

Microwave Antennas and Circuits Modeling Using Electromagnetic Field Simulator

Pavel HAZDRA, Milan POLÍVKA, Vratislav SOKOL

Dept. of Electromagnetic Field, Czech Technical University, Technická 2, 166 27 Praha, Czech Republic

hazdrap@fel.cvut.cz, polivka@fel.cvut.cz, sokolv@fel.cvut.cz

Abstract. *Electromagnetic field simulators have become a widely used tool in a design process of microwave circuits and systems. A proper usage of electromagnetic (EM) field simulators allows substantial reduction of the design time providing reliable results. In such case the required parameters of the designed circuit can be reached even at the first manufactured prototype in spite of high complexity of the structure. However, EM simulation as a numerical process suffers from systematic and random errors similar to measurement using real equipment. Thus the setting of the EM-field simulator such as a frequency range, mesh properties, usage of PEC and PMC walls etc. has to be done with the highest attention and the simulation results have to be always verified using well-established techniques. The aim of the paper is to demonstrate the selected capability of EM-field simulators with a few examples of antenna and circuit modeling. Also an issue of reliability and simulation errors will be discussed.*

Keywords

CAD, EM-field simulation, microwave modeling, numerical methods.

1. Introduction

The successful engineering realization of high-frequency components and systems for microwave applications is heavily dependent on computer-aided design (CAD) in numerous respects. EM field simulators are an efficient tool for modeling of passive parts of microwave circuits with various geometry and material properties such as antennas, SMD components, filters, packages, resonators, directional couplers, multilayer circuits and so on. Even effective nonlinear design remains highly dependent on reliable and accurate characterization of the linear portions of circuits.

The intention of the paper is to demonstrate the selected capability of widely used EM-field simulators with a few examples of antenna and circuit modeling and focus on specific features that are able to simplify the EM modeling of the structure without lose of fundamental precision

and/or decrease computational time. Thus, the aim of the paper is not to present near-perfect agreement between measured and simulated results or try to find what the best tool in the market is. The aim is rather to offer some “know-how” about EM modeling using commercial simulators.

2. Overview of EM Field Simulators

Although it is not easy to give a comprehensive overview of commercially available EM field simulators on limited space let’s try to briefly classify them according to usually used criteria [1]. Also a brief description of numerical methods that are used to treat Maxwell’s equations will be offered.

2.1 Classification by Solution Domain

Numerical methods can be divided into three categories: frequency domain, time domain and eigenmode or modal solvers.

Solvers in frequency domain generally solve Maxwell’s equations in their integral form. Frequency domain solvers discretize the solution domain, build a matrix, invert matrix or use iteration to find the solution. The matrix building and its solution must be repeated for each frequency. The numerical method is typically Method of Moments (MoM) [2], [3], Finite element Method (FEM) [4], [5] or Spectral Domain Method (SDM) [6]. Frequency domain solvers are generally the choice for highly resonating structures. When broadband data are needed many frequencies must be computed or “fast sweep” option based on finding a rational polynomial that describes the solution behavior using a minimum set of computed frequency points. Care must be then paid to the reached solution particularly at the extremes of the frequency range.

The following list shows the most considerable ones: Zeland IE3D [7], Ansoft HFSS [8], Ansoft Designer [9], FEKO [10], SuperNEC [11], Momentum-ADS [12], EM-Sight-AWR [13], Sonnet Suite [14], Microwave Wizard [15] and FEMLAB [16] (which is a very general FEM based multiphysic system able to solve almost arbitrary PDE, thus the enlistment here isn’t unambiguous).

Time domain solvers employ Maxwell's equations usually (not in all cases like the finite integration technique) in their differential form. Basically, after the discretization the simulated structure is excited by time domain waveform, algebraic equations are used to update the field quantities as a function of space and time until convergence is reached and the reflected waves at a port or other point of interest are monitored. Using the discrete or fast Fourier transform, it's easy to convert the response to the frequency domain over the desired band. Indeed, highly resonant structures generate very long time response which may lead to uncomfortable simulation time. Both frequency and time domain solver assume solution problem with external source of energy, typically at a port. Typical methods are Finite Difference Time Domain (FDTD) [17], [18] Finite Integration Technique (FIT) [19] and Transmission Line Matrix (TLM) method [20], [21].

State-of-the-art time domain simulators are: CST Microwave Studio (CST MWS) [22], Semcad [23], IMST Empire [24], Fidelity [25], QuickWave [26], Mefisto [27].

Eigenmode or modal solvers calculate the eigennumbers (related to eigenfrequencies) and eigenmodes (eigen field distributions) of the related homogenous wave equation. This is usually the choice for modeling of resonators and some kind of antennas. Except the above, 2D modal analysis is also often used to calculate and identify modes of transmission lines before the main simulation starts. Modal analysis is available namely in the following solvers: Ansoft HFSS, FEMLAB and CST MWS (the algorithm used here allows also to calculate the response over the given band, using the so-called Fourier method).

2.2 Classification by Geometry

Another way of classification of numerical techniques and EM tools is based on the number of independent space variables/dimensions upon which the field and source functions depend. In all further mentioned categories we can again distinguish between frequency and time domain formulations.

1D methods – are used for solving problems where the field and source functions depend on one space dimensions only. Typical examples are transmission line problems, uniform plane wave propagation or spherically or cylindrically symmetrical problems with only radial dependence.

2D methods – are used for solving problems where the fields and source functions depend on two dimensions. Typical applications are planar structures such as cross-section problems in transmission lines and waveguides, coaxial TEM problems, spherical problems depending only on radius and azimuth or radius and elevation.

2.5D methods – these are methods for solving problems where the fields depend on three dimensions, while their sources are mainly confined conductive planes with two space dimensions. Arbitrary a number of infinite di-

electric layers with conductive interconnections among them is allowed. These tools are sometimes called 3D planar solvers by some vendors. Such a specification can be considered when arbitrary 3D conductive interconnection can be modeled.

3D – these are methods for solving problems where both fields and source functions depend on three space dimensions. This category comprises all volumic full-wave general-purpose formulations. The most used 3D frequency domain methods are FE, FD, and MoM. Among the 3D time domain methods, the FDTD, FIT, and TLM formulations dominate.

3. Antenna Modeling

3.1 Cassegrain Antenna

A following structure is an example of modeling electrically large (dimensions larger than several wavelength) objects. The selected structure is a so-called Cassegrain antenna consisting of main paraboloid, rectangular-horn feeder and hyperbolic subreflector (see Fig. 1 for details). Antenna's operating frequency range is 20 to 40 GHz with the highest simulation frequency f_{\max} of 35 GHz. The main dimensions are (in terms of wavelength at 35 GHz): the diameter of the paraboloid $D=32.7 \lambda$ and the height $h=10.5 \lambda$. Full-wave calculation was performed by using time domain EM simulator CST MWS.

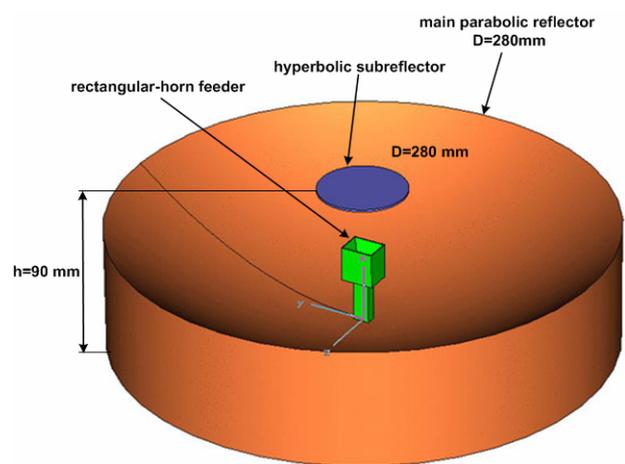


Fig. 1. Geometrical structure of the Cassegrain antenna.

Simulation of the whole structure is a very difficult task even for current computers. However, because the structure has 2 symmetry planes, it is possible to reduce the calculation domain (and also the simulation time) by factor of 3/4 using perfect electric condition (PEC) and perfect magnetic condition (PMC) symmetries (Fig 2). Including the distance between the bounding box and the structure (which is needed for proper forming of radiated waves), dimensions of the reduced calculation domain are now around $16.5\lambda \times 16.5\lambda \times 10.5\lambda$. Using 10 cells per wavelength for meshing,

the number of meshcells is 3.2 millions resulting in about 500MB of RAM demands.

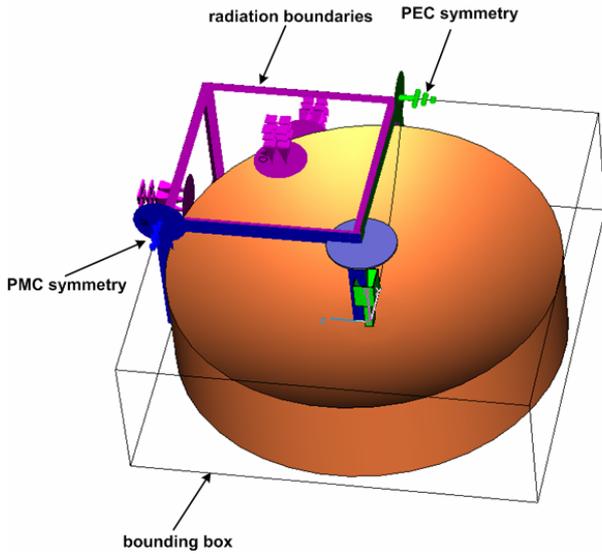


Fig. 2. Planes of symmetry used to reduce calculation domain.

Fig. 3. shows another specific result – the distribution of the near $|\mathbf{E}|$ field in the E-plane ($y=0$) cut with characteristic diffraction patterns. Calculated far-field performance is compared to the measured result, both at the frequency of 33 GHz (see Figs. 4 and 5). An agreement between simulation and measurement is satisfactory, even if we take into account that measurement has been done in an open-area test site; reflections from buildings around affected mainly the low-amplitude sidelobes. The agreement between calculated and measured directivity (37.6 dBi) and B_{-3dB} (2.28°) is excellent.

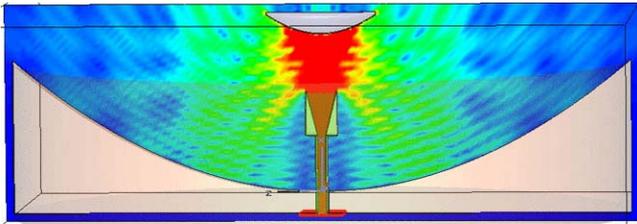


Fig. 3. E-plane cut showing near electric field @ 33GHz.

3.2 Modal Analysis of Dualband Patch Antenna Loaded by Slots

The simplest numerical approach to study patch antennas is to treat them as a 2D planar resonator lying in the XY plane [28]. We therefore assume that:

- the height of the patch above the ground plane is small thus only one component of the electric field (E_z) is dominant, the others are omitted,
- there is no fringing field at the boundary of the patch (i.e. the height of the patch above the ground is small). This condition results in Neumann boundary

condition for electric field $\partial E_z / \partial n = 0$, where n is the outer normal to patch's boundary,

- there's no feeder connected to the antenna, we are interested only in the eigen behavior.

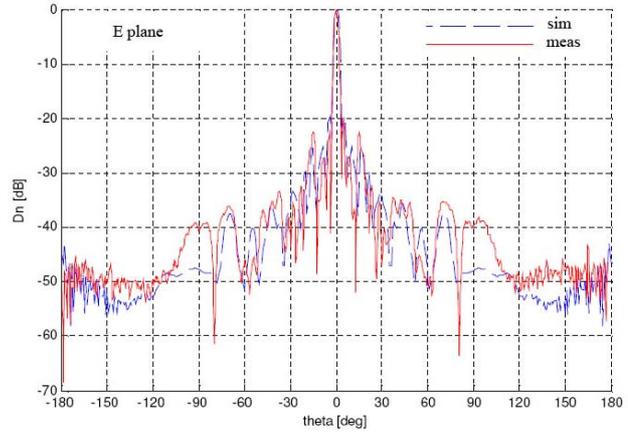


Fig. 4. E-plane farfield pattern cut @ 33GHz.

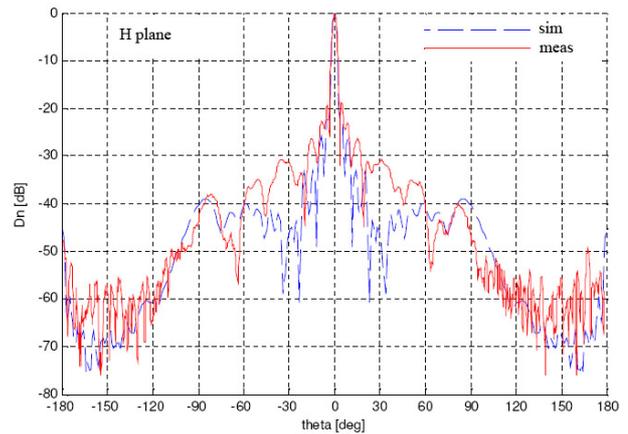


Fig. 5. H-plane farfield pattern cut @ 33GHz.

Under the above assumptions we are now able to write down the scalar Helmholtz wave equation

$$\frac{\partial E_z}{\partial n} + k^2 E_z = 0$$

and solve the PDE (usually employing FEM) for the set of eigenfields $E_{z,n}$ and eigenfrequencies

$$f_n = \frac{c \cdot k_n}{2\pi \sqrt{\epsilon_{\text{reff}}}}$$

Magnetic field (which is proportional to the surface currents flowing on the patch) is then obtained by the following equation

$$\vec{H} = \frac{1}{j\omega\mu_0} \vec{z}_0 \times \nabla E_z$$

Let's now consider a patch antenna with outer dimensions $L=29.62\text{mm}$, $W=18.78\text{mm}$ containing the two T-shaped notches (see Fig. 6.)



Fig. 6. The layout of the studied microstrip patch antenna with two T- notches.

Modal analysis (using FEMLAB software) was used to calculate the first 3 modes and compared with accurate results from full-wave EM simulator IE3D, based on MoM. (Fig. 7.) of the T-shaped notch microstrip patch antenna [28].

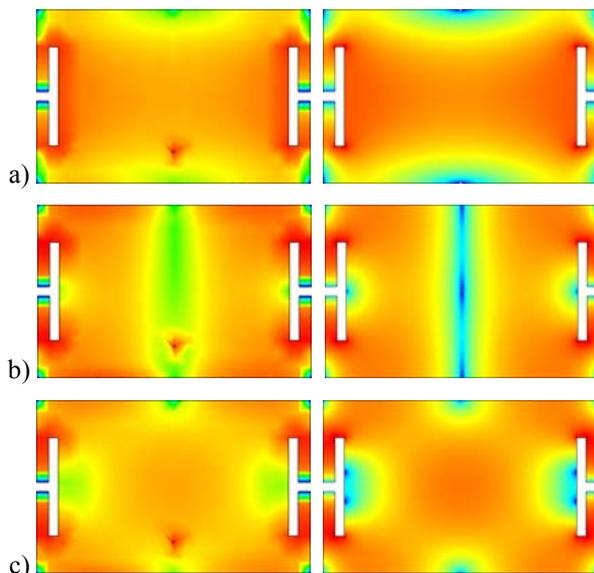


Fig. 7. Surface electric current \mathbf{J} of the first three modes, calculated by full-wave MoM (left) and using modal analysis with FEMLAB (right).

We have to note that the full-wave MoM solves the structure with feeder (coaxial probe in this case) included, thus the feeder is affecting the overall current distribution in some way [29]. Eigenmode solution of course doesn't take any feeding system into account.

Modal calculation of resonant frequencies is quite more sensitive due to several facts below:

- when microstrip antenna radiates, fringing field along its boundary is making electrical dimensions longer than physical ones. Fringing field effect decreases with lowering of substrate height. The Neumann boundary condition is only an approximation.
- radiation represents losses, thus the resonant frequency is shifted, depending on radiation Q_R of each mode.
- modal analysis doesn't take into account any coupling inside the structure. Accuracy can be decreased for structures with a lot of slots etc.

The following two tables show the results of resonant frequencies calculations performed by the modal analysis and full-wave MoM which is used as the accurate reference here. Tab. 1. is for the height of the patch above the ground plane $h=1$ mm, Tab. 2 for $h=0.5$ mm. Even that the studied antenna contains two T-shaped slots which are responsible for internal coupling, accuracy of the modal analysis is sufficient, with the average error 10.7 %. Better results (see Tab. 2) could be achieved by further decreasing of the height h , with the average error 6.9 % in resonant frequency estimation.

mode no.	f_{fullw} [GHz]	f_{modal} [GHz]	error [%]
1	6.10	6.94	13.8
2	8.67	9.35	7.8
3	9.838	10.86	10.4

Tab. 1. The comparison of resonant frequencies calculated by full-wave MoM and modal analysis for $h=1$ mm.

mode no.	f_{fullw} [GHz]	f_{modal} [GHz]	error [%]
1	6.44	6.94	7.8
2	8.87	9.35	5.4
3	10.09	10.86	7.6

Tab. 2. The comparison of resonant frequencies calculated by full-wave MoM and modal analysis for $h=0.5$ mm.

3.3 Aperture Coupled Microstrip Patch

Another feature that is able to simplify the model and to decrease computational time is the use of so called magnetic currents that actually represent voltage in slots. This feature is demonstrated in Zeland IE3D on modeling of aperture coupled patch antenna. In classical formulation the coupling slot is modeled as an aperture in a conductive plate forming the ground below the patch. In magnetic current formulation the only coupling slot is modeled (see Fig. 8 right). The patch dimensions are $W = 30$ mm, $L = 25$ mm, dimensions of coupling slot are $l_s = 20$ mm, $w_s = 0.5$ mm, dimension of the slot end extension is $w_{se} = 2$ mm, $l_{se} = 3$ mm. The patch itself is etched on the bottom side of the supporting substrate (Umatext 222, $\epsilon_r = 4.3$, $\tan\delta \sim 0.02$). The height of the patch over the ground plane of dimensions 120x120 mm with the coupling slot is 10 mm. Feeding microstrip line is at the bottom of dielectric substrate (GML 1000, $\epsilon_r = 3.2$, $\tan\delta \sim 0.003$) placed below the ground plane. The second metal plane working as a reflection plane has the dimensions 120x120 mm again and is placed 10 mm below the feeding microstrip line. The overlap of the feeding line over the coupling slot is 3 mm. The microstrip line is designed for $Z_{ch} = 50 \Omega$ and its width is 1.7 mm. In the EM model the following simplification has been done. In case of electrical modeling both the ground plane with the coupling slot and the reflecting plane have been modeled with real physical dimensions while in case of magnetic modeling both planes are supposed to be

infinitely large. The dielectric substrate is supposed to be infinitely large in the used MoM implementation as well. The antenna structure has been meshed with 16 cells per wavelength with the edge cells ratio of 0.05. The amplitude of the reflection coefficient for the mesh density of 16 cell per lambda are in a good agreement of both kinds of models however magnetic current modeling is not able to predict the resonance of the ground plane of a finite dimension (see the first local minima in frequency approx. 3.3 GHz in Fig. 9). The ratio of simulation time of electric to magnetic current modeling methods is 2173:2 (solution time in seconds per frequency).

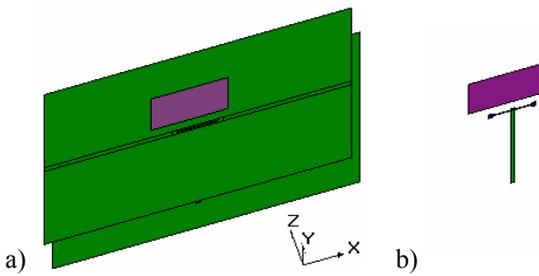


Fig. 8. Layout of aperture coupled patch antenna, a) electric current modeling, b) magnetic current modeling (reflection plane is not visible).

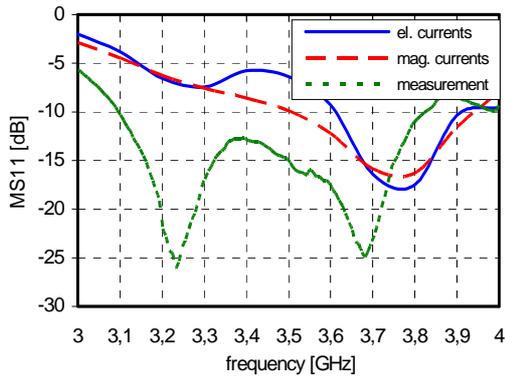


Fig. 9. Reflection coefficient of aperture coupled patch antenna, a) electric current modeling, b) magnetic current modeling, 16 cells/ λ with edge cells were used.

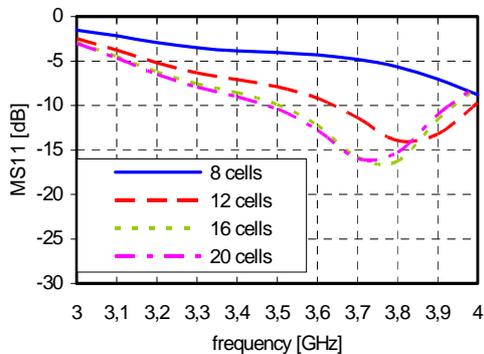


Fig. 10. Comparison of antenna reflection coefficient with magnetic current modeling of coupling slot with mesh density of 8, 12, 16 and 20 cells/ λ .

The difference between the measured and simulated (both electric and magnetic current modeling) magnitudes of the reflection coefficient can be seen. The sources of this disagreement can be found in the implementation of modeling of the finite ground plane when moment method is used.

Simulated reflection coefficients vs. frequency with magnetic current modeling with a mesh density as a parameter can be seen in Fig. 10. It can be seen that 16 cells can be quite enough for reasonable precision of modeling.

4. Circuit Modeling

4.1 SMD Resistor

Surface mounted devices are commonly used in a low-cost wireless microwave circuits in a frequency range up to tens of GHz. Although vendors often provide the s-parameters of SMDs the validity is restricted to the particular layout and substrate parameters. Nevertheless, general cases can be solved using either measurement based models [30], [31] or an EM field simulator.

The computation time can be reduced by exploiting the geometrical symmetry of the analyzed component in a computation domain. Thus, the PMC wall can be applied in this case, which is illustrated in Fig. 11. Moreover, the structure can be exited only once due to s-parameters symmetry. The most important difficulty is a modeling of the thin resistive layer of the SMD resistor. Usually, only one mesh cell is set by default leading to wrong simulation result. Therefore the local mesh refinement has to be applied over the thickness of the resistive layer. The mesh has to be generally set so that the additional refinement of the mesh influences the simulation result negligibly. Also the special technique dealing with improvement of the thin sheet simulation should be switched on, when it is provided by EM field simulator [32].

Finally, the multi-substrate equivalent circuit can be extracted based on resulting s-parameters [33], which is much faster process in comparison to measurement based models.

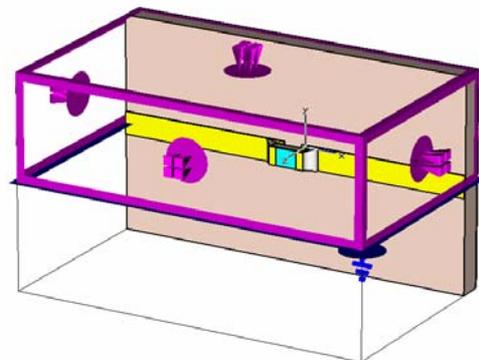


Fig. 11. Model of SMD resistor in CST Microwave Studio.

4.2 Transmission Lines

A proper modeling of microwave circuits strongly depends on correct characterization of transmission lines, namely impedance and effective permittivity.

The impedance evaluation is mainly influenced by discretization of a port cross-section and a size of its surface. The dependence of the microstrip line port impedance on a number of transverse cells over the strip width is illustrated in Fig. 12. The port impedance is generally given also by the port cross-section, which has to be scaled properly to the near-field region of the transmission line, see Fig. 13. The area of the port is too small in the first case ($k=1$) and the impedance value is wrong due to cropping of the near field region, while the impedance is computed correctly for two times longer base edge of the port ($k=2$), see Fig. 14. The conductive enclosure of the CPW port was applied in this case in order to ensure an excitation of the fundamental mode only. Similarly the air-bridges are used in real structures for this purpose.

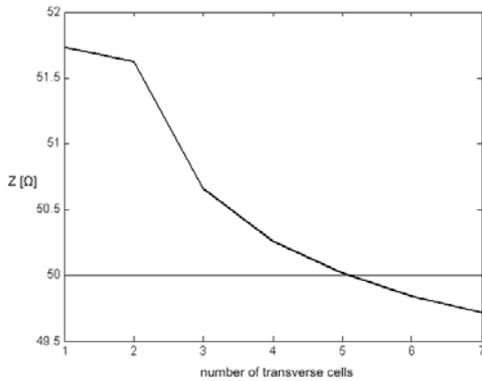


Fig. 12. Port impedance dependence on a number of transverse cells over the width of a strip at microstrip line ($w=0.635$ mm, $h=0.635$ mm, $\epsilon_r=10$).

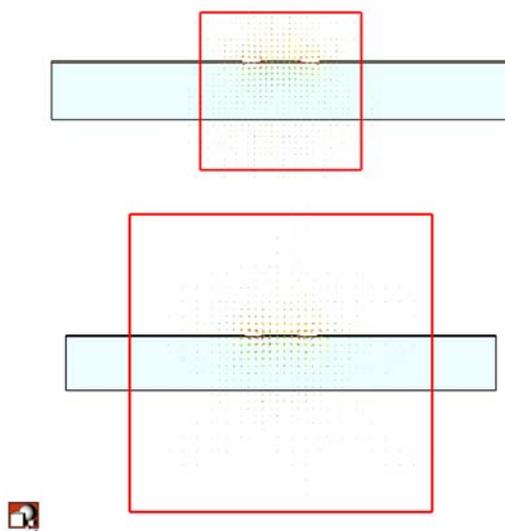


Fig. 13. Cross-section area of the CPW port. Scaling factor $k=1$ (top), $k=2$ (bottom).

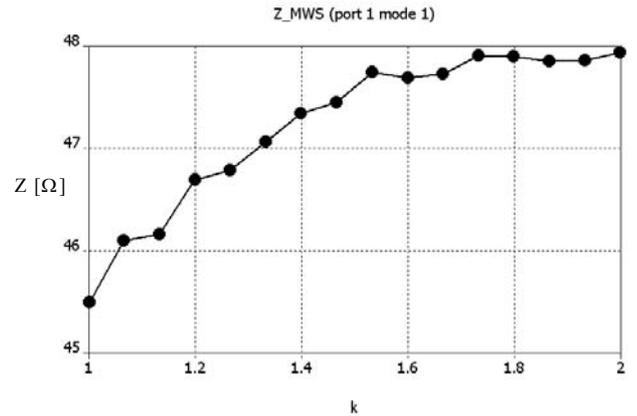


Fig. 14. Dependence of the CPW port impedance on the cross-section area (scaling factor k).

Next issue is frequency dependence of effective permittivity and impedance. Microstrip line is used as an example for the demonstration purpose. While the definition of effective permittivity is explicit the definition of impedance depends on quantities that are used for the evaluation [34], [35]. The effective permittivity is a growing function and all of CAD tools used in this test give nearly the same result depicted in Fig. 15. On the other hand, the impedance dependence is qualitatively different. While TXLine and AppCAD applications give a growing function both simulators CST MWS and HFSS give a decreasing function, see Fig. 16. The reason is that EM field simulators usually use for definition of the impedance power and squared current in contrast to the CAD applications, which use definition based on voltage-to-current ratio [35]. Therefore, the credibility of the numerical result has to be considered in conjunction with corresponding definition.

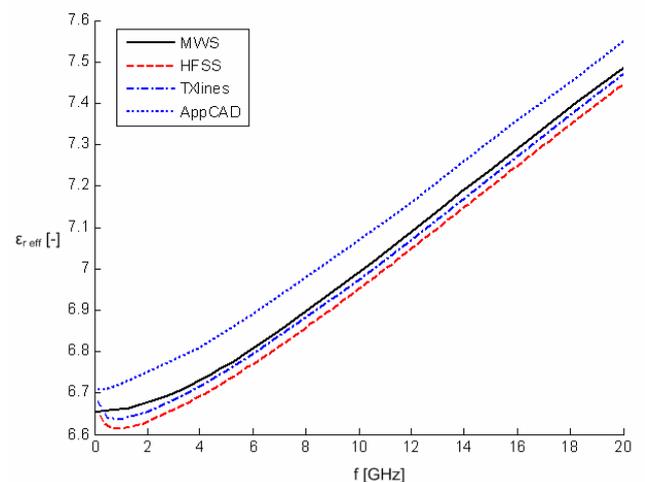


Fig. 15. Frequency dependence of effective permittivity for microstrip line ($w=0.635$ mm, $h=0.635$ mm, $\epsilon_r=10$) calculated in commercial softwares MWS [22], HFSS [8], TXLines [13], AppCAD.

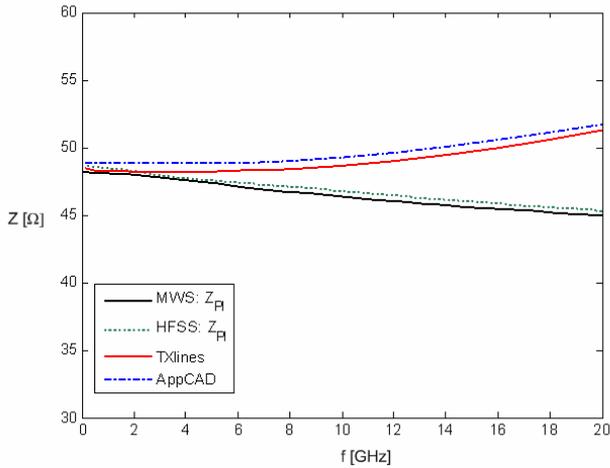


Fig. 16. Frequency dependence of impedance for microstrip line ($w=0.635$ mm, $h=0.635$ mm, $\epsilon_r=10$) calculated in commercial softwares MWS [22], HFSS [8], TXLines [13], AppCAD.

4.3 Dielectric Resonator

A dielectric puck is widely used as a high quality resonator for oscillators in a low noise block of satellite receivers. Usually, the dielectric resonator (DR) is coupled via magnetic field to the microstrip line, which is terminated by 50Ω load (usually SMD resistor), see Fig. 17. A behavior of the DR concerning fundamental resonant mode $TE_{01\delta}$ can be modeled as a parallel resonant circuit coupled via serial connected transformer to the microstrip line in certain reference planes [31].

Since the mechanical dimensions of the DR are smaller than its near field region the simulation setup has to be done with special care for port location. When the port is placed too close to the DR the energy is drawn out through the port cross-section resulting to lower quality factor. Moreover, the resonant frequency is shifted down. The problem of the near field disturbance by the port is illustrated in Fig. 18.

It seems to be the best solution to place the port far enough from the DR. However, the phase error problem of the de-embedding process will appear when the shift of the reference plane is applied over long distance [36]. Fortunately, the phase error can be determined using so called “thru test” where the zero length transmission line is used [1], [36].

Another difficulty of the DR simulation is its high quality factor (approximately thousands) when it is analyzed in a time-domain electromagnetic field simulator. In this case it is strongly recommended to apply autoregressive filter in order to decrease the computation time obtaining smooth curves of s-parameters [32]. The issue of the suitable solver choice for highly resonant structures is discussed in more detail in the next example.

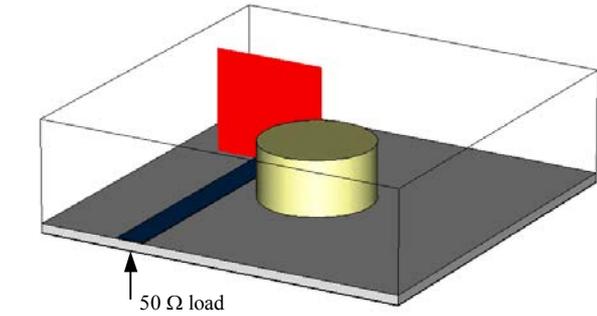


Fig. 17. Dielectric resonator coupled to the thru microstrip line. ($D=6$ mm, $h=3$ mm, $\epsilon_r=30$). The port cross-section is marked by the red rectangle.

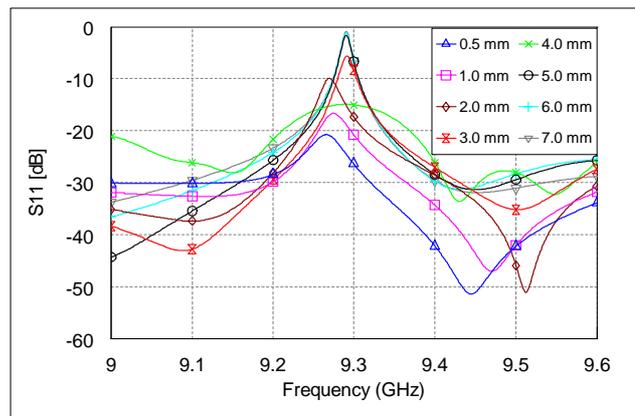


Fig. 18. Simulated reflection coefficient for variable spacing 0.5 - 7.0 mm of the port from a mechanical border of the dielectric puck.

4.4 Planar Microstrip Filter

Planar microwave filters represent a group of circuits where mechanical dimensions along with losses in dielectrics and conductors play an important role affecting essentially the performance. Although modern circuit simulators provide various tools for the design of planar microwave filters, an EM field simulator should be used for the performance verification because the full-wave simulation takes into account also the mutual coupling between sub-elements of the filter and the influence of the package.

A planar bandpass microstrip filter depicted in Fig. 19 was used for a comparison of time-domain and frequency domain EM field simulator. As a time-domain simulator CST Microwave Studio transient solver with autoregressive filter option was used while IE3D MoM solver was used as a frequency domain simulator. Both simulators take into account both dielectric and conductor losses.

The measured data obtained with Agilent E 8364A VNA (vector network analyzer) were used as a reference

solution, which is depicted along with simulated results in Fig. 20 and Fig. 21. A better agreement was achieved with the frequency domain IE3D simulator. The influence of attached SMA connectors is negligible at frequency 2 GHz, thus it was omitted in the comparison of s-parameters.

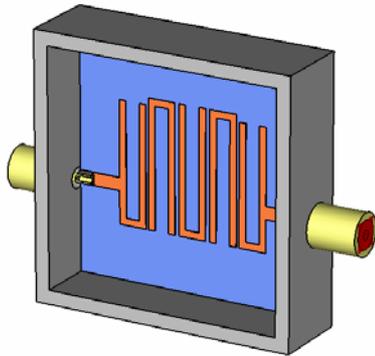


Fig. 19. Planar bandpass microstrip filter, center frequency 2 GHz.

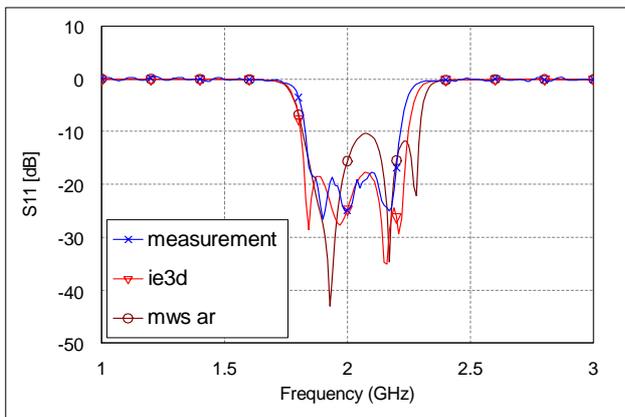


Fig. 20. Reflection coefficient of the planar bandpass filter calculated in IE3D, MWS (with autoregressive filter) and measured with Agilent E 8364A.

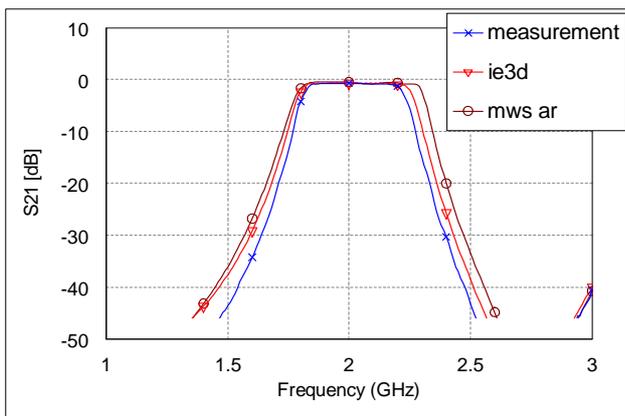


Fig. 21. Transmission coefficient of the planar bandpass filter calculated in IE3D, MWS (with autoregressive filter) and measured with Agilent E 8364A.

5. Reliability of the Results

A reliability of the computed results depends on many factors that can be seen from previous examples. However, a proper mesh setting can be considered as the most important factor. Unfortunately, there is not any general rule how to set optimally the mesh and the best setting depends strongly on experience of the operating personnel along with features of the particular EM field simulator. Even the widely spread opinion that the finer mesh ensures the more accurate result is questionable because modern EM simulators have often built-in techniques, which effectively reduce a number of the mesh cells ensuring the same or even better accuracy. Nevertheless, the problem of the accuracy estimation can be partially solved using either calibration process of an EM field simulator or mesh adaptation technique.

The optimal setting of the EM simulator, such as mesh density, frequency range and so on, can be evaluated using some geometrically simple structure with known analytical solution. This process where the result from the simulator is compared with the exact analytical solution is called calibration [1]. After that the same setting of the simulator is applied to the similar structure where the analytical solution does not exist.

Second possibility of the error estimation is a mesh adaptation technique. The mesh is gradually refined whereas the difference between current and previous iteration is evaluated. The difference error can be approximately considered as an error of the simulation. For instance CST MWS provides two refinement strategies. The first is based on field energy density while the second is expert based process [32].

Usually, the best verification of the simulated results is provided by combination of the calibration process, adaptive mesh technique and an experimentally obtained data.

6. Conclusion

A short list of commercially available EM field simulators along with brief overview of most often used numerical methods was provided in this paper. The capability of EM field simulators was demonstrated on a few selected examples of microwave antenna and circuit modeling. The weaknesses of the simulation and the importance of the verification process were also highlighted. An application of several useful techniques was illustrated on practical examples of the modeling. As it can be seen from the presented examples EM modeling requires some experience with proper setting of simulator to acquire reasonable results. Nevertheless, a proper usage of an EM field simulator provides a high-efficient design.

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About Authors...

Pavel HAZDRA was born in Prague in 1977. He graduated from the Czech Technical University in Prague in 2003. In the same year he joined the Department of Electromagnetic Field as an assistant. He is finishing his postdoctoral studies. His research activities cover microstrip antennas, fractal antennas and theory of EM field.

Milan POLÍVKA graduated from the Czech Technical University in Prague in 1996. In the same year he joined the Department of Electromagnetic Field as an assistant. He received his Ph.D. in Radio Electronics in 2003 at the Department of Electromagnetic Field. He is a member of IEEE. His research interests are in the field of antenna and radiating systems.

Vratislav SOKOL received the M.Sc. and Ph.D. degrees in Radioelectronics from the Czech Technical University in Prague in 2000 and 2004, respectively. He was a Post-Doctoral Research Associate with the Cork Institute of Technology in Ireland where he dealt with an ultra-wideband transceiver. His research focuses on microwave circuit design and modeling using electromagnetic field simulators. His research interests also include precise microwave measurement and calibration. He was also a consultant with the Radar Department at the Czech Hydrometeorological Institute in 2001-2004.