

# An Efficient, Low-profile Antenna Employing Lossy Magneto-Dielectric Materials

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## Abstract

It has been shown that a magnetic core in a loop antenna excludes energy from the interior reducing radiation  $Q$  and that a core with magnetic losses still provides this reduction of stored energy without excessive dissipation provided that the magnitude of the complex permeability is sufficiently large. We present a broadband, low-loss, low-profile antenna based on a lossy magneto-dielectric material, a Ni-Zn ferrite operating in its dispersion region. The antenna behaves as a magnetic dipole and a radiation efficiency of almost 65 % is obtained at a frequency for which the loss tangent of the material is greater than 10.

## 1 Introduction

It is well known that the introduction of a magnetic core into an electrically-small generalized loop or TE<sub>1</sub> mode spherical wire antenna tends to exclude magnetic energy from the interior of the antenna thus reducing the radiation  $Q$  [1,2]. More recently, it has been shown for the specific case of the TE<sub>1</sub> mode spherical wire antenna with a homogeneous core, that the internal magnetic energy in the core cannot be forced to zero, but rather a minimum value of stored internal magnetic energy is effected using a finite permeability magnetic core [3-7]. The analysis in [3-6] also indicates that the dielectric polarizability of the core is detrimental to the minimization of internal stored energy. Specifically, the quality of a magnetic material for loop antenna applications is determined by the ratio of the relative permeability to the relative permittivity,  $\frac{\mu_R}{\epsilon_R}$ . Finally, very recently, it has been shown that the optimum lossless magnetic core for the TE<sub>1</sub> mode spherical wire antenna is actually composed of a thin magnetic shell or mantle covering a conducting kernel [5-7]. While the spherical geometry considered in refs. [3-7] is clearly the most appropriate for exciting only the lowest order TE<sub>1</sub> spherical mode and thus exhibiting the lowest external radiation  $Q$ , it is not a particularly practicable geometry for an antenna, especially if conventional magnetic materials, most of which exhibit a specific gravity of approximately 5, are to be employed. Of course, a magnetic dipole located an electrically-short distance away from and with its axis parallel to a ground plane exhibits a constructive image in the ground plane. Therefore, the kernel/mantle geometry in refs. [5-7] is considerably more practicable if the conducting kernel is realized as a hollow metallic hemispherical shell or even a more generalized convex deformation in a conducting ground plane. Recent efforts in the development of broadband, electrically-small magnetic dipoles have elicited renewed interest in the development of magnetic materials with low magnetic loss in the UHF and microwave range. Unfortunately, conventional magnetic materials are intrinsically lossy and all practical ferrite materials exhibit pronounced magnetic loss near their ferromagnetic resonance frequencies. However, as has been shown, this does not destroy of the ability of the material to facilitate the transfer of electromagnetic power and it has also been shown that a lossy magnetic core, that is a core with magnetic losses, still provides this beneficial reduction of stored energy in a TE<sub>1</sub> mode spherical wire antenna without incurring excessive dissipation provided that the magnitude of the complex permeability is sufficiently large [8].

A useful analogy for understanding the action of a lossy core is the lossy flux-coupled transformer. While the vast majority of power and RF signal transformers employ extremely low-loss cores, some RF flux-coupled transformers operate almost entirely in the dispersion regime of the constituent ferrite material of their cores. The operation can be qualitatively explained by considering the idealized operation of two magnetically-coupled coils, without leakage inductance or parasitic inter-winding capacitance. That is, consider two coupled inductances,  $L_1$  and  $L_2$ , with mutual inductance  $M$ . The insertion loss or forward

transfer scattering parameter in a system with characteristic impedance  $Z_0$  is:

$$S_{21} = \frac{2j\omega M}{Z_0 + j\omega(L_1 + L_2)}. \quad (1)$$

The values of circuit elements,  $L_1$ ,  $L_2$ , and  $M$ , must be determined from the electromagnetics of the coupled coils, but it is worthwhile to consider the action of an idealized toroidal coil. The self inductance of such a coil is  $L(\omega) = \mu_0\mu_R(\omega)nA/L_m$  where  $n$  is the number of turns,  $A$  is the cross sectional area, and  $L_m$  is the magnetic path length. Precisely the same arguments used to derive the self inductance also hold for the mutual inductance of two such coils wound on the same core and a complex core permeability causes both the self and mutual inductances to become complex. Note that when the magnitude of  $\omega L_1$  and  $\omega L_2$  are large compared to  $Z_0$ , the insertion loss becomes:

$$S_{21} \approx \frac{2M}{(L_1 + L_2)} \quad (2)$$

which is independent of the phase angle of the self and mutual inductances. Apparently, the transformer would function even in the hypothetical case of a purely imaginary permeability, provided the magnitude of the permeability were sufficiently large. One might think of the loop antenna as a transformer in which the “secondary” consists of the lines of electric displacement current surrounding the antenna that are associated with the near field. Thus, the “secondary” occupies a volume around the antenna with the density of the secondary windings diminishing with distance from the antenna. The problem with borrowing concepts from flux-coupled transformer design is that the “secondary” of the magnetic loop antenna (and the load on it) cannot be easily adjusted by the designer of the antenna. Clearly, it depends on the geometry of the antenna and thus for a spherical wire antenna is fixed for a given electrical size. Therefore, in order to improve upon the results obtained in [8] a different approach was deemed necessary.

## 2 Practical Ferrite-based Magnetic Dipole Antenna

The magnetic dipole antenna considered here is not spherical but instead a flat array of magnetic cores of rectangular cross section on a conducting ground plane each being driven by co-phased shielded loops at two locations symmetrically located about the centerline as shown in Fig. 1.

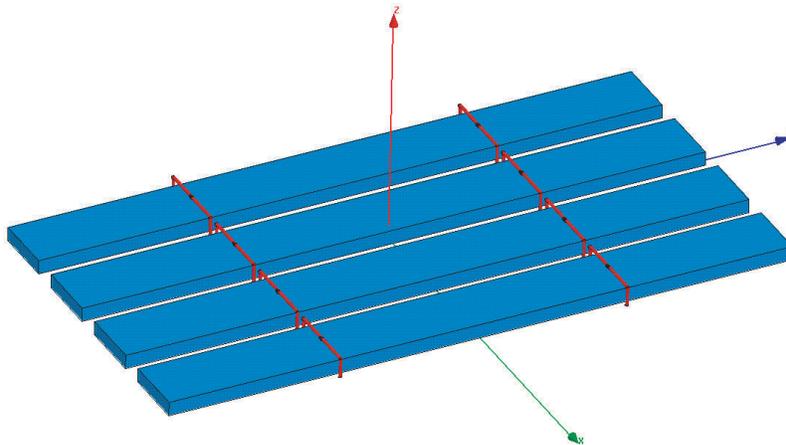


Figure 1: Concept low profile antenna consisting of 4 magnetic “loop-stick” elements. Each element has two shielded loop feeds located symmetrically with respect to its center. The dimensions of each ferrite bar are 200mm (wide) x 2000mm (long) x 45mm (thick)

If only one such core were considered and the core were driven in its center, the antenna geometry would degenerate to that of a (rather large) loop-stick antenna placed flat against a conducting ground plane.

The ferrite material employed here is a Ni-Zn ferrite with a low-frequency initial relative permeability of about 3200. The peak in  $\mu''_R$  occurs at about 2 MHz. The relative permittivity is essentially constant and real ( $\epsilon_R \approx 14.0$ ) over the frequency range considered. In the simulation, complex, frequency-dependent permeability, permittivity, and conductivity were represented using a look-up table which, in turn, was based on a large quantity of measured data. The antenna was characterized using a commercial finite element simulation, HFSS. In Fig. 2, the numerically-predicted radiation efficiency of the antenna is plotted along with the magnetic loss tangent of the material. As can be seen, the radiation efficiency varies from about 34-64% over the range of 30-100 MHz even though the magnetic loss tangent ranges from 5-14 over the same frequency range. In Fig. 3, the principal plane patterns are plotted. From the pattern data, it is clear that the antenna exhibits essentially the pattern of a magnetic dipole. The maximum directivity of the antenna is 1.8 which indicates that there is a slight admixture of higher order spherical modes in the pattern. The antenna is not truly electrically small, but it is nevertheless very low profile.

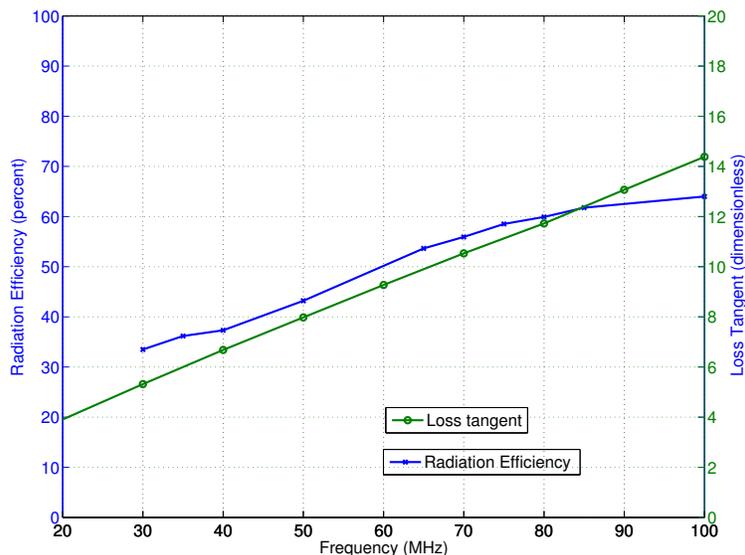


Figure 2: Radiation efficiency of the antenna (blue) and the magnetic loss tangent,  $\frac{\mu''_R}{\mu'_R}$ , of the material (green). Thus, at 100 MHz the imaginary component of the relative permeability is more than 14 times as large as the real part. The ferromagnetic resonance of this material occurs at approximately 2 MHz and the entire frequency range considered here lies in the dispersion region of the ferrite.

### 3 Conclusion

A very low profile, essentially flat, magnetically-loaded antenna conformal to a conducting surface has been proposed, described, and characterized using a numerical simulation based on experimental material data. A differentiating feature of this antenna is that the constituent magnetic material is predominantly lossy. The loss tangent of the magnetic material ranges from about 2.5 at 10 MHz to greater than 14 at 100 MHz. In this frequency range, the same ferrite material can be used effectively as a planar absorber with a normal incidence return loss of about 20 dB at 30 MHz increasing to over 30 dB at 100 MHz. However, when driven as an antenna, the radiation efficiency ranges from 35-64 percent over this same frequency range. The pattern of the antenna in the frequency range is essentially that of a magnetic dipole with the dipole axis being the axis of the loop-stick elements or the  $y$  axis in Figure 1. As the aperture is made larger while the feed spacing is maintained, one obtains essentially a sheet of magnetic displacement current over a ground plane uniform in the  $x$  direction and with a sinusoidal variation in the  $y$  direction.

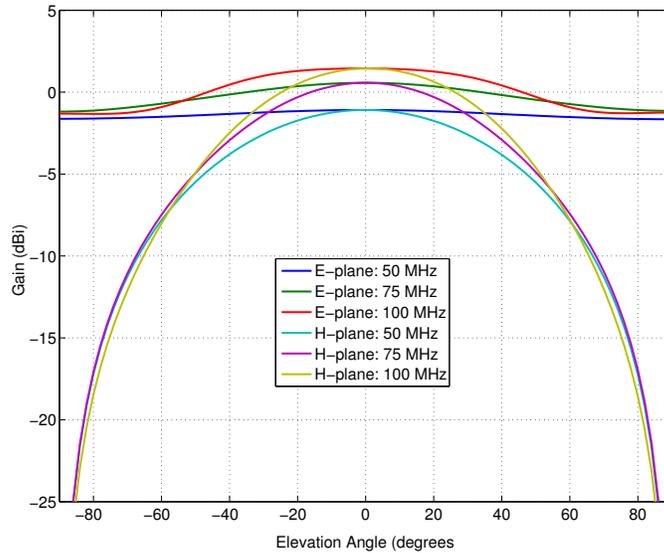


Figure 3: E and H-plane patterns of antenna. Over the frequency range of 20-100 MHz, the pattern remains essentially that of a magnetic dipole.

## 4 References

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