Efficient Modeling of Towel Bar Antennas Using Model of Distributed Loading along Wires

Milos M. Jovicic [1], Saad N. Tabet [2], and Branko M. Kolundzija [3]

[1] WIPL-D d.o.o., Belgrade, Serbia
milos.jovicic@wipl-d.com

[2] NAVAIR, Patuxent River, MD, USA
saad.tabet@navy.mil

[3] Chair of General Electrical Engineering School of Electrical Engineering, Belgrade, Serbia
kol@etf.bg.ac.rs

Abstract **─** This paper presents an efficient technique to determine equivalents of towel bar antenna dielectric standoffs in the form of wires with distributed loadings using WIPL-D Pro (3-D EM solver) software. Starting from the product, we will determine its basic characteristics and propose simplifications in modeling for further analysis. Benefits of this technique are simplicity of modeling and fast, but still accurate, simulations.

Index Terms **─** Antenna modeling, dielectric standoff, distributed loading, WIPL-D Pro.

I. INTRODUCTION

Let us consider a simple towel bar antenna, mounted above a Perfect Electric Conductor (PEC) ground plane using dielectric standoffs, as shown in Fig. 1. The antenna is excited at one end and short-circuited to the PEC ground at the other. Hence, the antenna uses the PEC ground to operate as a loop antenna. In general, the larger the conductivity of the ground below the antenna tube is, the better the antenna efficiency will be [1].

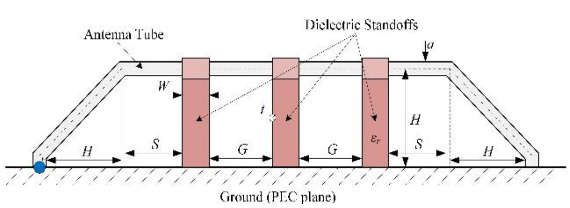

Fig. 1. Towel bar antenna mounted above a PEC ground plane.

The antenna tube is capacitively coupled to the ground. The capacitive coupling along the parts of the tube containing standoffs is different from those without standoffs. The influence of each dielectric standoff can be emulated by a loaded wire with properly determined equivalent radius. The equivalent radius and distributed loading of the wire are determined by the cross-sectional dimensions of the dielectric standoff and its electrical properties (i.e., relative dielectric constant).

In order to determine the equivalent values of the distributed loading on equivalents wires, the relative permittivity of the antenna standoffs should be determined. Capacitance per unit length of the part supported by the standoffs can be deduced from the measurements of the commercially available sample. The capacitance per unit length can be used to determine the relative permittivity of the dielectric standoff in the following way. The model of the antenna part supported by the standoffs is simulated with a variable value for the relative permittivity. The value of the capacitance per unit length is extracted from the simulations and compared to the value obtained by measurements. The relative permittivity of the standoff part is found for the case where the extracted value matches the value obtained by measurements. The capacitance per unit length is calculated using equation (1),

$$C' = \frac{1}{cZ_C}, \tag{1}$$

where c is the speed of the TEM wave propagating on the transmission line and Zc is the characteristic impedance of the line. These two parameters can be obtained from WIPL-D Pro numerical simulations.

II. EVALUATION OF CAPACITANCE PER UNIT LENGTH

If the part of the antenna supported by dielectric standoffs is set to be an eighth-wavelength long transmission line, as presented in Fig. 2, the characteristic impedance and the speed of the TEM waves can be determined using the simple transmission line equation

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provided in equation (2):

$$Z(D) = Z_C \frac{Z_P + jZ_C \tan(\beta D)}{Z_C + jZ_P \tan(\beta D)}. \qquad (2)$$

For the case where the length of the transmission line D is chosen to be:

$$D = \lambda / 8, \qquad (3)$$

the tangent in equation (2) is equal to one, and the input impedance depends only on the transmission line termination and the characteristic impedance, as shown in equation (4):

$$Z(D) = Z_C \frac{Z_P + jZ_C}{Z_C + jZ_P}. \qquad (4)$$

The choice of short and open circuit terminations can considerably simplify the calculation of the line characteristic impedance.

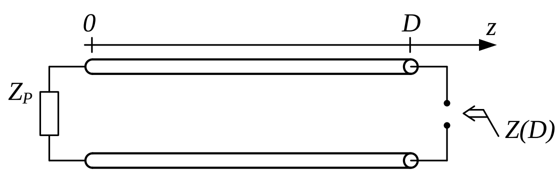

Fig. 2. Model of the transmission line section.

A. Characteristic impedance determination

For a short circuit termination, equation (4) reduces to:

$$Z_{SC}(D) = jZ_C. \qquad (5)$$

On the other hand, when an eighth-wavelength transmission line is terminated in an ideal non-radiating open circuit, the input impedance becomes:

$$Z_{OC}(D) = -jZ_C. \qquad (6)$$

In both cases, the magnitude of the input impedance is equal to:

$$|Z_{SC}(D)| = |Z_{OC}(D)| = Z_C. \qquad (7)$$

Over a narrow frequency range, the characteristic impedance (Z_C) is determined by locating the point of intersection between the input impedance magnitude of the short circuit and open circuit termination plots, as shown in Fig. 3.

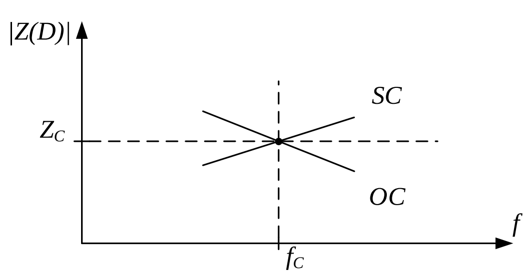

Fig. 3. Input impedance of an eighth-wavelength transmission line in a narrow frequency range for short and open circuit terminations.

B. Speed of TEM wave on transmission line

The magnitudes of the input impedance for open and short circuit terminations are equal, only when the length of the transmission line is an eighth-wavelength. Therefore, the speed of the TEM wave propagating on the transmission line can be determined from:

$$c = 8Df_C. \qquad (8)$$

C. Thin wire above an infinite PEC ground plane

1) Theoretical approach

As a proof of concept for computing the per unit length equivalent capacitance, we will use the thin wire above an infinite PEC ground plane model shown in Fig. 4. This is a good benchmark, since the capacitance per unit length for this case is computable in closed form using equation (9),

$$C' = \frac{2\pi\varepsilon_0}{\ln \frac{2h}{a}}. \qquad (9)$$

Some relevant dimensions used in this model are shown in Table 1.

Table 1: Model dimensions of a thin wire above an infinite PEC ground plane

Parameter [unit]	Value
a [mm]	12.7
h [mm]	609.6

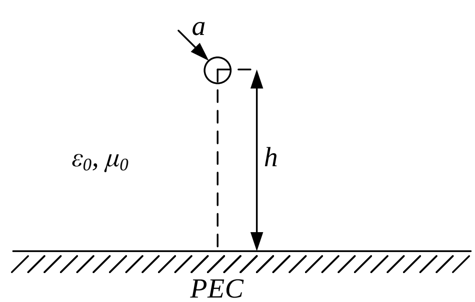

Fig. 4. Cross section of thin transmission line above the PEC infinity plane.

For the case of a thin wire above an infinite PEC ground plane with the dimensions as listed in Table 1, the capacitance per unit length is:

$$C'_{Theory} = 12{,}18 \text{ pF/m}. \qquad (10)$$

2) Numerical approach

In order to illustrate the calculations of the capacitance per unit length from simulations, two separate models of an eighth-wavelength transmission line were created. The first corresponding to a short circuit termination (PEC), and the second corresponding to an open circuit termination (Perfect Magnetic Conductor - PMC). The transmission lines in both models were 3.75 m long, with the same cross-section as the model of the thin wire above an infinite PEC ground plane discussed earlier. The narrow band frequency range of interest for both models is from 7.5 to 12.5 MHz. Both models are shown in Fig. 5.

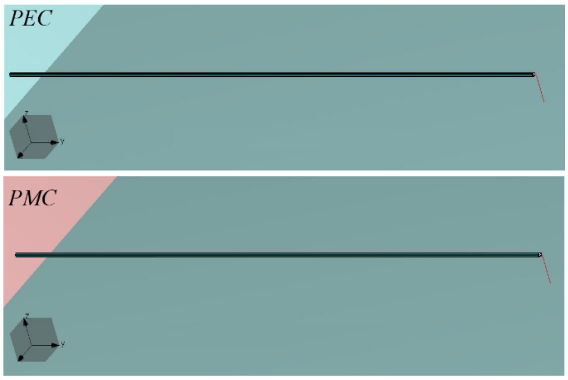

Fig. 5. Model of section of a thin wire above an infinite PEC ground plane in WIPL-D.

For short circuit termination, we used an infinite PEC plane, which forces the tangential component of the electric field to zero at the location of the PEC plane. A PMC infinite plane was used to model an ideal open circuit termination, since it forces the tangential component of the magnetic field to zero. It also enables the open circuit termination in precisely defined plane without radiation at the end.

Input impedance magnitude plots for short and open circuit termination models are shown in Fig. 6. The coordinates of their point of intersection are provided in Table 2.

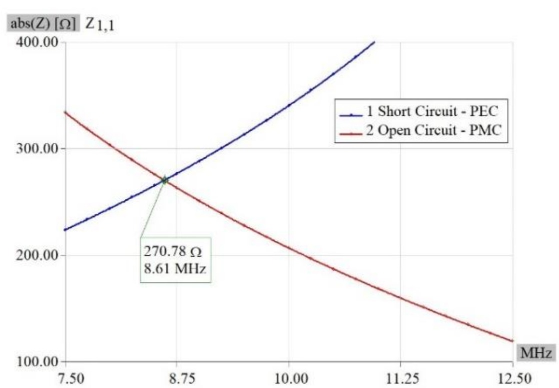

Fig. 6. Input impedance magnitude plots for short and open circuit terminations.

Table 2: Coordinates of short and open terminations point of intersection

Parameter [unit]	Value
Z_C [Ω]	270.78
f_C [MHz]	8.61

Placing the values in Table 2 into equation (8), the speed of the TEM wave in the transmission line becomes:

$$c_{WIPL-D} = 2.583 \cdot 10^8 \text{ m/s}. \tag{11}$$

The computed speed of the TEM wave in the transmission line obtained from the WIPL-D simulations is lower than the speed of light, even though the models were fully placed in vacuum. The reason for this disagreement is mainly caused by the feed region of the transmission line. Namely, instead of using plane wave excitation, we used a spatial delta generator in WIPL-D. This means that the generator is placed at one single point at the end of the feed wire, closest to the PEC plane placed below the transmission line. The total length of the conductor above the infinite PEC plane is equal to the sum of the transmission line length and the feed wire length.

If we combine the speed of the TEM wave in the transmission line obtained from equation (11), and the characteristic impedance from Table 2 into equation (1), we obtain the capacitance per unit length in equation (12) below:

$$C'_{WIPL-D} = 14.207 \text{ pF/m}. \tag{12}$$

As expected, due to the incorrect value of the speed of the TEM wave in the transmission line, the calculated value of the capacitance per unit length is also incorrect. However, if the value of the speed of light is used in equation (1), instead of the speed of the TEM wave obtained from the WIPL-D simulations, the result becomes:

$$C'_{WIPL-D\_corrected} = C'_{WIPL-D} \cdot \frac{c_{WIPL-D}}{c_0} = C'_{WIPL-D} \cdot K_{cal} = 12.23 \text{ pF/m}, \tag{13}$$

where K_{cal} represents a calibration factor defined as follows:

$$K_{cal} \triangleq \frac{c_{WIPL-D}}{c_0} = 0.86. \tag{14}$$

The relative error between the capacitance per unit length obtained from the equation and that obtained via CEM modeling is:

$$\rho = \frac{\left| C'_{Theory} - C'_{WIPL-D\_corrected} \right|}{C'_{Theory}} = 0.41\%. \tag{15}$$

III. EMULATION OF THE RELATIVE PERMITTIVITY OF THE DIELECTRIC STANDOFF

The measured capacitance per unit length of the dielectric standoff of the towel bar antenna is approximately:

$$C'_{Standoff} \cong 16.6 \text{ pF/m}. \tag{16}$$

Two new WIPL-D models representing a towel bar antenna supported by dielectric standoffs were devised in a similar fashion to the models of a thin wire above infinite PEC and PMC planes, using the dimensions listed in Table 3. The new WIPL-D models are illustrated in Fig. 7.

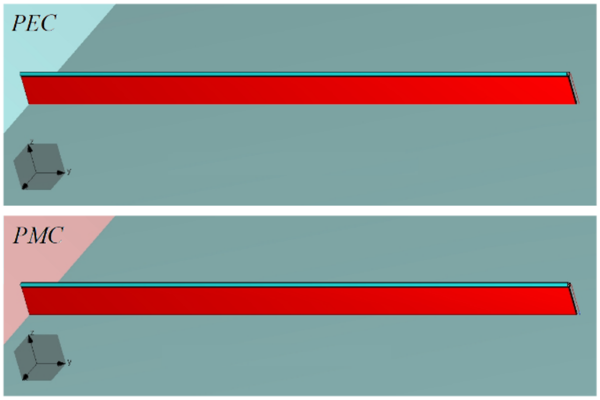

Fig. 7. Transmission line models a towel bar antenna supported by dielectric standoffs, over PEC and PMC planes, in WIPL-D Pro.

Table 3: Model dimensions for the WIPL-D models in Fig. 7

Parameter [unit]	**Value**
a [mm]	12.7
h [mm]	609.6
t [mm]	12.7

In order to determine the relative permittivity of the dielectric standoff, a set of simulations was run using the "Sweep" option in WIPL-D Pro. The relative permittivity value was swept from 8 to 20, in increments of 4.}. For each of the relative permittivity values analyzed, the characteristic impedance and the frequency at the point of intersection were determined. A plot showing the functional dependency of the extracted capacitance per unit length versus relative permittivity is presented in Fig. 8.

For a dielectric permittivity of 16, the extracted value of the capacitance per unit length, $C'_{Emulated} = 16.7\ \text{pF/m}$, approximately equals the dielectric standoff's measured value.

Finally, when all the dimensions and the relative permittivity of the dielectric standoffs of the towel bar antenna are known, a full three-dimensional (3-D) model of the towel bar antenna can be created.

IV. MODELS OF A TOWEL BAR ANTENNA IN WIPL-D PRO

A. Metal-dielectric model of a towel bar antenna

A full 3-D metal-dielectric (M-D) model of a towel bar antenna in WIPL-D Pro is shown in Fig. 9. In order to reduce the required number of unknowns and simulation time in WIPL-D, two symmetry planes are used; PEC in the *x0y* plane and symmetry in the *y0z* plane.

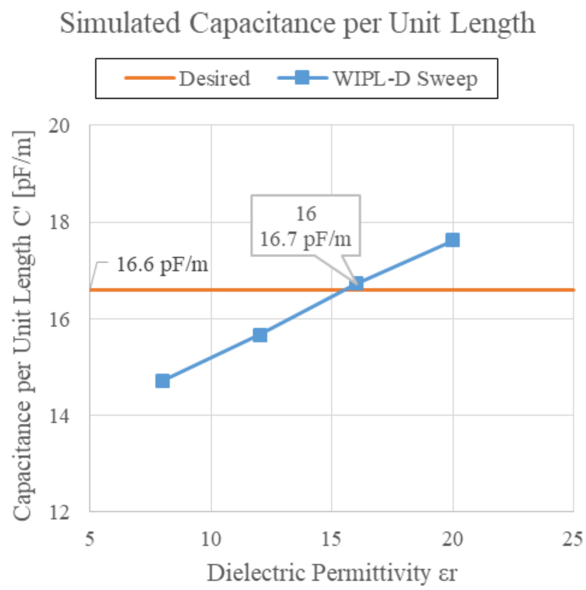

Fig. 8. Simulated capacitance per unit length.

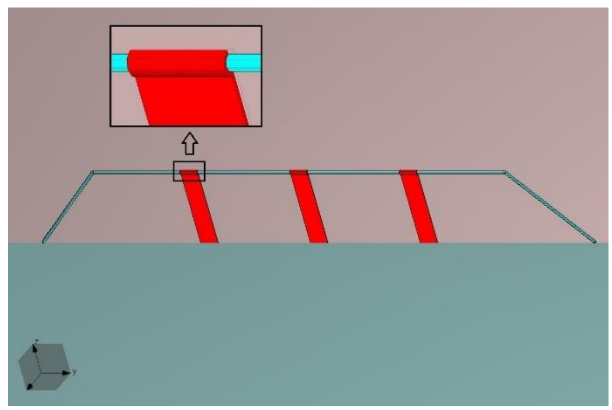

Fig. 9. Metal-dielectric model of a towel bar antenna in WIPL-D Pro.

The dimensions of the model and the electrical properties of the dielectric standoffs are provided in Table 4.

Table 4: Dimensions of the model and electrical properties of the dielectric standoffs

Parameter [unit]	**Value**
a [mm]	12.7
H [mm]	609.6
S [mm]	750
W [mm]	150
t [mm]	12.7
G [mm]	800
εr	**16**

B. Equivalent wire model of a towel bar antenna

An equivalent wire model of an M-D towel bar antenna, in WIPL-D Pro, is shown in Fig. 10. In this model, only PEC in the *x0y* plane is applied.

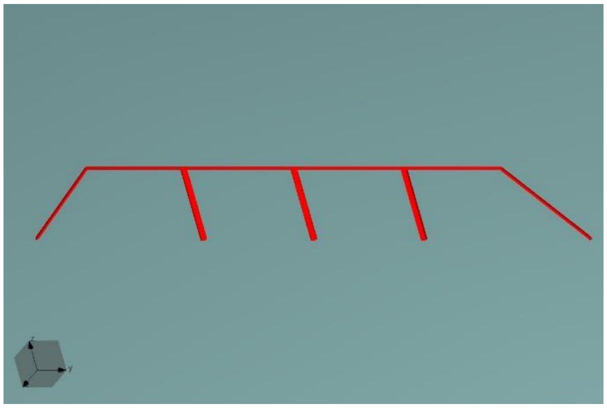

Fig. 10. Equivalent wire model of a towel bar antenna in WIPL-D Pro.

As a rule of thumb, the surface area of the cross-section of the dielectric standoff must be equal to the surface area of the cross-section of its equivalent wire. Therefore,

$$R_{equivalent} = \sqrt{\frac{W \cdot t}{\pi}} \,. \tag{17}$$

Also, the surface capacitance of the equivalent wire is related to the standoff's relative dielectric constant as shown in equation (18) below [2]:

$$C_{S,equivalent} = \frac{\varepsilon_0 (\varepsilon_r - 1) R_{equivalent}}{2} \,. \tag{18}$$

V. RESULTS

The values of the equivalent radius obtained from equation (17) and the surface capacitance obtained from equation (18), respectively, are provided in Table 5. All other dimensions remained the same as those of the M-D model. It is to be noted that the equivalent wires are placed along the center of the standoffs.

A comparison of the real and imaginary components of the reflection coefficient, for the models in Figs. 9 and 10, is shown in Fig. 11.

Table 5: Equivalent radius and surface capacitance

Parameter [unit]	Value
$R_{\text{equivalent}}$ [mm]	24.625
$C_{\text{S,equivalent}}$ [pF]	1.634

Furthermore, the surface capacitance in the equivalent wire model can be manually tuned to achieve better matching to the results of the M-D model. Figures 12 and 13 show results for the reflection coefficient and radiation pattern comparisons, respectively, for the M-D model and the equivalent wire model with $C_{S,equivalent}$ = 2.452 pF.

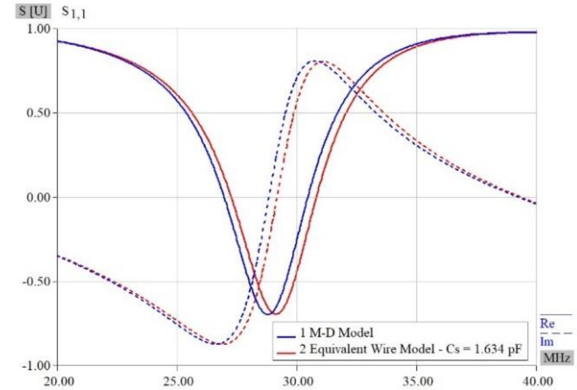

Fig. 11. Comparison of real and imaginary components of the reflection coefficient of the M-D and equivalent wire models, $C_{S,equivalent}$ = 1.634 pF.

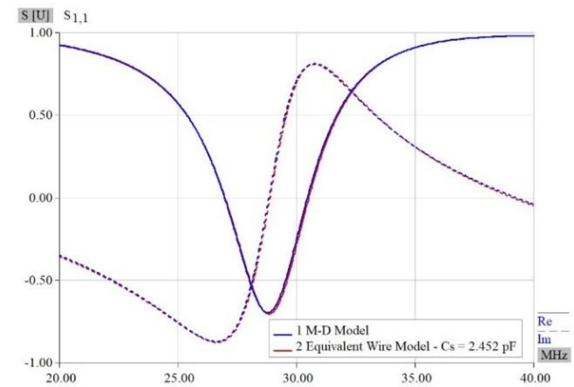

Fig. 12. Comparison of real and imaginary components of the reflection coefficient of the M-D and equivalent wire models, $C_{S,equivalent}$ = 2.452 pF.

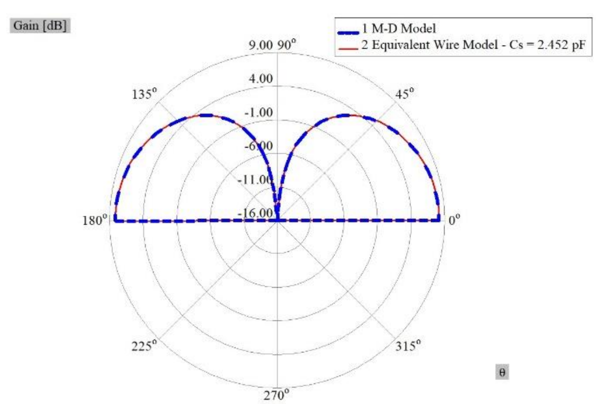

Fig. 13. Comparison of the radiation patterns of the M-D and equivalent wire models, $C_{S,equivalent}$ = 2.452 pF.

VI. SUMMARY AND CONCLUSION

The simulations of the M-D and equivalent wire models, as obtained through the provided equations, yielded very similar results. Moreover, tuning the equivalent capacitance of the equivalent wire model led to a near identical reflection coefficient and radiation pattern results when compared to those of the M-D model.

The mounting of a wire on an arbitrary platform does not require adapting the mesh of the platform, as is the case for the full 3-D M-D model. In WIPL-D Pro, the only requirement for connecting wires to a 3-D surface mesh of a platform is to properly position the wires and ensure their connectivity to the 3-D surface mesh, using non-trivial junctions at the connection points.

Moreover, the M-D model required a much larger number of unknowns, to estimate the currents over the elements, than the equivalent wire model did to approximate the axial currents along the wires. Hence, the M-D model simulation runtime far exceeded that of the equivalent wire model.

REFERENCES

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[2] WIPL-D Pro v14, Software and User’s Manual, WIPL-D d.o.o., Belgrade, 2017.

Milos M. Jovicic was born in Belgrade, Serbia, in 1992. He received his BSc. degree in Electronics and MSc. degree in Microwave Techniques from the School of Electrical Engineering, Belgrade, Serbia in 2016 and 2018. Currently, he works for WIPL-D d.o.o., Belgrade, Serbia as an Application and Technical Support Engineer. His main research activities are electromagnetics modeling and simulation of various types of antennas and microwave devices.

Saad N. Tabet received his B.S., M.S. and Ph.D. in Electrical Engineering from Mississippi State University in 1982, 1984 and 1989, respectively. He has more than 30 years of experience in analytical and computational electromagnetics (CEM). For the past 20-plus years, he has been employed at Naval Air Systems Command (NAVAIR), Naval Air Station (NAS), Patuxent River, MD, USA. He currently serves as the Senior CEM Technologist and Co-Director of the Radar and CEM Modeling (RACEMM) Laboratory, which he was instrumental in creating in 2007. He leads all CEM modeling efforts, to include in-situ antenna/array radiation patterns, coupling, cosite, and link budget analyses. His in-situ modeling and analyses efforts span a score of manned and unmanned complex fixed and rotary wing aircraft, shipboard, and other application specific structures.

Branko M. Kolundzija received his B.Sc., M.Sc. and Ph.D. in Electrical Engineering from the School of Electrical Engineering, Belgrade, Serbia in 1981, 1987 and 1990, respectively. Currently, he works as a full-time Professor at the School Electrical Engineering, Belgrade, Serbia. He is the head developer of the WIPL-D software, as well as, the author of several books and scientific papers and articles in his main areas of interest: computational electromagnetics, antennas, microwaves and EMC.