FDTD SIMULATION OF A WIDE-BAND HALF VOLUME DRA

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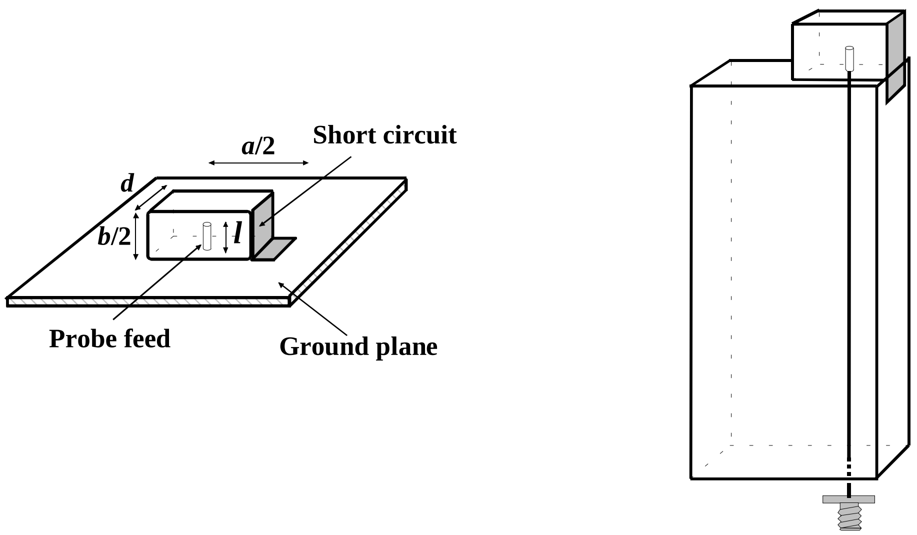
1. Introduction

The half volume dielectric resonator antenna (DRA) [1] is an interesting candidate to serve as a small wide-band antenna for new generation mobile phone handsets. The DRAs have relatively low losses and therefore they provide good radiation efficiency. In a previous study [2] it is observed that the bandwidth of a DRA is approximately proportional to the volume of the antenna in wavelengths. Thus, the main restriction in the miniaturization of the DRA is the availability and cost of suitable low-loss high-ε dielectric materials.

In this work we discuss the DRA design from the computational point of view. As a representative case we study a DRA manufactured from a material having $\varepsilon_r{}'=16$ and $\tan\delta=7\cdot10^{-4}$. We considered two models: 1) the antenna was placed above a finite ground plane, and 2) the antenna was placed on the top of a metal box representing a coarse mobile phone model. The configurations and dimensions are shown in Figure 1. We were mostly concerned with the S_{11}-parameters, which were measured with a vector network analyzer.



model 1) model 2)

Figure 1. Antenna configurations. In both models the resonator dimensions are $a/2$ = 17.5 mm, $b/2$ = 15.9 mm, d = 8 mm, and the feed probe length is l = 8.8 mm. In model 1) the ground plane size is 150 mm × 150 mm (in the simulation an infinite ground plane was used). In model 2) the phone model dimensions are 100 mm × 40 mm × 10 mm.

2. FDTD simulation

In the simulation of the structures, we put special effort on the proper modelling of the coaxial feed. The DRA was excited by the extended inner conductor of a 50 Ω SMA connector. The inner conductor itself, being of sub-cell diameter, was modelled by a thin-wire approximation [3]. The coaxial line was modelled as a separate 1D transmission line and the injection of the excitation to the FDTD volume was done following [4]. The S_{11}-parameters were extracted in two steps: first, only the 1D transmission line was simulated and the incident voltage was recorded at a certain reference point. This takes only a fraction of a second. Next, the actual FDTD volume was connected to the transmission line and the total voltage was again recorded in the transmission line at the same point. Subtracting the incident voltage from the total one we obtained the reflected voltage. Performing FFT to the reflected and incident voltages, we finally got the S_{11}-parameters for each frequency component.

We have one main supplement to the model in [4]: we do not have to assume that the outer radius of the coaxial line equals Δx. The only condition is that the inner radius $R_i<\min(\Delta x/2, \Delta y/2)$. Figure 2 shows one quarter of the cross-cut of the coaxial line in the level of the ground plane, along with the grid parameters. The parameters of Figure 2 are used in the model 1), while in the model 2) the parameters are Δx=1.333 mm, Δy=1.25 mm, R_i=0.465 mm and R_o=1.54 mm. Δz=0.795 mm in both models. The black dots represent the electric field locations in the aperture that transfer the TEM wave from the feed line to the FDTD volume. Because the electric field of the TEM mode in the coaxial line has the form

$$\vec{E} = \frac{V}{\ln(R_o / R_i)} \frac{1}{r} \vec{u}_r, \qquad (1)$$

we can easily evaluate the corresponding electric field components from the voltage of the feed line. The H_z-component is set to zero in the aperture.

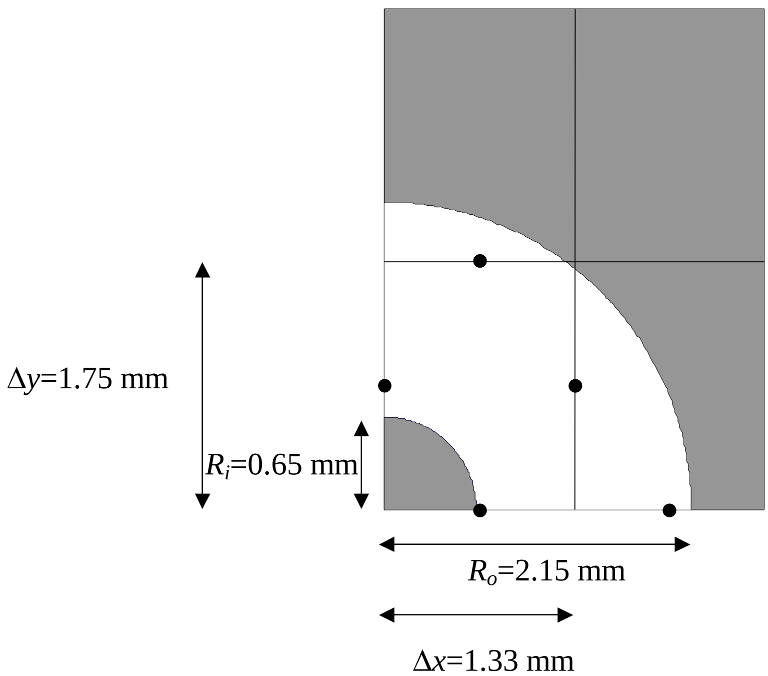


Figure 2. One quarter of the coaxial feed in the ground plane level and the corresponding FDTD grid. The black dots represent special electric field locations, where the field value is obtained from the voltage in the feed line according to Equation (1).

We studied the sensitivity of the S_{11}-parameters to several simulation parameters to find out the critical manufacturing and design factors. Let us consider model 1) first:

i) Due to short wavelengths in the DRA, we used very small grid parameters compared to free space wavelength. To avoid excessive simulation time, we brought the absorbing boundary conditions (8-cell PML) only 10 cells from the antenna, or about $\lambda/10$. Making this distance larger had no essential effect to the results.
ii) We studied the effect of numerical dispersion through a special compensation procedure, but no impact due to numerical dispersion was observed.
iii) We compared the models with and without the dielectric losses. Again, no essential differences were found.
iv) We included the conductivity losses of the metal surfaces at certain frequency, but this had also only a negligible effect.
v) We added an air gap near the joint of the short circuit plate and the ground plane, because this joint is not absolutely tight in practise. However, neither this had any essential effect.
vi) We varied slightly the width of the short circuit plate – without considerable effect.
vii) Finally, we added an **air gap between the feed probe and the dielectric**, resulting a significant impact to the results. A similar sensitivity is reported also in [5]. An efficient way to model thin coverings accurately with FDTD is given in [6], but in our problem the air cavity is modelled accurately enough by simply defining ε_r=1 in the electric field locations immediately adjacent to the probe axis.

There are two main resonances in the interesting band 2…4.5 GHz. Around 2.2 GHz we have the short circuit's first resonance, while at about 3.8 GHz there is the dielectrically loaded feed probe's monopole-type quarter-wave resonance. Because the dielectric dice has a high permittivity value, it is rather obvious that the volume of the air gap around the feed probe affects the loading critically. Indeed, perturbing the air gap moves the location and depth of the probe resonance several hundreds of MHz and several dB. In the point of view of the manufacturing process, it is difficult if not impossible to prevent the existence of the air gap, and because of the high sensitivity, repeatability problems are expected.

In the model 2), we found that the air gap has a less dominating effect, affecting mainly the resonance level, not so much the frequency. The matching has been improved by inserting a 2 mm thick piece of Teflon under the DRA (in the simulation the thickness equals $3{\cdot}\Delta z$=2.385 mm). The air gap in the model 2) prototype is smaller than in the model 1) prototype, and the simulation without any air gaps agrees well with the measurements.

The simulated and measured S_{11}-parameters are shown in Figures 3 and 4. The impedances in the Smith charts are referenced just below the DRA in both models. We observed that a very small change in the air gap influences the phase of the reflection a lot, explaining the systematic discrepancy between measured and simulated results in the Smith chart

3. Conclusions

The DRA has turned out to be a very interesting candidate for a small antenna of mobile handsets. Through proper design, an exceptionally wide bandwidth is obtainable, while still having good radiation efficiency. The FDTD method is well suited for modelling dielectric structures. In this paper we have studied the use of FDTD in the design of DRA. Especially, we have adopted an efficient way to model the coaxial feed, and discussed the significance of several physical factors in the DRA models. The most important factor turned out to be the air gap between the feed probe and the dielectric dice.

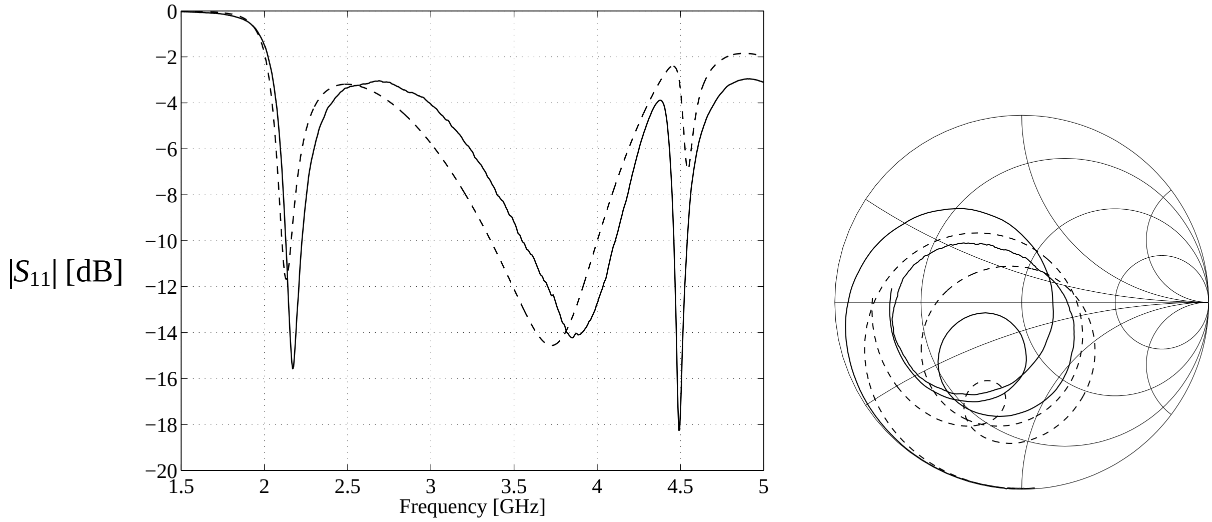


Figure 3. Simulated (dashed line) and measured (solid line) S_{11}-parameters. Model 1).

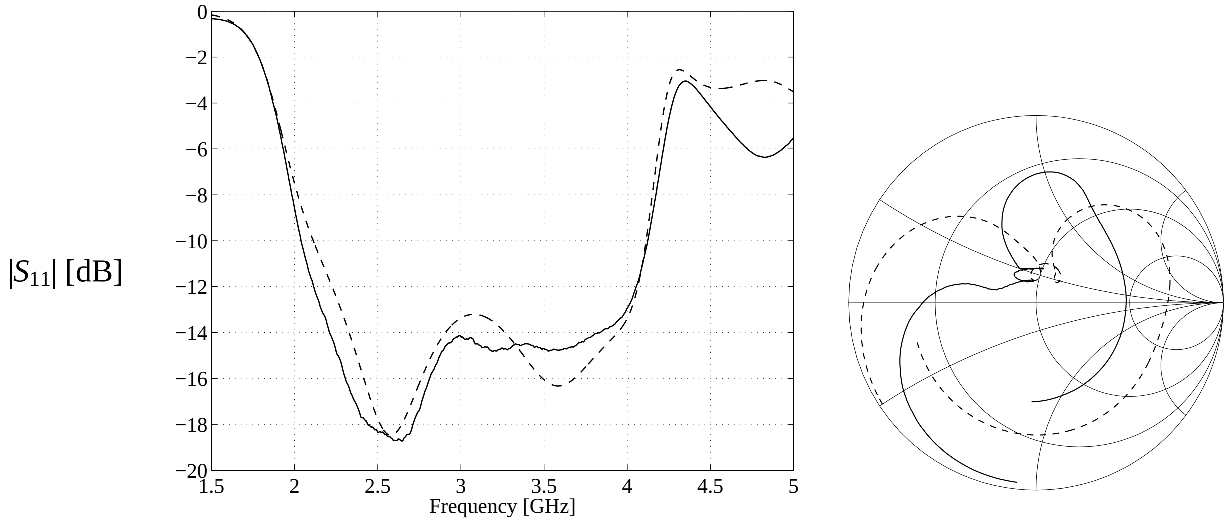


Figure 4. Simulated (dashed line) and measured (solid line) S_{11}-parameters. Model 2).

4. References

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