# Measurement and modeling of effective cable parameters of unshielded conductors

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# Abstract

Purpose – Cables are ubiquitous in electronic-based systems. Electromagnetic emission of cables and crosstalk between wires is an important issue in electromagnetic compatibility and is to be minimized in the design phase. To facilitate the design, models of different complexity and accuracy, for instance, circuit models or finite element (FE) simulations, are used. The purpose of this study is to compare transmission line parameters obtained by measurements and simulations.

Design/methodology/approach - Transmission line parameters were determined by means of measurements in the frequency and time domain and by FE simulations in the frequency domain and compared. Finally, a Spice simulation with lumped elements was performed.

Findings – The determination of the effective permittivity of insulated wires seems to be a key issue in comparing measurements and simulations.

Originality/value - A space decomposition technique for a guided wave on an infinite configuration with constant cross-section has been introduced, where an analytic representation in the direction of propagation is used, and the transversal fields are approximated by FEs.

Keywords Time-domain-reflectometry (TDR), Numerical simulation, Cable parameter extraction, Finite element method (FEM). Spice, Circuit analysis, S-parameter analysis

Paper type Research paper

# 1. Introduction

One main goal of the present work is to match measurement data, Spice (LTSpice, 2020) and finite element (FE) simulations to create models of different complexity. A configuration as simple as a single insulated wire above ground needs to be modeled, built and tested to

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COMPEL - The international journal for computation and mathematics in electrical and electronic engineering Vol. 41 No. 3, 2022 pp. 1041-1051 Emerald Publishing Limited 0332-1649 COMPEL extract the equivalent RLGC-parameters, characteristic impedance  $Z_0$ , velocity factor  $v_f$ , as well as delay time  $t_d$ . Naturally, simulation data and measurement data should match, but 41.3 due to imperfections of the physical test setup, such as connectors, stray capacitances, imperfect return conductors (e.g. copper ground plane), measurement inaccuracies, imperfect terminations and models for simulation that use a 0.5 mm copper plane for current return, have no connectors and stray effects involved, this is not easy to accomplish. Theory 1042and exemplary simulation data of micro-strips of a printed circuit board (PCB) can be found in Hollaus et al. (2008), Caniggia and Maradei (2008), Chaturvedi et al. (2016) as well as in Eisenstadt and Eo (1992). Careful consideration of parasitic effects is necessary, and concise analysis of their order is a prerequisite to obtaining reliable results. Because the setup here appears to be simple, it serves as a basis for more complex models which involve coupling effects between high impedance traces or low impedance loops on a PCB. Measured data and simulation, as well as underlying models, still need to match.

#### 2. Measurement setups

Each test setup requires a specific configuration and termination. Apart from the Spice simulation (LTSpice, 2020), the setups use either a polyvinyl chloride (PVC) or silicone insulated single wire above a copper ground plane (1.0 m times 1.0 m and 0.5 mm thick), which can be considered infinite. The PVC insulated wires have a copper-conductor of 1.5 mm<sup>2</sup> and red and black insulation with a diameter of 3.0 mm. The silicone insulated wires have a copper conductor of 1.0 mm<sup>2</sup> and red and black insulation with a diameter of 2.3 mm. All four wires are 1.0 m long. The wires are fixed to the ground plane with adhesive tape to minimize the distance between insulation and copper – there should ideally be no air gap between cable and ground plane. The test equipment is connected via SMA-connectors with their ground terminals soldered to the ground plane and the inner terminal to the cable under test (CUT).

# 2.1 Time domain reflectometry

The time domain reflectometry (TDR)-test uses an HP8753D 6 GHz vector network analyzer (VNA) which subsequently converts the acquired response into the time domain by means of an inverse fast Fourier transform. The calibration requires a test cable that is terminated with the system impedance of 50  $\Omega$ , an open and a short standard. The test requires the far end to be open to provide a quasi-full positive reflection of the incident signal. The TDR-test set is connected to the near end of the CUT.

### 2.2 Frequency domain testing – S-parameters

The acquisition of the S-parameters uses a Rohde & Schwarz ZVL6 as VNA. The calibration requires a full two-port through-open-short-match procedure. The measurement requires the CUT to be connected to both ports of the VNA. It represents a fully terminated CUT.

All data from the VNA consist of input reflection coefficient  $S_{11}$ , forward transmission gain  $S_{21}$ , reverse transmission gain  $S_{12}$  and output reflection coefficient  $S_{22}$  with magnitude in *dB* and the respective phases in degrees. A conversion to a linear scale and into radians [as in Tuerk *et al.* (2020)] needs to be accomplished before application of the following set of equations (1) and (2).

With  $Z_l$  being the characteristic impedance of the CUT

$$Z_{l} = \sqrt{Z_{0}^{2} \cdot \frac{(1+S_{11})^{2} - S_{21}^{2}}{(1-S_{11})^{2} - S_{21}^{2}}}$$
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and the propagation constant  $\gamma$ 

$$e^{-\gamma l} = \left(\frac{1 - S_{11}^2 + S_{21}^2}{2S_{21}} \pm K\right)^{-1}$$
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with *l* being the length of the CUT and *K* given by the following

$$K = \sqrt{\frac{\left(1 + S_{11}^2 - S_{21}^2\right)^2 - \left(2S_{11}\right)^2}{\left(2S_{21}\right)^2}} \tag{3}$$

the per unit length (PUL) parameters of the CUT can be extracted from the S-parameters.

Considering ambiguities, the  $\pm K$ -term must be accounted for in the following extraction of the RLGC-parameter set

$$\begin{aligned} R' &= Re(\gamma Z_l), \\ L' &= \frac{Im(\gamma Z_l)}{\omega}, \\ G' &= Re\left(\frac{\gamma}{Z_l}\right) \quad \text{and} \end{aligned} \tag{4}$$
$$C' &= \frac{Im\left(\frac{\gamma}{Z_l}\right)}{\omega}, \end{aligned}$$

respectively.

# 3. Simulation methods

3.1 Spice simulation

Because the CUT is a sample of 1.0 m in length and the maximum frequency is 500 MHz in the case of the S-parameter measurement, the simulation model has been split into pieces of 0.1 m for the purpose of the simulation. The PUL data acquired are divided by 10 as far as L'and R' are concerned, C' and G' are multiplied by 10. In total, the CUT in the simulation represents 1.0 m, and the condition of segments being significantly shorter than  $\lambda/10$  is still satisfied.

The model in Figure 1 shows 10 sections, each of them with 4 lumped elements (R1, L1, L1)C1, G1 to R10, L10, G10, C10), the termination of the model represented by R11 and the respective parameters which are applied to all elements. To get S-parameters from the schematic, the directive .net I (R11) V1 is used. This parameter set provides the result given in section 4 in Figure 6.

### 3.2 Finite element method

The benchmark with details is presented in Figure 2 and consists of a PVC or silicon insulated copper wire above a copper plate. The length of the wire has been selected to be

f 1 conductors



Figure 2. Insulated copper wire above ground, not to scale

**Notes:** PVC: cross section  $A_{Cu} = 1.5 \text{ mm}^2$ , diameter of insulation  $d_{ins} = 3.0 \text{ mm}$  and air gap  $d_0 = 0.1 \text{ mm}$ . Silicon:  $A_{Cu} = 1.0 \text{ mm}^2$ ,  $d_{ins} = 2.3 \text{ mm}$  and  $d_0 = 0.1 \text{ mm}$ 

 $\Gamma_{\rm air}$ 

Cu

1.0 m, and the copper plate with a thickness of 0.5 mm is assumed to have infinite dimensions. The air gap represents an estimated average in contrast to the physical setup of the measurements.

**GND** 

The basic boundary value problem to be solved in the frequency domain reads

curl 
$$\mu^{-1}$$
curl  $A + j\omega(\sigma + j\omega\varepsilon)A = 0$  in  $\Omega$ ,  
 $A \times \boldsymbol{n} = \boldsymbol{\alpha}$  on  $\Gamma_D$ 

where *A* is the magnetic vector potential (MVP),  $\mu$  the magnetic permeability,  $\sigma$  the electric conductivity,  $\varepsilon$  the electric permittivity, *j* the imaginary unit,  $\omega$  the angular frequency,  $\alpha$  some Dirichlet boundary condition,  $\Omega$  the problem domain,  $\Gamma_D$  the Dirichlet boundary and *n* the unit normal vector, see Figure 2. The domain  $\Omega$  comprises air, copper wire, insulation and GND, respectively, where  $\Gamma_D$  is its outer boundary. The following considerations for the FE simulations are based on a Cartesian coordinate system. The

transversal fields are lying in the drawing plane, compare with Figure 2. The direction of propagation corresponds to the z-coordinate. The problem for the FE simulations is assumed to be infinitely long in the direction of propagation. To take into account the specific nature of the present problem, the MVP has been decomposed into two components, a transversal component  $A_{\tau}(x, y)$  lying in the drawing plane and  $A_{z}(x, y)$ . which points into the direction of propagation

$$A = (A_{\tau}(x, y), A_{z}(x, y))^{T} e^{-\gamma z},$$
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where  $\gamma = \sqrt{j\omega\mu(\sigma + j\omega\varepsilon)}$  with  $A_{\tau,h} \in H(\text{curl}, \Omega_{2D})$  and  $A_{z,h} \in H^1(\Omega_{2D})$  for the FE approximation indicated by the additional index h, see Schöberl and Zaglmayr(2005). Because the dependence of A in equation (5) with respect to z is known, a two-dimensional FE model suffices. The lumped elements R', C', etc., have been calculated based on the corresponding losses and energies PUL in the electromagnetic field. For instance, C' has been determined with the electric energy  $W'_{e,\tau}$  due to the transverse electric field  $\mathbf{E}_{\tau}$ , which is shown in Figure 7, and R' with the aid of the losses  $P'_z$  caused by currents due to the electric field component  $\mathbf{E}_z$ . The lumped elements are obtained by

$$R' = \frac{2P'_z}{I^2},$$

$$L' = \frac{2W'_{m,z}}{I^2},$$

$$G' = \frac{2P'_\tau}{U^2} \quad \text{and}$$

$$C' = \frac{2W'_{e,\tau}}{U^2},$$
(6)

respectively. The voltage U between the copper wire and ground plate is prescribed to excite the problem; the current I is obtained by means of simulation. The z-components of the current densities in the wire and in the ground plate are opposed. The current in the copper wire should theoretically match the current in the GND plane except for the sign. The crosssections of the copper wire and of the GND-plane are completely different; consequently, their FE-meshes too. Therefore, one cannot expect an exact match of the two currents because of numerical inaccuracies. For instance, prescribing U = 1 V yields, I = -0.01411 - 0.01411 $i_{0.00013}$  A in the wire and  $I = 0.01461 + i_{0.00061}$  A in the GND-plane, which is satisfactorily accurate.

#### 4. Results

All data shown were acquired with CUTs 1.0 m in length or referred to as 1.0 m as far as simulation data are concerned.

#### 4.1 Frequency domain data

The measurement in the frequency domain was performed using a 2-port VNA (ZVL6). Both ends of the CUT are connected to the VNA, which represents a termination with 50  $\Omega$  at either end. The obtained set of S-parameters was subsequently converted into RLGC-data. The results with the given PVC-insulated conductor can be seen in Table 1.

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Measurements of a red and black silicone-insulated conductor show a similar picture, and the obtained results are given in Table 2.
It is important to note that the conductors with different colored insulators exhibit
differing RLGC parameters. This will also be addressed in Section 4.2.
The data extraction requires some arithmetic: The test point chosen is the lowest
frequency minimum of the imaginary part of $Z_{S11}$ , the impedance of the CUT derived from
$S_{11}$ , as can be seen in Figure 3 for the conductor with the black PVC insulator. The impedance seen at Port 1 of the VNA equals

	Parameter	PVC black	PVC red	Unit
	f	59.46036	57.46873	MHz
	$Z_l$	65.21131	63.88990	Ω
	R	1.325222	4.935247	Ω
	L	293.0912	285.8608	nH
	С	68.92185	70.03098	pF
	G	1068.469	704.8839	μS
	$v_p$	$2.224949 \cdot 10^8$	$2.235000 \cdot 10^8$	m/s
Table 1.	$v_f$	0.7416498	0.7450001	1
Frequency domain,	$t_d$	4.494484	4.474273	ns
PVC	${\cal E}_{r,eff}$	1.818035	1.801720	1

	Parameter	Silicone black	Silicone red	Unit
Table 2. Frequency domain, silicone	$f$ $Z_l$ $R$ $L$ $C$ $G$ $v_p$ $v_f$ $t_d$ Exects	$59.3338870.325182.415429307.7858639.87322.284874 \cdot 1080.76162474.3766091.723923$	$\begin{array}{c} 55.78086\\ 69.65358\\ 1.786379\\ 305.5429\\ 62.97747\\ 284.2796\\ 2.279667 \cdot 10^8\\ 0.7598888\\ 4.386607\\ 1.731809\end{array}$	$\begin{array}{c} \text{MHz} \\ \Omega \\ \Omega \\ \text{nH} \\ \text{pF} \\ \mu \text{S} \\ \text{m/s} \\ 1 \\ \text{ns} \\ 1 \end{array}$



**Figure 3.**  $Z_{S11}$ , real and imaginary part

$$Z_{S11} = Z_0 \cdot \frac{1 + S_{11}}{1 - S_{11}}$$
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and considers the magnitude and phase of the real and imaginary parts of  $Z_{S11}$ .

#### 4.2 Time domain data

As mentioned above, the TDR test set is used to acquire the "flight time" of a pulse incident on the CUT. The effective permittivity  $\varepsilon_{r,eff}$  can directly be derived from the time delay  $t_d = t_{flight}/2$  as given by the TDR test set using

$$\varepsilon_{r,eff} = \left(\frac{1}{v_f}\right)^2 \tag{8}$$

with  $v_f$  being the velocity factor of the wave on the CUT, which is always  $\leq 1$ , and

$$v_f = \frac{1}{c_0 \cdot t_d} \cdot l \tag{9}$$

with  $c_0$  being the speed of light in a vacuum and *l* being the length of the CUT.

Tables 3 and 4 show the parameter "unit" which has not been explained so far. TDR measurements usually normalize the amplitude of the reflected wave to the system impedance. A full reflection from an open terminated cable is then represented by a unit of 2, whereas no reflection at the end of a cable, i.e. a proper termination with the impedance of the conductor, will show a unit of 1.

A fraction or a multiple of a unit is consequently a measure of the impedance of the cable, as outlined in Figure 4.

The "flight time" mentioned earlier is the time a pulse incident into the CUT takes until it returns from the open end of the CUT. The pulse "travels" two times the cable

Parameter	PVC black	PVC red	Unit	
Flight time	9.732	9.568	ns	
$t_d$	4.866	4.784	ns	
v <sub>f</sub>	0.6855	0.6973	1	
$\mathcal{E}_{roff}$	2.128	2.057	1	
Υ <sub>f</sub>	0.145	0.136	1	Table 3.
$\dot{Z}_l$	66.95	65.74	Ω	Time domain, PVC

Parameter	Silicone black	Silicone red	Unit	
Flight time	9.78	9.624	ns	
$t_d$	4.89	4.812	ns	
v <sub>f</sub>	0.6821	0.6927	1	
$\mathcal{E}_{roff}$	2.149	2.0839	1	Table 4.
ν <sub>f</sub>	0.189	0.187	1	Time domain
$Z_l$	73.30	73.00	Ω	silicone

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length; therefore, the delay time  $t_d$  is half the "flight time". The "reflection factor" mentioned in Figure 4 can be converted into an impedance value of the CUT by means of the following equation

$$Z_l = Z_0 \cdot \frac{1 + r_f}{1 - r_f},$$
(10)

where  $Z_l$  represents the characteristic impedance of the CUT and  $r_f$  the reflection factor with a sign.

A representative measurement is shown in Figure 5.

# 4.3 Spice results

To verify the applicability of the model data and their match to measured results, the acquired cable data of "PVC black" have been used. It is clear that the idealized simulation because no parasitics are considered, shows minor deviations. Yet, a first verification with the resonant frequency shows a satisfying match, as given in Figure 6.

The trace shows  $Z_{S11}$  derived from  $S_{11}$  using equation (7) but a blank and the resonance at 66.14 *MHz*, which is a close match to the measurement results.



**Figure 5.** Exemplary TDRpattern



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Figure 6. Resonance match for PVC black

# 4.4 Finite element simulations

Homogenous Dirichlet boundary conditions  $\mathbf{A}_{\tau} \times \mathbf{n} = \mathbf{0}$  and  $A_z = 0$  are selected on  $\Gamma_D$ . The shortest line from the copper wire to GND consists of  $\Gamma_{iso}$  and  $\Gamma_{air}$ , compared with Figure 2. A voltage of U = 1 V as excitation is prescribed by the tangential component of  $\mathbf{A}_{\tau}$  along  $\Gamma_{iso}$  and  $\Gamma_{air}$  corresponding to  $\epsilon_r$  of air and insulator and the lengths of  $\Gamma_{iso}$  and  $\Gamma_{air}$ , respectively.

Simulations have been carried out with Netgen/NGSolve (Schöberl, 2021). The transversal electric field is obtained by

$$E_{\tau} = -j\omega A_{\tau},\tag{11}$$

a field plot is presented in Figure 7. The measurement and simulation results are summarized in Table 5. No losses have been considered in the insulation; G' solely takes



 $\begin{array}{c} \mbox{Figure 7.} \\ \mbox{Transverse electric} \\ \mbox{field } | \mathbf{E}_{\tau} | \mbox{ for the PVC} \\ \mbox{insulated wire, detail.} \\ \mbox{The FE model} \\ \mbox{consists of 13,160} \\ \mbox{elements} \end{array}$ 

		Measu	urement	Simu	ilation	
Parameter	Unit	PVC	Silicone	PVC	Silicone	
f	MHz	59.3	59.5	59.3	59.5	
R'	Ω/m	1.33	2.42	1.19	1.45	
Ľ	nH/m	293	308	306	293	<b>T</b> 11 <b>T</b>
C	pF/m	68.9	62.2	81.8	72.9	Table 5.
$G^{'}$	mS/m	1.07	0.64	0.12	0.14	Measurement and
$Z_l$	Ω	65.2	70.3	61.1	63.4	simulation results

COMPEL<br/>41,3into account losses due to the transverse field in copper, which explains the large<br/>difference between measurement and simulation. The true values of several parameters<br/>are not known. To study the influence on transmission line parameters (TLPs), several<br/>simulations have been carried out. The uncertain parameters have been varied in a<br/>meaningful and feasible range. Results are presented in Tables 6–8. The influence of  $\varepsilon_r$ <br/>on the TLPs is big, whereas that of  $d_0$  and  $d_{ins}$  is rather moderate. However, it is worth<br/>mention that for a very small  $d_0$  and a feasible FE mesh, the limit value C' = 93.5 pF has<br/>been found.

#### 5. Conclusions

Measurements in the frequency domain and in the time domain show a reasonably accurate match of the results.

It needs to be said that both approaches complement and confirm each other. The FD testing requires a terminated line (with the system impedance), the TD testing needs an open line end, i.e. without termination matching the system impedance.

Future work will extend the model to extract estimated emissions from a single conductor as well as bus structures at frequencies of interest. This will be examined with and without a grounded plane with subsequent verification and matching to get reliable simulation models. Three-dimensional FE models will be used to appropriately simulate relevant emissions. The simulation models shall be optimized for fast and reliable

	$\frac{\varepsilon_r}{1}$	$R^{'}$ $\Omega$	L <sup>′</sup> nH/m	C' pF/m	G' mS/m	$Z_l$ $\Omega$
<b>Table 6.</b> Relative permittivityof the insulator	3	1.15	302	71.0	0.11	65.2
	4	1.19	306	81.8	0.12	61.1
	5	1.22	309	90.5	0.12	58.4

	$d_0$ mm	$\stackrel{R^{'}}{\Omega}$	L΄ nH/m	C' pF/m	G' mS/m	$Z_l$ $\Omega$
Table 7	0.08	1.21	304	84.4	0.13	60.1
Air gap width	0.1 0.12	1.19 1.17	306 307	81.8 79.5	0.12 0.11	61.1 62.2

	d <sub>ins</sub> mm	$\stackrel{R^{'}}{\Omega}$	L <sup>′</sup> nH/m	C' pF/m	G <sup>′</sup> mS/m	$Z_l$ $\Omega$
Table 8.	2.9	1.21	298	82.7	0.12	60.1
Diameter of the	3.0	1.19	306	81.8	0.12	61.1
insulator	3.1	1.17	313	81.0	0.11	62.2

simulations based on a sensitivity analysis showing the impact of parameter variations like  $d_0$  in Figure 2.

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