oc}}{Z_L + Z_T}\right|^2 \mathrm{Re}\, Z_L. \quad (6)$$

If we use the Thévenin equivalent circuit, we find that the internal power dissipated in the network is given by

$$P_D = \frac{1}{2}\left|\frac{V_{oc}}{Z_L + Z_T}\right|^2 \mathrm{Re}\, Z_T, \quad (7)$$

while use of the Norton equivalent circuit gives

$$P_D = \frac{1}{2}\left|\frac{V_{oc}}{Z_L + Z_T}\right|^2 \left|\frac{Z_L}{Z_T}\right|^2 \mathrm{Re}\, Z_T. \quad (8)$$

These two results are different, except for the special case when the magnitude of the load impedance is equal to the magnitude of the Thévenin equivalent impedance. Thus, this result clearly shows that the Thévenin and Norton equivalent circuits can not be relied upon for calculating the internal power dissipation in the network shown in Figure 2 (which of the two expressions, Equations (7) or (8), should one use?).

The correct result for the internal power dissipation in the network shown in Figure 2 can be solved for. The result is different from that shown in Equations (7) or (8) and is quite complex, so it will not be given here. However, we can note that since the mutual impedance, Z_{12}, is usually very small, the power in the receiving-antenna load impedance will be very small compared to that delivered by the source generator to Z_g and Z_{22}. Thus, the total internal power dissipated in the network will be much greater than that given by either Equation (7) or (8), since those expressions are of the same order of magnitude as the power delivered to the receiving-antenna load. Silver [3] and Ramo and Whinnery [5] both state that the internal power dissipation as found for the Thévenin and Norton equivalent circuits should not be equated to the scattered power from the receiving antenna. For the special case when the load impedance is the complex conjugate of the Thévenin impedance, both Equations (7) and (8) give the result that the internal power dissipation is equal to the absorbed power in the load impedance. If the internal dissipated power was equated to the scattered power for this case the result would be correct for a dipole antenna, but not, in general, for other antennas. This special case does not prove that the internal dissipated power is equal to the scattered power. Another interesting special case is that of an open-circuited receiving antenna. If the receiving antenna is open circuited, then there is no internal power dissipation in the Thévenin equivalent circuit, which would suggest the incorrect result that an open-circuited antenna does not scatter any power. On the other hand, the Norton equivalent circuit has substantial internal dissipated power, even when the receiving antenna is open circuited. In general, as the two different results obtained in Equations (7) and (8) would suggest, the internal power dissipation in the Thévenin and Norton equivalent circuits does not have a physical meaning. There is no known power-conservation theorem associated with these equivalent networks.

Even though the Thévenin and Norton equivalent circuits are limited in the sense described above, they have some interesting properties that enable one to find the reradiated portion of the total field scattered by a receiving antenna. In the Appendix to this paper, we show that the total field scattered by a receiving antenna, in the presence of a transmitting antenna, is given by either of the two following expressions:

$$\mathbf{E}_s(Z_L) = \mathbf{E}_s(Z_L = 0) + \frac{I_{sc} Z_L}{Z_L + Z_A} \mathbf{E}_r, \tag{9}$$

$$\mathbf{E}_s(Z_L) = \mathbf{E}_s(Z_L = \infty) - \frac{V_{oc}}{Z_L + Z_A} \mathbf{E}_r, \tag{10}$$

where $E_s(Z_L)$ is the total scattered field when the receiving antenna is terminated in a load impedance Z_L; $\mathbf{E}_r$ is the field radiated by the receiving antenna for unit input current, in the presence of the transmitting antenna with its source generator short circuited, i.e., $V_g = 0$; $\mathbf{E}_s(Z_L = 0)$ is the field scattered by the receiving antenna when its terminals are short circuited, in the presence of the transmitting antenna; and $\mathbf{E}_s(Z_L = \infty)$ is the field scattered by the receiving antenna, but with its terminals open circuited. I_{sc} is the Norton equivalent current source, V_{oc} is the Thévenin equivalent open-circuit voltage source, and Z_A is equal to the Thévenin impedance Z_T, which is the input impedance of the receiving antenna when it is used for transmitting in the presence of the transmitting antenna with its voltage source short circuited.

An examination of Equation (9) shows that $Z_L I_{sc}/(Z_L + Z_A)$ is the portion of the current supplied by the Norton current source to the antenna input impedance Z_A. Thus, the interpretation that the current in Z_A gives rise to a reradiated field that is part of the total scattered field from the receiving antenna can be made. Similarly, in Equation (10), the term $-V_{oc}/(Z_L + Z_A)$ is the current flowing into the antenna impedance Z_A due to the Thévenin equivalent open-circuit voltage source. Hence, in the Thévenin equivalent circuit, the current produced in Z_A can be interpreted to produce the reradiated field that is a part of the total field scattered by the receiving antenna. Note that in addition to the reradiated field, the total scattered field includes the field scattered by the short-circuited antenna or the open-circuited antenna, respectively, for the two equivalent circuit models.

The result shown in Equation (9) is an old fundamental theorem in antenna scattering. It has been derived by Aharoni [6], King and Harrison [7], Stevenson [8], and by Collin using Dicke's scattering-matrix representation of a receiving antenna [9]. Aharoni used the compensation theorem for his derivation, and called the term $Z_L I_{sc} \mathbf{E}_r/(Z_L + Z_A)$ the reradiated field. Since the total scattered field is the superposition of two fields, it is clear that the scattered power cannot be found from the reradiated field component by itself. Thus, the calculation of the power dissipation in Z_A cannot be equated to the scattered power. A further point to note is that the far-zone radiation pattern of the field scattered by the open-circuited or short-circuited antenna is generally different from that of the reradiated field, which has the same pattern as that of the antenna when it is used as a transmitting antenna.

Aharoni notes that the total scattered power is primarily of theoretical interest, since due to interference with the radiated field from the transmitting antenna, the power radiated away to infinity has an interaction term as shown earlier in Equation (2), and thus the scattered power does not exist as a separate entity.

We will end this paper with some comments on the alternative equivalent circuit that Love presents in his paper [1]. Love's alternative equivalent circuit applies to a receiving antenna for which a physical aperture and the aperture efficiency can be defined. Love's equivalent circuit has both an equivalent voltage source and an equivalent current source. The circuit has the property that for an impedance-matched antenna, the received power is equal to the incident power flux multiplied by the product of the physical geometric area of the antenna and the aperture efficiency. The internal power dissipated in Love's equivalent circuit is equal to the incident power on the physical geometric area of the antenna minus the received power. This is interpreted as reradiated power. Even for an impedance-mismatched antenna, the circuit has the property that the sum of the absorbed power and the power dissipated internally in the network equals the incident power on the physical geometric area of the antenna. This author does not agree with Love that the internal power dissipated in his equivalent circuit can be equated to the reradiated or scattered power from the antenna. For example, if we had a large parabolic antenna with an aperture efficiency of 90 percent, the antenna would receive nine times as much power as it scatters, under impedance-matched conditions. This result is inconsistent with the physical requirement that the antenna must scatter much more power than that in order to create the deep shadow region behind the large parabolic-reflector antenna. Most antennas scatter more power than they absorb, unless they are minimum-scattering antennas, in which case the scattered power is equal to the absorbed power [10]. The analysis carried out in this paper has clarified the properties of the Thévenin and Norton equivalent circuits, and has showed that they can be used to find a reradiated electromagnetic field that is only a part of the total scattered field. At this time, it is not clear if a similar property would be true for Love's equivalent circuit.

Appendix

In this Appendix, we provide a derivation of the Norton and Thévenin equivalent circuits and the scattering formulas, Equations (9) and (10). In order to avoid unnecessary details, we assume that both the receiving and transmitting antennas are connected to their respective terminations by coaxial transmission lines, with inner radius a and outer radius b. The derivation can be easily generalized to include waveguides through use of equivalent currents and voltages for a circuit description. For coaxial transmission lines, the radial electric field at a terminal plane is given by

$$E_r = \frac{V}{r\ln(b/a)},$$

and the azimuthal magnetic field by

$$H_\phi = \frac{I}{2\pi r},$$

where V is the voltage across the line, and I is the current on the center conductor. We also have $E_r = Z_0 H_\phi$ and $V = Z_c I$, where $Z_0 = \sqrt{\mu_0/\varepsilon_0}$ is the intrinsic impedance of free space, and $Z_c = \frac{Z_0 \ln a/b}{2\pi}$ is the characteristic impedance of the coaxial transmission line. The positive direction of current flow is in the direction of the unit normal, **n**, which points into the volume outside of the antenna structures. If the coaxial transmission line is terminated in an impedance Z at a terminal plane, then the impedance boundary condition at the termination will be $V = -IZ$, since the positive direction for current flow is outwards, not into the termination. The corresponding impedance boundary condition on the transverse electric and magnetic fields is

$$\frac{E_r}{H_\phi} = \frac{V}{r\ln a/b}\frac{2\pi r}{I} = -\frac{ZZ_0}{Z_c}.$$

We will generally express the impedance boundary conditions at the terminal planes in terms of circuit quantities, but these can be translated into impedance boundary conditions for the transverse fields, as shown above. On the lossy metal surfaces of the antennas, we use the impedance boundary condition

$$\mathbf{E}_{\tan} = Z_m \mathbf{n} \times \mathbf{H},$$

where Z_m is the surface impedance of a lossy conductor, given by $Z_m = (1+j)/\sigma\delta_s$, with σ being the conductivity of the metal and δ_s being the skin depth.

In the derivation to follow, we will introduce three different field solutions. Solution 1 is the field $\mathbf{E}_a, \mathbf{H}_a$, produced when the transmitting antenna is operative and the receiving antenna has a short circuit at its terminal plane, T_1. Solution 2 is the field $\mathbf{E}_b, \mathbf{H}_b$, and is the total field when the receiving antenna is transmitting with unit input current $I_0 = 1$ in the presence of the original transmitting antenna with its voltage-source generator short circuited, i.e., $V_g = 0$. Solution 3 is the total field $\mathbf{E}_c, \mathbf{H}_c$, produced by the transmitting antenna in the presence of the receiving antenna when it is terminated in a load impedance Z_L. In all three cases, the field solutions are to be constructed so that the impedance boundary conditions on the metal surfaces of the antennas are satisfied, the radiation condition at infinity is satisfied, and the appropriate impedance boundary conditions on the terminal planes are satisfied. The imposed boundary conditions are sufficient to guarantee that the solutions will be unique. Hence, it will not be necessary to repeat the uniqueness arguments again. The first two solutions require the solution of a pair of coupled integral equations that will determine the unknown currents induced on the antenna surfaces. The third solution is obtained by using superposition. We do not give the detailed solutions, since it is sufficient for our purposes to only know the general composition of the field solutions.

For the first solution with the receiving-antenna terminals short circuited, the boundary condition on the terminal plane T_1 is

$$V_{1a} = 0 \text{ or } \mathbf{n} \times \mathbf{E}_a = 0, \tag{A1}$$

while on the terminal plane T_2 we have

$$V_{2a} = V_g - I_{2a} Z_g. \tag{A2}$$

We can construct the solutions for the fields $\mathbf{E}_a, \mathbf{H}_a$ by assuming that we know $H_\phi = I_{2a}/2\pi r$ on the terminal plane T_2. The solution will then determine a value for E_r, and thus a value for V_{2a}. The ratio V_{2a}/I_{2a} gives us the input impedance for the transmitting antenna in the presence of a receiving antenna with its input terminals short circuited. We fix I_{2a} in terms of the a priori specified source voltage V_g by using

$$V_{2a} = V_g - I_{2a} Z_g = Z_{in} I_{2a}, \tag{A3}$$

where Z_{in} is the transmitting-antenna input impedance. The solution will determine a value for H_ϕ on the terminal plane T_1 from which we obtain the current

$$I_{1a} = 2\pi r H_\phi = -I_{sc}, \tag{A4}$$

where we have defined the short-circuit current I_{sc} at the terminal plane to be into the shorted termination. For the second solution, when the receiving antenna is transmitting with unit input current $I_0 = 1$, the boundary condition on the tangential magnetic field on the terminal plane T_1 is

$$H_\phi = \frac{I_0}{2\pi r}. \tag{A5}$$

The solution will give a value for E_r and $V_{1b} = I_0 Z_A$ on the terminal plane T_1, where Z_A is the input impedance of the receiving antenna when it is transmitting in the presence of the original transmitting antenna with its voltage source short circuited. On the terminal plane T_2, the boundary condition is

$$V_{2b} = -I_{2b} Z_g. \tag{A6}$$

For the third solution, for the case when the receiving antenna is terminated in a load impedance Z_L, we will show that the total field is the superposition of the fields $\mathbf{E}_a, \mathbf{H}_a$ and $\mathbf{E}_b, \mathbf{H}_b$ in the form

$$\mathbf{E}_c = \mathbf{E}_a + \frac{I}{I_0}\mathbf{E}_b, \qquad \mathbf{H}_c = \mathbf{H}_a + \frac{I}{I_0}\mathbf{H}_b,$$

with I to be chosen so that the impedance boundary condition on the terminal plane T_1 is satisfied. Clearly, the field $\mathbf{E}_c, \mathbf{H}_c$ satisfies the impedance boundary conditions on the metal surfaces of the antennas and the radiation condition at infinity, since both the a and b fields satisfy these boundary conditions. On the terminal plane T_2, the required boundary condition is

$$V_{2c} = V_{2a} + \frac{I}{I_0}V_{2b} = V_g - \left(I_{2a} + \frac{I}{I_0}I_{2b}\right)Z_g = V_g - I_{2c}Z_g. \quad \text{(A7)}$$

When we substitute for V_{2a} and V_{2b} from Equations (A2) and (A6), we find that this boundary condition is satisfied independently of the value of I. In order to satisfy the impedance boundary condition on the terminal plane T_1, we require

$$V_{1c} = -I_{1c}Z_L = V_{1a} + \frac{I}{I_0}V_{1b}$$
$$= 0 + \frac{I}{I_0}I_0 Z_A = -\left(I_{1a} + \frac{I}{I_0}I_0\right)Z_L \quad \text{(A8)}$$

This equation may be solved for I to give

$$I = -\frac{I_{1a}Z_L}{Z_L + Z_A} = \frac{I_{sc}Z_L}{Z_L + Z_A}. \quad \text{(A9)}$$

The current into the load Z_L is the total current $-I_c$, which is given by

$$I_L = -I_c = I\frac{Z_A}{Z_L} = \frac{I_{sc}Z_A}{Z_L + Z_A}. \quad \text{(A10)}$$

We note that this is the current that would be produced by the current source in the Norton equivalent circuit. Also, Z_A was calculated as the input impedance seen at the terminals of the receiving antenna with the voltage source at the transmitting antenna short circuited, and thus is equal to the Thévenin impedance, Z_T, used in the equivalent circuit. Thus, we have obtained results that can clearly be interpreted in terms of the Norton equivalent circuit.

We will now proceed to determine the field scattered by the receiving antenna. The induced currents on the receiving antenna, due to the field $\mathbf{E}_a, \mathbf{H}_a$ can be found and used to calculate the scattered electric field from the receiving antenna under short-circuit conditions. We will designate this field as $\mathbf{E}_s(Z_L = 0)$. Likewise, we can find the radiated electric field from the induced currents on the receiving antenna and the input current when it is used to transmit with unit input current. We will designate this field as $\mathbf{E}_r$. Then, by superposition, we find that the total scattered field is given by

$$\mathbf{E}_s(Z_L) = \mathbf{E}_s(Z_L = 0) + I\mathbf{E}_r = \mathbf{E}_s(Z_L = 0) + \frac{I_{sc}Z_L}{Z_L + Z_A}\mathbf{E}_r, \quad \text{(A11)}$$

which is the scattering formula given by Equation (9). The derivation of the Thévenin equivalent circuit and the scattering formula, Equation (10), proceeds in a similar fashion. The only change is that the field $\mathbf{E}_a, \mathbf{H}_a$ is calculated for an open-circuited termination of the receiving antenna. The boundary condition on the terminal plane T_1 is $\mathbf{n} \times \mathbf{H}_a = 0$ in place of $\mathbf{n} \times \mathbf{E}_a = 0$. The solution determines a value for the electric field on the terminal plane T_1, and a corresponding terminal-plane voltage, which we call V_{oc}. The fields $\mathbf{E}_b, \mathbf{H}_b$ and $\mathbf{E}_c, \mathbf{H}_c$ are calculated as before, and I is chosen so that the boundary condition

$$V_{1c} = -I_{1c}Z_L = V_{oc} + \frac{I}{I_0}V_{1b} = V_{oc} + \frac{I}{I_0}Z_A = -\frac{I}{I_0}I_0 Z_L \quad \text{(A12)}$$

is satisfied on the terminal plane T_1. We can solve for the current I, which is found to be

$$I = -\frac{V_{oc}}{Z_A + Z_L} = -I_L, \quad \text{(A13)}$$

since the total current at the terminal plane T_1 is simply I. This result shows that the load current is equal to $V_{oc}/(Z_L + Z_A)$, which can be interpreted in terms of a Thévenin equivalent circuit. As before, we separate out the electric field scattered and radiated by the receiving antenna in terms of induced currents on the receiving antenna, and thereby obtain

$$\mathbf{E}_s = \mathbf{E}_s(Z_L = \infty) - \frac{V_{oc}}{Z_L + Z_A}\mathbf{E}_r, \quad \text{(A14)}$$

which is the scattering formula given by Equation (10).

References

1. A. W. Love, "Comment: On the Equivalent Circuit of a Receiving Antenna," *IEEE Antennas and Propagation Magazine*, **44**, October 2002, pp. 124-125.

2. J. D. Kraus, *Antennas, Second Edition*, New York, McGraw-Hill Book Company, 1988, Chapter 2.

3. S. Silver, *Microwave Antenna Theory and Design*, (MIT Radiation Laboratory Series, Volume 12), New York, McGraw-Hill Book Company, 1949, Chapter 2.

4. R. E. Collin, *Antennas and Radio Wave Propagation*, New York, McGraw-Hill Book Company, 1985, Chapter 5.

5. S. Ramo and J. Whinnery, *Fields and Waves in Modern Radio*, New York, John Wiley and Sons, Inc., 1962, Chapter 12.

6. J. Aharoni, *Antennas*, Oxford, Clarendon Press, 1946, pp. 164-176.

7. R. W. P. King and C. W. Harrison, Jr., "The Receiving Antenna," *Proc. IRE*, **32**, January 1944, pp. 18-34; see also pp. 35-49 for a related paper by the same authors.

8. A. F. Stevenson, "Relations Between the Transmitting and Receiving Properties of Antennas," *Quart. Appl. Math.*, **5**, January 1948, pp. 140-148.

9. R. E. Collin and F. J. Zucker, *Antenna Theory, Volume 1*, New York, McGraw-Hill Book Company, 1969, Chapter 4.

10. C. G. Montgomery, R. H. Dicke, and E. M. Purcell, *Principles of Microwave Circuits*, (MIT Radiation Laboratory Series, Volume 8), New York, McGraw-Hill Book Company, 1948, Chapter 9.

In Memoriam: Richard C. Johnson

Richard C. Johnson, a Georgia Tech Research Institute (GTRI) retiree, world-renowned antenna expert, and exceptional mentor to young engineers, died in January, 2003, after a long battle with Parkinson's disease. He was 72.

Johnson invented and patented the compact antenna range, which allows installations of microwave antennas to be measured and tested accurately indoors. Compact ranges are used worldwide today. Johnson also designed and improved antennas for ship surface-search radars, hundreds of which were installed aboard US Navy vessels, said Jim Cofer, director of GTRI's Business Development Office.

"Most designers of that era concentrated on the main beam region of an antenna," Cofer said. "Dick recognized that most interference/susceptibility of a system occurred in the other 99.9 percent of the antenna's spherical domain. Therefore, he included these considerations in his designs.

"Johnson's widespread recognition was directly responsible for establishing the threat simulator research and development base at GTRI," said Cofer. "Cumulative funding for this area is now well over $200 million, making it possibly the largest continuously funded research area at GTRI."

Johnson was ahead of his time: many of his simple yet elegant antenna designs were based on theoretical applications he'd postulated 20 years earlier, said GTRI senior research engineer Rickey Cotton. He also believed in passing on his knowledge to younger engineers. To that end, Johnson organized Friday afternoon "Antenna Bull Sessions" in his office in the late 1960s for a select group of younger GTRI researchers, including Cofer, Neal Alexander, and Don Bodnar, assigning homework projects and teaching them to solve difficult, real-world antenna problems.

"In addition, when riding the daily shuttle from our facility in Cobb County to campus in the mornings and evenings, Dick used to quiz and tutor the co-ops also riding the shuttle that were working with us on the programs," Cotton recalled. "He would quiz them on the frequency limits of the radar bands, and typical waveguide sizes associated with each."

A Georgia Tech alumnus, Johnson taught electrical engineering at Georgia Tech and wrote several books, including two editions of the *Antenna Engineering Handbook*, continuing to share with others what he'd learned. Several years after retiring, Johnson even gave Cotton his library of microwave books – which Cotton continues to use in his work.

For his many contributions to antenna research, design and applications, Johnson was recognized by numerous organizations, including the IEEE, which elected him to the grade of Fellow in 1975. In 1988, Johnson also became the first Georgia Tech research faculty member to receive Board of Regents-approved emeritus status. Johnson was a member of the AP-S AdCom (1978-84), an AP-S Distinguished Lecturer (1978-79), President of AP-S (1980), and Editor of the AP-S *Newsletter*. He received the Distinguished Achievement Award from AMTA in 1989.

Memorial contributions may be made Johnson's name to Shepherd Center, 2020 Peachtree Road, NW, Atlanta, GA 30309 USA.

[The above article was provided by Lea McLees of GTRI (e-mail: Lea.McLees@gtri.gatech.edu). Information about AP-S activities was added.]