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Abstract – **We present a fully planar realization of a low profile dipole antenna placed on top of a high impedance substrate. The structure is made by three metal layers and it does not requires vias between the layers. The antenna is matched to 50 Ohms and has a gain of 7.1 dBi including all feed network losses.**

1 INTRODUCTION

Since the advent of electromagnetic bandgap materials and metamaterials researchers have tried to realize artificial magnetic conductors [1]-[5]. They could provide an effective shielding of radiation without requiring the vanishing of the tangential electric field. Various applications have been devised, such as for example, a shielding wall for a very low profile electric dipole [6]-[12].

In this work an artificial reflective surface is realized using a printed periodic structure on a planar substrate. This structure is similar to the one presented in [5] though its modeling is substantially different since in this case a subwavelength thickness is used for the dielectric layers.

The high impedance substrate is based on a new fully planar metamaterial comprising of pairing planar conductors, whose coupling provides the support of an anti-symmetric mode [13-16].

The present proposed design for the low profile dipole uses a three metal layer structure where the top layer allows the placement of any planar radiating element or array that can benefit from the artificial high impedance substrate (HIS) underneath, which is made by the other two metallic layers.

The HIS is used here to enhance the input impedance performance and gain of a transverse printed dipole, located very close to the metallic ground plane. The design is centered around 5.4 GHz corresponding to the wireless LAN frequency band.

First we introduce the HIS and its effective modeling using transmission line theory, then we present the design of a low profile folded dipole placed on top of the HIS.

2 REACTIVE HIGH IMPEDANCE SUBSTRATE

To design the HIS an equivalent transmission line model [10, 11], [15] is used at first, and results are compared with full wave simulations.

The equivalent transmission line (TL) network consists of a shunt L-C resonate circuit (Z_A in Figure 2) that resonates at the magnetic frequency [11, 12, 15], whereas the TL with characteristic impedance Z_0 represents plane wave propagation.

As an example, we consider the response of a layer of copper dogbones (Figures 1 and 2) printed on a dielectric grounded substrate with relative dielectric constant $\varepsilon_r = 2.2$ and losses tan $\delta = 0.0009$ (e.g., RT Duroid 5880). The various geometrical parameters characterizing the unit cell of the considered metamaterial (cf. Figure 1(b)) are as follows (in mm): $A = 8$, $B = 8$, $A1 = 0.5$, $B1 = 4$, $A2 = 7.8$, $B2 = 0.5$, $H = 1.6$. The thickness of the metal layers is 35 μm.

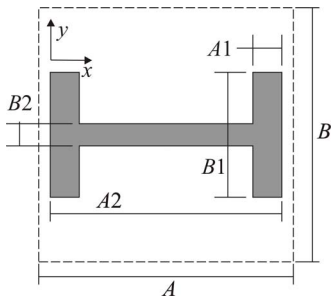

Figure 1: Unit cell of the HIS composed of a periodic array of dogbone-shaped copper conductors on a grounded dielectric substrate. *A* and *B* are the dimension of the periodic cell.

The dimension of the unit cell along the *z*-direction is $C = 4.0$ mm, and therefore the port is defined at $z = 4$ mm. The phase of the reflection coefficient *R* in Figure 3 predicted by the synthesized TL network is in good agreement with those obtained numerically with the commercial software CST Microwave Studio. The magnetic frequency is $f_m \approx 6.1$ GHz. Note that at the frequency of approximately 6GHz the phase is zero degrees, showing therefore that the artificial substrate

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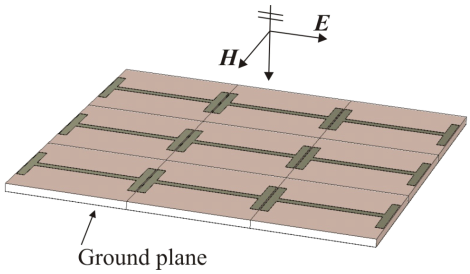
can be used as an artificial magnetic conductor. Some authors also pointed out that it could be advantageous to work in the region where the phase of the reflection coefficient is near 90 degrees [7] for antenna matching purposes. Other authors have however used different operation regimes to realize low profile antennas [9].



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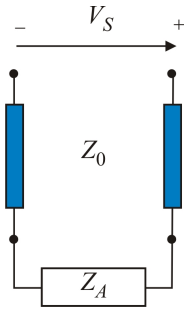


(b)

Figure 2: (a) HIS composed of a periodic array of dogbone-shaped conductors printed on a PEC-backed dielectric material and illuminated by a plane wave. (b) TL representation of the plane wave incidence on the HIS represented by Z_A.

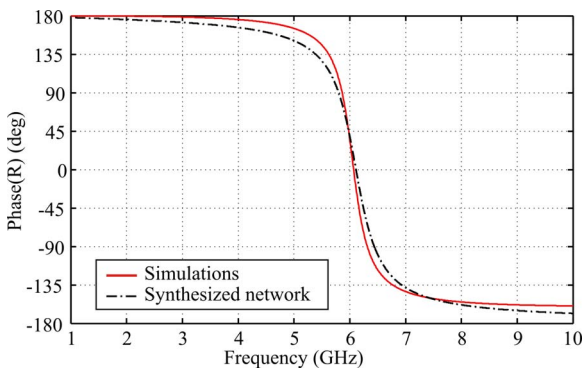
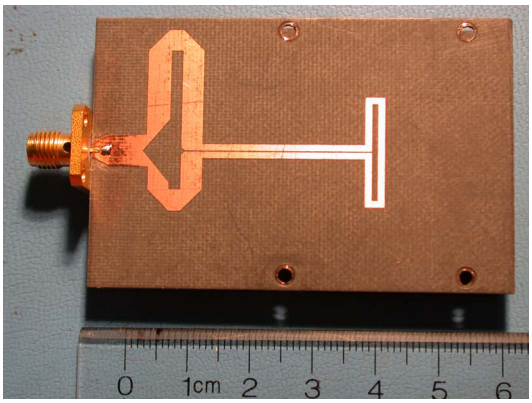


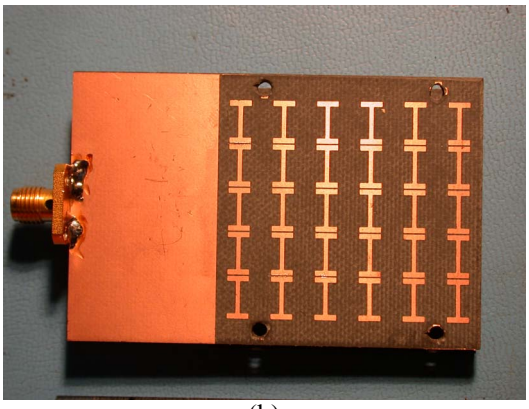
Figure 3: Phase of the reflection coefficient (at z = 4 mm from the bottom ground plane) for a normal plane wave incidence on the HIS with thickness H = 1.6 mm and permittivity $\varepsilon_r = 2.2$ with tan $\delta = 0.0009$. Results obtained through the equivalent L-C circuit approach (dashed-dotted lines) are compared with data from full-wave simulations (red continuous lines).

3 LOW PROFILE FOLDED DIPOLE ON A RECATIVE HIGH IMPEDANCE SUBSTRATE

The equivalent circuit developed above can be applied to the design of a low profile antenna consisting of a dipole located on top of a reactive high impedance substrate [8]. The high impedance exhibited by the artificial substrate permits to locate the metallic dipole very close to the surface itself without inhibition of the dipole radiation. Furthermore, the substrate prevents radiation from travelling across, resulting in a low profile antenna with high efficiency, even though the antenna system (comprising the dipole and thin metamaterial substrate) is integrated on top of an external lossy structure.



(a)



(b)

Figure 4: Fully planar realization of a low profile dipole antenna on top of a HIS. (a) top metal layer with a folded dipole and a balun from the unbalanced microstrip feed line to the balanced twin line. (b) intermediate metal layer (on the opposite side of the dipole substrate) showing the ground plane for the balun and the dogbone layer under the folded dipole. The dogbones are then located on top of a grounded substrate, not shown in this figure.

The structure has been designed to work at 5.5 GHz (for wireless LAN applications). The HIS composed of a periodic array of dogbones on a grounded dielectric substrate with thickness H = 1.6 mm and permittivity $\varepsilon_r = 2.2$ (namely, RT Duroid 5880) has been dimensioned to resonate slightly above this frequency. and is shown in Figure 4. The geometrical parameters in millimeters characterizing the unit cell of the considered periodic material (cf. Fig. 2) are as follows: A = 7, B = 7, $A1$ = 0.875, $B1$ = 3.5, $A2$ = 6.83, $B2$ = 0.7.

The geometry of the dipole over the finite-size HIS is shown in Figure 4. The finite-size HIS is composed of a 5×6 array of dogbone-shaped conductors. The folded dipole antenna has a length of 17.25 mm, gap of 1 mm, and strip width of 1 mm. The dipole is excited by a twin line with width of 1mm and a gap of 0.2 mm with a simulated characteristic impedance of 112 Ohm. The twin line is fed by a balun as shown in Figure 4. Note that the balun requires a large area, however, such printed dipole could also be fed directly with the balanced output of differential amplifiers (not shown here). The spacing of the dipole and the HIS is 1.6 mm. The thickness of the HIS is also 1.6mm, therefore the total thickness of the antenna (including the dipole and the HIS) is 3.2 mm. Further miniaturization can be accomplished, though not attempted here in this first fully planar realization.

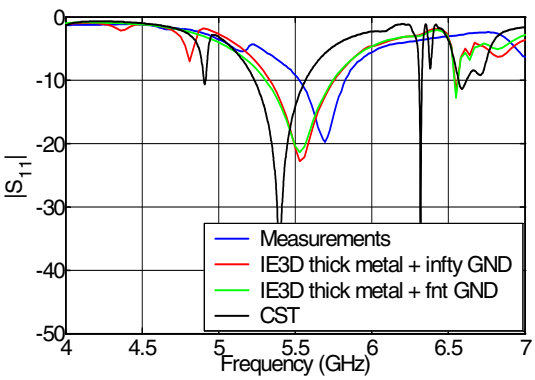
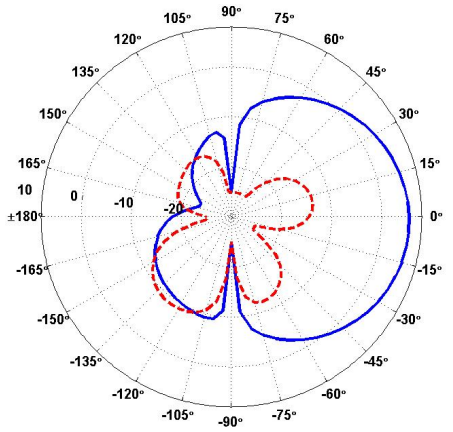
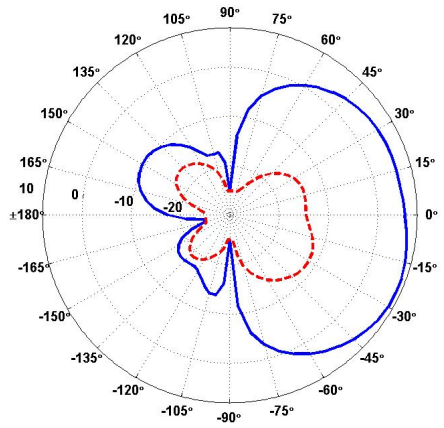


Figure 5: Simulated and measured reflection coefficient at the antenna connector (see Figure 4). Simulations are performed with CST Microwave Studio and with IE3D. In both cases the copper has a thickness of 35 µm. In IE3D the dielectric layer is infinite and both cases with infinite and finite ground plane have been simulated. The IE3D mesh is made by approximately 20 sub domains per wavelength (at a frequency of 7GHz).



(a)



(b)

Figure 6: Simulated radiation pattern: (a) E- and (b) H-planes are shown. Co-polar and cross-polar pattern components are plotted in solid and dashed lines, respectively.

The return loss of the structure was measured using an HP8510C vector network analyzer, and measured results are compared with simulated one in Figure 5. Good agreement is observed between data using two completely different simulators. The first one is based on time domain (CST from Microwave Studio) and the second one on method of moments (IE3D from Zeland)

The simulation with IE3D are carried out using infinite ground plane (red curve) as well finite ground (green curve). The simulation with CST instead takes into account also the finiteness of the dielectric substrates. All simulations and measurements show that at the center frequency the return loss is as good as −20dB. IE3D it is closer to the measured data in terms of resonance frequency and depth of the return loss. The frequency shift between simulations and measurements may be due to fabrication tolerances, since we used copper laminate having thickness of 35 µm, and etched away the metal where not needed. However the space between the twin lines is 200 µm

and between contiguous dogbones is 170 μm, and therefore comparable with the metal thickness. This high aspect ratio provides poor control in the gaps.

Figure 6 shows the radiation pattern in the *E* and *H* planes simulated using IE3D. The simulated peak gain is 7.1 dB including all losses in the feed network. The small tilt in the *H*-plane is likely due to the coupling (via scattering) between the balun and the dipole.

4 CONCLUSION

We have shown simulated and measured results of a fully planar realization of a low profile dipole antenna. The paper shows a working prototype and we believe that the total thickness realized here (3.2 mm at 5.5 GHz) can be further miniaturized. The measured return loss and simulated gain (7.1dB) confirm the good performance of our prototype and prove that the developed concept can be used to design antennas with miniaturized thickness in a full planar technology.

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