

lation level was obtained, maintaining the level at the desired signal constant. The resulting radiation pattern in Fig. 3(b) shows nulls in the interference directions, without sidelobe pattern degradation in the other directions. This radiation pattern features the desirable performance common to the Applebaum-Howells array.

#### V. CONCLUSION

A systolic array architecture for the Applebaum-Howells array has been derived. The proposed procedure consisted of two steps: the input element signal orthogonalization and the feedback loop elimination procedures. It has been shown that the orthogonality in output signals from the Gram-Schmidt processor, which was employed as the preprocessor, can eliminate the global signal feedback loop needed for the conventional Applebaum-Howells processor and that the Applebaum-Howells array can be effectively implemented by using the systolic array architecture. The simulation results also show that the proposed processor features the desirable characteristics of the radiation pattern with low sidelobe level common to the Applebaum-Howells array. Regarding the importance of the Applebaum-Howells array in the adaptive array field, it is concluded that the result obtained here has derived a significant solution to overcome the systolic array implementation problem for the adaptive array.

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### Theoretical Increase in Radiation Efficiency of a Small Dipole Antenna Made with a High Temperature Superconductor

ROBERT J. DINGER AND DAVID J. WHITE, MEMBER, IEEE

**Abstract**—An electrically small wire dipole antenna with a shunted stub matching network has been analyzed to determine the improvement in radiation efficiency that can be achieved by making the metallic components from high temperature superconducting (HTS) material. The analysis shows that significantly higher radiation efficiencies can be obtained with HTS material only if the dielectric losses are kept at very low values. A loss tangent of less than  $10^{-4}$  is necessary for any dielectric materials in the transmission lines and support structures for the

antenna; a loss tangent of at least  $10^{-5}$  is desirable, particularly if the dipole length is 0.1 wavelength or less.

#### INTRODUCTION

The discovery of ceramic materials such as Y-Ba-Cu-O that are superconducting at temperatures high enough to permit efficient cooling of relatively large structures leads one to consider their application to antennas. The improved antenna performance arises from a potentially much lower value of surface resistance  $R_s$ , which leads to lower ohmic losses and improved radiation efficiency. To date the measured values of  $R_s$  at microwave frequencies are about one order of magnitude lower than copper [1], and there is the promise that improved material processing will eventually lead to  $R_s$  values many orders of magnitude lower than copper.

The primary candidate antenna for a high temperature superconductivity (HTS) application is one in which low efficiency is a result of ohmic losses comparable to or larger than the radiation resistance. This situation is found with electrically small antennas, for which the radiation resistance decreases much faster than the conductor losses as the antenna is made smaller. To assess the potential improvement in efficiency that is possible by replacing the normal conductors with HTS, we have analyzed in detail the electrically small antenna shown in Fig. 1. A short dipole of overall length  $L$  is driven from a balanced twin-lead transmission line. A shunt short-circuit stub is used to match the impedance of the transmission line to the antenna. The twin-lead transmission line is enclosed in a dielectric, which is required in a practical application both for support and efficient cooling of the conductors. The transmission line size parameters were selected as sizes that are practical for antennas in the 100 MHz to 1 GHz frequency range and that would give a 50- $\Omega$  impedance with a reasonable value of dielectric constant.

The antenna and matching network in Fig. 1 is probably a feasible configuration for a practical HTS antenna; indeed, it is similar to the HTS antenna for which measurements were presented in [2]. However, this configuration also serves as a surrogate for a wide variety of electrically short antennas, such as a wire monopole driven by coax line and a printed circuit dipole driven by a microstrip transmission line.

#### ANALYSIS

The performance parameter of interest is the radiation efficiency:

$$\eta = \text{radiated power} / \text{input power} = P_r / P_{in} = P_r / (P_r + P_{loss}) \quad (1)$$

where  $P_{loss}$  is the power dissipated in conductor ohmic losses and in dielectric losses. The input power is calculated at the input to the matching network, rather than the antenna terminals; this is always the best way to calculate antenna efficiency, but is particularly appropriate for electrically short antennas because of the large (and potentially lossy) reactive elements in the matching network needed to match the antenna reactance. The antenna input impedance is given by

$$Z_a = R_r + R_c + jX_a \quad (2)$$

where  $R_r$  is the radiation resistance,  $R_c$  is the conductor loss in the antenna, and  $X_a$  is the antenna reactance. A suitable approximation to  $R_r$  for this dipole is given by [3]

$$\begin{aligned} R_r &= 20\pi^2 (L/\lambda)^2, & 0 < L < 0.188\lambda \\ R_r &= 24.7(\pi L/\lambda)^{2.4}, & 0.188\lambda < L < \lambda/2. \end{aligned} \quad (3)$$

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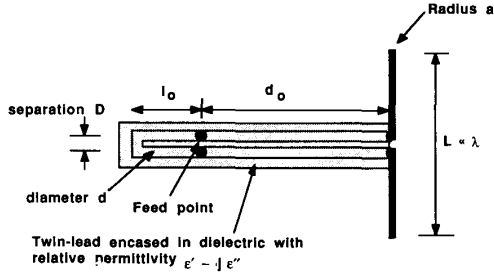


Fig. 1. Short dipole antenna with twin-lead stub matching network. The twin-lead transmission line is encased in a dielectric jacket. For the calculated curves the following values were used:  $D = 0.012\lambda$ ,  $d = 0.01\lambda$ ,  $a = 0.0005\lambda$ ,  $\epsilon' = 2.23\epsilon_0$ .

For  $L$  up to about  $0.4\lambda$  (the largest  $L$  of interest in this communication), we can assume a triangular current distribution, for which  $R_c$  can be approximated by [3]

$$R_c = R_s L / 6\pi a \quad (4)$$

where  $R_s$  is the surface resistance of the antenna conductor and  $a$  is the antenna wire radius. To approximate the input reactance of the dipole, we use the best-fit straight line to the curve presented in [3] for  $a = 0.0005\lambda$ , with the result that

$$X_a = -1450 + 3140(L/\lambda) \Omega. \quad (5)$$

The analytical expressions given in [4] could be used for  $X_a$ , or the method of moments could be used to calculate  $X_a$ , but the resulting curves shown below would have essentially the same shape and trend, with similar conclusions.

The conventional impedance matching analysis leads to the following equations for the load distance  $d_0$  and the stub length  $l_0$  [5]:

$$d_0 = (1/\beta) \tan^{-1} \frac{-X_a Z_c + \{R_a Z_c [(Z_c - R_a)^2 + X_a^2]\}^{1/2}}{Z_c (Z_c - R_a)} \quad (6)$$

and

$$l_0 = (1/\beta) \tan^{-1} [S^{1/2}/(S - 1)]. \quad (7)$$

In (6) and (7),  $\beta = 2\pi\sqrt{\epsilon'}/\lambda$ ,  $Z_c$  is the characteristic impedance of twin-lead line, given by

$$Z_c = \pi^{-1}(\mu_0/\epsilon')^{1/2} \cosh^{-1}(D/d), \quad (8)$$

$S = (1+\rho)/(1-\rho)$ , and  $\rho$  is the magnitude of the reflection coefficient  $\Gamma_a$  of the antenna load:

$$\Gamma_a = (Z_a - Z_c)/(Z_a + Z_c). \quad (9)$$

The propagation constant on the twin-lead line is given by  $\gamma = \alpha_d + \alpha_c + j\beta$ ; the dielectric attenuation is given by  $\alpha_d = \pi(\epsilon''/\epsilon')/\lambda$  and the conductor loss attenuation is given by  $\alpha_c = R/2Z_c$ , where

$$R = \frac{2R_s D}{\pi d^2 [(D/d)^2 - 1]^{1/2}}. \quad (10)$$

The dielectric loss tangent is given by the ratio  $\epsilon''/\epsilon'$ .

The derivation of the equation for  $\eta$  is a straightforward application of impedance matching procedures discussed, for example, in [5]; hence, we only give the final equations. We assume unit incident power on the transmission line connected to the antenna and matching section. The power dissipated in the entire matching structure and antenna (due to radiation and dissipative losses) is given by

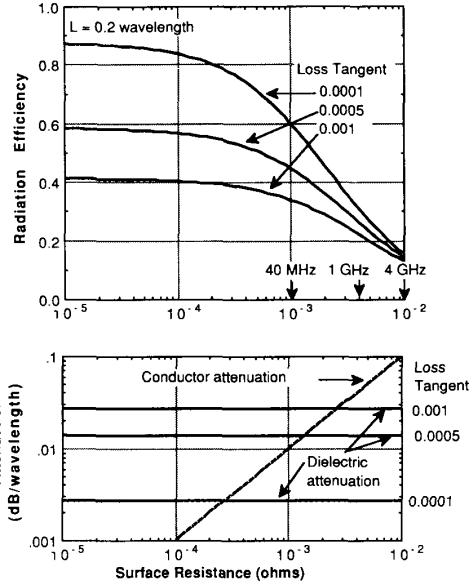


Fig. 2. Radiation efficiency and attenuation versus surface resistance for  $L = 0.2\lambda$ . The arrows on the abscissa of the radiation resistance plot indicate the surface resistance for 77°C copper at the frequency shown.

$$P_{in} = (1/2) \operatorname{Re} \left[ \frac{|T_0|^2 (1 + \Gamma_a e^{-2\gamma d_0}) (1 - \Gamma_a^* e^{-2\alpha d_0 + j2\beta d_0})}{Z_c |1 - R_0 \Gamma_a e^{-2\gamma d_0}|^2} + Y_s' \right] \quad (11)$$

where

$$T_0 = 2Y_c / (Y_s' + 2Y_c)$$

$$R_0 = -Y_s' / (Y_s' + 2Y_c)$$

$$Y_s' = Y_c \coth(\gamma l_0)$$

and

$$\alpha = \alpha_c + \alpha_d.$$

The radiated power is given by

$$P_r = \frac{R_r |T_0|^2 |1 + \Gamma_a|^2 e^{-2\alpha d_0}}{2|Z_a|^2 |1 - R_0 \Gamma_a e^{-2\gamma d_0}|^2}. \quad (12)$$

The efficiency is then calculated from (1).

#### DISCUSSION

Figs. 2 and 3 present some plots of the efficiency as a function of the various antenna and matching network parameters. In Fig. 2 we plot the efficiency as a function of the surface resistance for a dipole length of  $0.2\lambda$ , with the dielectric loss tangent as a parameter. Since all of the length parameters can be scaled to wavelength in the matching equations, the curves shown are independent of absolute frequency; for reference, the frequency at which copper at a temperature of 77°K has the  $R_s$  value on the abscissa is indicated at several points. The three-decade range of  $R_s$  shown encompasses the surface resistance values that can reasonably be expected from improved HTS material. As  $R_s$  is lowered, the efficiency improves as expected, but it levels off at a value of  $R_s$  where dielectric losses begin to dominate. This value occurs when  $\alpha_d$  begins to exceed  $\alpha_c$ , as shown in Fig. 2. Unless the loss tangent is  $10^{-4}$  or less, only a rather modest improvement in radiation efficiency can be realized.

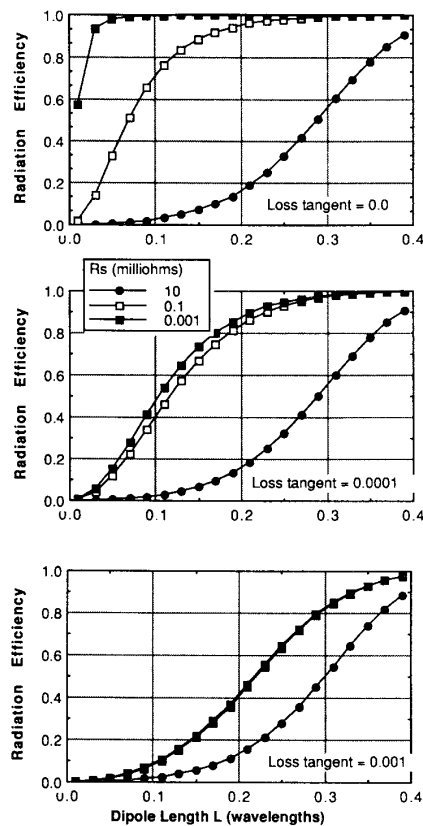
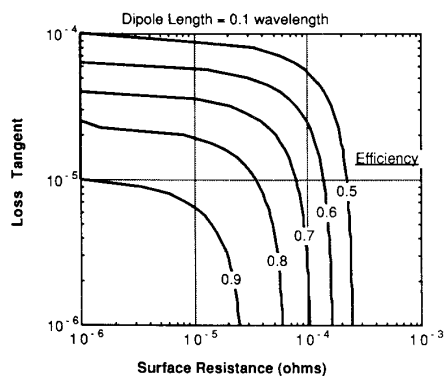


Fig. 3. Radiation efficiency versus dipole length for three loss tangents.

Fig. 4. Locus of surface resistance and loss tangent necessary to yield a given efficiency for a dipole length of  $0.1\lambda$ .

The importance of a very low loss tangent is also made evident in Fig. 3, in which we plot the radiation efficiency as a function of dipole length. For no dielectric losses, radiation efficiencies approaching 100% can be obtained with dipoles as short as  $0.1\lambda$ . However, even a loss tangent as low as  $10^{-4}$  cuts into the enhanced efficiency substantially, limiting the high efficiency region to  $0.2\lambda$  or longer. The same point is made in Fig. 4, in which we have fixed the dipole length at  $0.1\lambda$ , and plotted the locus of  $R_s$  and loss tangent values that produce a given radiation efficiency. Clearly, the antenna and all supporting and material structures must use very low loss dielectrics and ceramic materials to realize the enhanced efficiency that HTS promises.

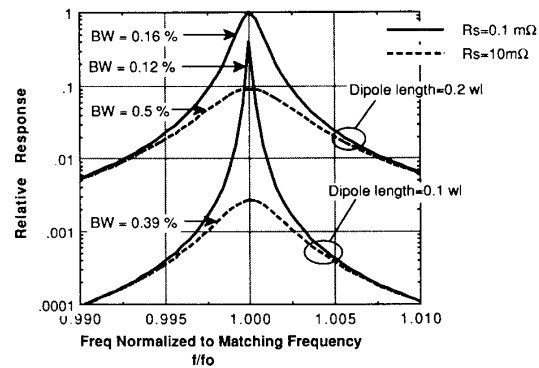
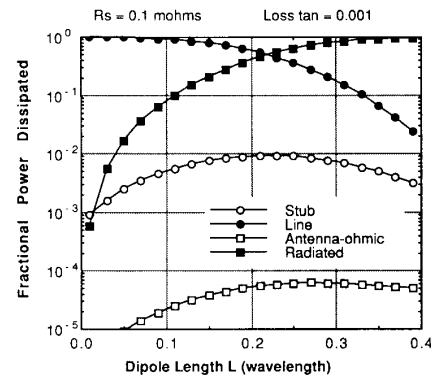
Fig. 5. Relative frequency response of matching circuit and antenna. The 3 dB bandwidth are indicated for each curve. Loss tangent =  $10^{-4}$  for these curves.

Fig. 6. The fraction of power dissipated in the indicated portions of the antenna and matching network. The portion labeled "line" refers to the section of transmission line between the stub and the antenna.

The frequency response of the matching circuit and antenna is shown in Fig. 5 in which the relative magnitude of the radiated power is plotted. The bandwidths are very narrow, as is typical for electrically short antennas. The relative improvement offered by a lower  $R_s$  is inversely proportional to dipole length; for  $L = 0.1\lambda$ , a 20 dB improvement in radiated power can be achieved, compared to  $10\lambda$  for  $L = 0.2\lambda$ .

Low loss materials in the transmission line between the matching stub and the antenna are especially critical. In Fig. 6 we plot the fractional power losses in the various portions of the antenna and matching network for a representative pair of values of  $R_s$  and loss tangent. The large standing waves between the stub and antenna terminals produce most of the loss; the antenna ohmic losses actually account for the smallest fraction of the losses. These results indicate that most of the 6 dB improvement in radiated power measured for the HTS Y-Ba-Cu-O antenna in [2] came from the HTS material in the matching network, rather than the antenna proper. Fig. 6 also suggests that only the matching network needs to be made from HTS material.

#### CONCLUSION

We have analyzed a typical electrically small wire dipole antenna with a shunted stub matching network to determine the improvement in radiation efficiency that can be achieved by making the metallic components from HTS material. The analysis shows that significantly higher radiation efficiencies can be obtained with HTS material only if the dielectric losses are kept at very low values. A loss tangent of

less than  $10^{-4}$  is necessary for any dielectric materials in the transmission lines and support structures for the antenna; a loss tangent closer to  $10^{-5}$  is desirable, particularly for dipole lengths of  $0.1\lambda$  (or less). Most of the losses (dielectric and conductive) occur in the section of line between the stub and the antenna terminals, so that low loss dielectric material is especially important here. In fact, a significant improvement in efficiency can be obtained by making only the matching network from HTS material, since only a small fraction of the losses occur in the antenna proper.

In principle, the best solution is to avoid dielectric materials as much as possible by using, for example, hollow waveguides. For frequencies much below 10 GHz, however, hollow waveguides are impractical. Another factor is the low thermal conductivity [6] of the HTS materials; efficient cooling of an antenna and matching structure will require heat transfer media of some type that will introduce dielectric loss. These two conflicting requirements will complicate the design of practical HTS electrically small antennas.

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### Comments on "A Simplification of the Synthesis of Parallel Wire Antenna Arrays"

HARVEY K. SCHUMAN, SENIOR MEMBER, IEEE

Apparently the authors and reviewers of this paper<sup>1</sup> were unaware of earlier work by Hirasawa and Sinnott [1] and Sanzgiri and Cummins [2], both published in the IEEE TRANSACTIONS ON ANTENNAS AND PROPAGATION. The useful relationship between point source array theory and a method of moments formulation of a parallel wire array antenna was clearly described in the earlier work. At the very minimum, the authors should have referenced this work and identified in what way their own work contributed something new.

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<sup>1</sup>K. Siakavara and J. N. Sahalos, *IEEE Trans. Antennas Propagat.*, vol. 37, pp. 936-940, July 1989.

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### Author's Reply by J. N. Sahalos, Senior Member, IEEE

I thank Dr. Schuman for his interest, comments, and for directing our attention to some interesting papers. With no objection his comment is a very positive statement and with no qualms we could reference [1] and [2]. Some points of comparison of our contribution<sup>1</sup> with [1] and [2] are the following: 1) we give two different formulations for the purpose of simpler use (one for linear and one for planar arrays) and 2) we relate the point source excitations with the corresponding voltages (see (17)), instead of the currents, of the wire antennas.

Reference [1] is directly related to our communication<sup>1</sup> but in contrast to [1], which makes use of the far-zone vector potential, we use the far-zone electric field.

Our misrepresentation had nothing to do with the characterizations of Dr. Schuman. In the same area of our work<sup>1</sup> one can find several equally interesting articles summarized by the title, "Antenna design by the method of moments," [3]-[6]. Our main purpose was to extend the Kang and Pozar solution [7] for nonuniform wire antennas arrays and to find a simpler method than the orthogonal one [8]. At the same time, emphasis has been given to the engineering applications where we found an improvement in the design by using parasitic wire antennas. This explains the title and the reason for the appearance of the contribution in this TRANSACTIONS.

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