Survey on the Generation of Equivalent Circuit Cable Models for Transient Simulation

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Abstract—Especially in automotive and aerospace applications very long and complex cable harnesses are part of virtually all electronic systems. To precisely predict their electromagnetic behavior by simulation, accurate models of cables and harnesses are essential. In this work a semi-automatic workflow to derive a lumped circuit transmission line (LCTL) model from the 2D cross section of an arbitrary cable harness is presented. In contrast to conventional measurement based behavioral models (e.g. S-parameters) this approach provides flexibility to depict different harness compositions without the need for repeating measurements. The model generation process and involved limitations are discussed in detail. For validation, the simulation results obtained with the LCTL model are compared to measurements. 3D finite element method (FEM) and 3D hybrid multi transmission line (MTL) simulation in frequency and time domain. The considered frequency range is up to 1 GHz.

Keywords—cable harness; equivalent circuit; 3D EM simulation; finite element method (FEM); lumped circuit transmission line (LCTL); multi transmission line (MTL); per-unit-length parameters (p.u.l.)

I. INTRODUCTION

Cable harnesses are an intrinsic part of modern electronic systems. Especially in vehicles and aerospace applications (e.g. more electric aircraft (MEA)) long and complex harnesses are prevailing, but also for wired data transmissions the electrical performance of transmission lines is essential. For the prediction of suchlike system’s performance by simulation, the modeling of cable harnesses is an inevitable task - even if the device being designed is e.g. an integrated circuit (IC) and the attached harness connects a load. The generation of accurate and broadband electrical harness models is the first necessary step for system level simulation. Nonetheless, its complexity is often underestimated.

Generally speaking, due to the flexible nature of cables, measurements of their transmission properties are hardly repeatable. Stochastic approaches describing cross talk in terms of mean value and standard deviation are common, e.g. [1]–[6], where it is shown that measurement results may easily deviate by 10 dB depending on the (unknown) position of a wire within a harness. In simulation one needs to define a specific harness cross section and wire routing. This can never cover all possible variations of a real setup, hence the goal of cable harness simulation is not to precisely depict a single measurement result, but rather to predict a typical mean value. Behavioral data to describe the influence of transmission lines (TL) on electric signals is usually obtained by measurement in frequency domain (FD) using vector network analyzers (VNA). The resulting scattering parameter matrices (S-parameter) yield non-transparent black-box models when used for circuit simulations, which are known to frequently cause issues during time domain (TD) transient simulation [7]. Furthermore, S-parameter models cannot be transformed to depict a different harness composition than that of the original measurement. To increase stability of transient simulations, acquire more flexible harness models and to overcome the need for repetitive measurements, the use of lumped circuit transmission line (LCTL) models is proposed in this work.

Commercial EM solver tools to automatically generate LCTL models from multi-conductor transmission lines (MTL) exist. In this paper it is investigated if the same can be accomplished solely from the harness’ 2D cross section. The used cables are twisted wire pairs (TWP). Thus, the 3D information on the wire’s twist will be lost. The circuit topology and necessary parameter extraction from the cross section are already analytically described in [8], [9]. Often analytical examples tend to drastic simplifications on 2D cross section and material to ease calculations. With todays computational resources such are not necessary anymore. Per-unit-length (p.u.l.) parameters, i.e. RLCG/m, can be determined precisely from the detailed 2D cross section. Actually, a number of commercial simulation tools offer the functionality to generate (behavioral) TL models from 2D cross sections. This paper may be understood as a study on the usability and limitations for day-to-day cable modeling. Generating broadband models manually is complicated, error-prone and time consuming work. Hence, we present a workflow to semi-automatically generate a SPICE netlist from the 2D cross section of an arbitrary harness. The approach is motivated as follows:

First, p.u.l. parameter extraction in 2D is computationally faster than accomplishing the same from full 3D setups. However, the question remains if 3D parameters like cable routing, twisting and discontinuities at the cable ends need to be considered.

Second, the less RLCG components the LCTL netlist is made from, the faster is the circuit simulation. A minimalistic consideration of frequency dependent phenomena like skin or proximity effect will result in a simple netlist and reduce circuit
simulation times (however, also adversely affect the models effective bandwidth).

The paper is structured as follows: Section II gives an overview of state-of-the-art approaches to generate models of cable harnesses for transient simulation and clarifies the motivation for the proposed approach. In Section III a methodology to semi-automatically generate a minimalistic LCTL circuit from only the 2D cross section of a cable harness is presented. Limitations in terms of accuracy and frequency bandwidth are discussed. In Section IV the method is validated against time and frequency domain measurements in comparison to results obtained with commercial 3D electromagnetic (EM) simulation tools applying the finite element method (FEM) or hybrid multi transmission line (MTL) solvers. As example serves a harness consisting of three parallel twisted wire pairs (TWP). Section V concludes the work and summarizes known issues.

II. COMMON APPROACHES FOR THE GENERATION OF TRANSMISSION LINE MODELS

The authors distinguish between the following approaches to include models of cable harnesses into circuit simulation:

- **S-parameters (FD):** Non-physical behavioral descriptions by S-parameters can be obtained from measurement or simulation. Because these network parameters are defined in FD, transient TD simulation require to generate an equivalent circuit description fitted to the FD data in a pre-processing step [7]. This is usually computed by the circuit simulation software in a background process, without notice of the user. However, the performance of the fitted circuit depends on the capability of the used engine (especially to extrapolate a correct DC operating point), thus it might vary between circuit simulators (or even between versions of the same simulator). That means, the same set of S-parameters can lead to very different transient simulation results. As a consequence, many circuit designers prefer models that have real physical meaning.

- **Non-physical equivalent circuit model (TD+FD):** Instead of using the S-parameter raw data for simulation, above described equivalent models can be directly used as input to circuit simulation, i.e. as netlist, to overcome the issue of fitting-dependent simulation results. Most common is the use of so-called broadband SPICE (bbspice) models. Such models are composed of a set of ideal voltage and current sources which reflect the poles and zeros of the complex-valued frequency data. One example for the specific netlist generation process is for instance given in [10]. Actually, this is again a behavioral model, hence circuit simulation computes very fast. Furthermore, the model is well suited for transient simulation. However, suchlike equivalent circuit models might behave non-physically and could also cause convergence issues, depending on the quality of data that was used to generate the model.

- **Expression-based p.u.l. or RLCG/m models (FD):** Instead of using S-parameters themselves for the description of transmission lines, they can be expressed as frequency dependent per-unit-length (p.u.l.) parameters (or more generally transmission line (TL) parameters), i.e. RLCG/m(f). The works [11], [12] show how to convert S-parameters to RLCG/m(f). These can be used as input to e.g. the W-element of HSPICE or the similar mtline object of Cadence Spectre (to name only two examples). Respective circuit blocks use again rational fitting algorithms to generate a behavioral model from the RLCG/m(f) input data. As above, the simulation result depends on the quality of the source data and on the specific internal implementation. Alternatively, p.u.l. parameters can also be obtained by analytical expressions, e.g. [13], [14] for specific cable types. These expressions may be implemented as functions of frequency in the circuit simulator. However, due to the used modeling language, e.g. Verilog-A, the model will not be universally usable with any circuit simulation software.

**Physical equivalent circuit model (TD+FD):** The last approach mentioned here is to generate a physically meaningful equivalent circuit which consists of 'real' circuit components only, i.e. resistors, inductors, capacitors and conductors - in short RLCG. It is fairly obvious that a single set of RLCG cannot reflect broadband frequency dependency, hence the length of the harness is divided into lumped RLCG blocks derived from the p.u.l. values (as sketched in the center right schematic of Fig. 1). The component values correspond to the respective length of the transmission line. The result is a so-called lumped circuit transmission line (LCTL) model. Commercial software tools are able to automatically generate LCTL models from the 2D cross section of a cable harness while also taking 3D properties into account [15], [16]. The clear advantage of suchlike physically meaningful circuits is, that they are usable with any circuit simulator, sure to cause no convergence problems, behave well at DC point and the corresponding netlist is human interpretable. It is even possible to describe frequency dependency of p.u.l. parameters RLCG/m(f) with discrete RLC components, like e.g. the skin effect as exemplified in [17]. This, however, increases the complexity of the model substantially which slows down the simulation.

III. METHODOLOGY

The used methodology is depicted by Fig. 1. It was applied before in [18]. Per-unit-length RLCG/m parameters of the transmission line cross section (sketched in the top left of the figure) are extracted from quasi-static 2D simulation (ANSYS 2D Extractor [19]) and exported to MATLAB. From the given coupling matrices a script automatically generates a lumped elements SPICE netlist of N cascaded RLCG blocks, whereas each of the blocks has to be electrically short. According [20] (Chapter 5.1.2) the maximal length of transmission line that could be represented by a single lumped equivalent circuit is limited to $\Delta l \leq \lambda/10$, where $\Delta l$ is the length of one transmission line segment and $\lambda$ the wavelength related to the maximum frequency of interest and the isolation material. The wavelength in a transmission line with dielectric material between the conductors is described in [21] (Chapter 4.1) as

$$
\lambda = \frac{\lambda_0}{\sqrt{\varepsilon_r}} \quad \text{and} \quad \lambda_0 = \frac{c_0}{f}
$$

(1)

where $\varepsilon_r$ is the relative permeability of the isolation material, $c_0$ the speed of light and $f$ the frequency. A long transmission line is represented by cascading RLCG blocks of length $\Delta l$ until the desired overall length $l$ is obtained. To calculate the necessary number $N$ of segments, the aforementioned equations are put together to

$$
N \geq \frac{10 \cdot f_{\max} \cdot l \cdot \sqrt{\varepsilon_r}}{c_0}
$$

(2)
Since the used dielectric materials have $\varepsilon_r \leq 4$, (2) can be further simplified to:

$$N \geq \frac{20 \cdot f_{max} \cdot l}{c_0}$$  \hspace{1cm} (3)

### A. Work-Flow to Obtain Cable Parameters

Besides the real cable routing inside the harness, other typically unknown variables during simulation are the actual dielectric properties of the cable’s insulation, i.e. the (effective) relative permittivity $\varepsilon_r$ and the dielectric loss tangent $\tan(\delta)$. Furthermore, the 2D cross section can only be approximated by sample measurement or data sheet information. The workflow that is described below is based on measurements and can be used to obtain the correct insulation material parameters for the simulations shown in this work:

1) S-parameter measurements on a short piece of (twisted) wire. By rule of thumb, the used wire needs to be shorter than $\lambda/4$ of the maximum considered frequency to ensure that no reflections due to unmatched line termination invalidate the measurement.

2) Extraction of the dielectric material parameters from the measured S-parameters similar to [11], [12].

3) Obtained material parameters are applied to a nominal 2D cross section. The exact 2D geometry is fitted by utilizing an optimization task with the goal to decrease the deviation between simulation and reference result. For this work, the software ANSYS 2D Extractor [19] was used.

4) Verification of the material properties by measurement and 3D simulation (this work used EMCoS Hybrid MTL solver [22]) with the actual wire length and twisting.

For the automotive FLRY-A $0.5\text{mm}^2$ TWP mostly used in this work, the described procedure resulted in $\varepsilon_r = 3.354$ and $\delta_{loss} = 0.0178$, in contrast to the expected values for PVC of $\varepsilon_r = 2.7$ and $\delta_{loss} = 0.007$.

### B. Discussion of Limitations

The methodology described by Fig. 1 involves two drastic simplifications: First, the cross section is assumed to be constant over the whole length of the harness. Second, the RLCG/m parameters are extracted at a single frequency point, although it is well known that they are frequency dependent. This is mainly due to the dielectric properties of the cable insulation material, namely the dielectric loss tangent $\tan(\delta)$ and skin effect. The issue is visualized by Fig. 2 for two materials with different values of $\tan(\delta)$. The coupling capacitance $C_{12}$ and conductance $G_{12}$ between wires are influenced by $\tan(\delta)$, but the deviation of $C_{12}$ over frequency is negligible. Due to the vast frequency dependency of the conductance a LCTL model based on the extraction at a certain frequency cannot be broadband. For above described method of extracting RLCG/m from the harness cross section a solution frequency needs to be chosen. The problem is the p.u.l. parameters need to be exact at the DC point if the resulting LCTL model should be usable for transient simulations. Unfortunately, a low extraction frequency will yield incorrect simulation results at high frequencies.

Besides the change of inter-wire conductance $G_{12}$, Fig. 2 also shows the increase of series resistance $R_s$ within $500\text{MHz}$ which can be explained by the skin effect. Note that it is no function of the insulation material. The impact of the deviation of $R_s$ from $0\Omega$ to $2.5\Omega$ is negligible compared to the impact of increasing $G_{12}$ from $0\text{nS}$ to $2\text{nS}$, i.e. the isolation resistance from infinite to only $500\Omega$.

For the materials values used in Fig. 2, Fig. 3 shows simulations of the forward transmission S-parameter $S_{21}$ when respective dielectric properties are applied to a harness made from three similar TWPs. The dielectric loss leads to deviations increasing with frequency. The results shown in Fig. 3 were obtained with EMCoS Hybrid MTL solver [22] which is able to generate LCTL models reflecting frequency dependencies. With the proposed workflow this is not yet possible. Measurements on two types of TWP cables were conducted to investigate the consequence by experiment. Fig. 4 gives $S_{21}$ of an $1\text{m}$ Ethernet cable in comparison to an automotive FLRY-A $0.5\text{mm}^2$ type. It is visible, that the attenuation of the latter increases rapidly with frequency, while that of the Ethernet cable stays constant. The figure also shows the simulation results obtained with the proposed minimalistic LCTL model which neglects the frequency dependency of the insulation material. The conclusion is, that the presented modeling approach can work very well, as long as the considered cable type is frequency stable.

### IV. Verification

The proposed methodology is verified with an experimental setup of three FLRY-A $0.5\text{mm}^2$ TWPs above a large ground plane and in between metallic fixtures as illustrated by Fig. 5.

#### A. Frequency domain (FD): S-parameters

Results from S-parameter simulation and measurement are compared in Fig. 6. Two measurement results of forward transmission factors $S_{21}$ and $S_{65}$ are exemplified to display the deviation of measurements. One could be tempted to simplify the setup as perfectly symmetric with six equal forward
Fig. 2. $S_{21}$ of single TWP above ground plane: Frequency dependency of p.u.l. series inductance $L_{11}$, coupling capacitance $C_{12}$ and conductance $G_{12}$ between wires and series resistance $R_s$ as function of dielectric loss tangent $\tan(\delta)$ of the insulation material. Data extracted from frequency sweep based on a single 2D cross section, using ANSYS 2D Extractor [19].

![Graph showing $S_{21}$ vs. Frequency](image)

Fig. 3. $S_{21}$ of 1 m bundle of three TWP in between metallic fixtures and above a large ground plane (setup of Fig. 5): Impact on insulation material properties $\varepsilon_r$ and $\tan(\delta)$ on simulation result. Simulations conducted with EMCoS Hybrid MTL solver [22].

![Graph showing $S_{21}$ vs. Frequency for Ethernet and FLRY-A](image)

Fig. 4. Comparison of two types of TWP: Ethernet cable versus automotive FLRY-A. Shown is the forward transmission $S_{21}$ of 1 m TWP, where one of the wires is the GND reference. The Ethernet cable is more frequency stable. The idealized LCTL model neglects frequency dependent dielectric loss, hence it does not reflect the differences between the cables.

![Graph showing comparison of Ethernet and FLRY-A](image)

Fig. 5. Measurement and simulation setup for bundle of 3 TWP above ground plane [18]. During FD simulation, the fixture is respected by either including it to the 3D model or with an equivalent circuit. In TD simulation it has no noteworthy impact due to the limited bandwidth.
transmission factors but in reality, none of it equals another. The green trace ‘LCTL netlist simulation’ gives the result obtained with the presented approach. Additional simulations were conducted with the commercial 3D EM software tools ANSYS HFSS (FEM) and EMCoS (Hybrid MTL). It can be observed, that all of the simulation results are within ±5 dB of the measurement result for frequencies up to 700 MHz. An extension of the frequency range is possible with higher model generation effort but was outside the scope of this work. Note that using FEM, simulation time is in the range of hours and significantly affected by the model’s level of detail, hence the solution’s accuracy.

B. Time Domain (TD): Transient Pulses

To validate the usability and correctness of the generated netlist for transient simulation in time domain, an experiment with different terminations at the 12 ports of the three TWP harness as sketched in Fig. 7 was set up. Fig. 5(c) depicts the physical setup where ports 1-4 belong to TWP1, ports 5-8 to TWP2 and ports 9-12 to TWP3. Fig. 5(d) shows the measurement of port voltages with standard 500 MHz oscilloscope probes. In simulation their influence on the measured signal is modeled by the equivalent circuit of Fig. 8. At ports 1,3 and 5 a broadband (> 1 GHz bandwidth) pulse of 2 ns and 2.5 V with 5 MHz repetition rate was injected. The resulting pulse form is shown in the bottom subplot of Fig. 9. The fast rise time of the pulse guarantees well observable distortions at all other ports of the harness.

- Injection to TWP1 at ports P1 and P3 mimics common-mode (CM) disturbance into symmetric termination (2x 50 Ω at P2 and P4). This represents e.g. the idle-stated outputs of a symmetric line driver.
- One wire of TWP2 is grounded, while the other is terminated by 50 Ω. This condition represents e.g. a single-ended driver topology.
- TWP3 is on both sides terminated with different ohmic loads. Here common-mode to differential-mode conversion by coupling between cables is investigated. The voltages induced to P10 and P12 are higher than these at P9 and P11 due to the lower resistance values. (For brevity, only the voltages at P10 and P12 are plotted in Fig. 9.)

Simulations were conducted with the free SPICE simulation software LTspice [23] using either the SPICE netlist generated by the methodology described in Section III or the equivalent circuit output of the Hybrid MTL task processed with EMCoS. The comparison in Fig. 9 shows generally good agreement between measurement and both modeling approaches in both voltage amplitudes and time delay. Note that the measurement’s bandwidth was limited by the use of standard 500 MHz probes. Hence, the big deviations at higher frequencies observable from Fig. 6 have no impact on the TD results. A repetition of this experiment with higher bandwidth probes is planned for future investigations.

V. CONCLUSION

In this work, different approaches to obtain (equivalent circuit) models of cable harnesses were discussed with respect to their usability for transient circuit simulations. The advantage of physically meaningful LCTL models is, that they are reliably stable and usable with any circuit solver. As a matter of fact, the parameters of real harnesses are typically poorly defined or extremely variable, especially concerning the dielectric properties of the cable’s insulation and the cross section of the harness composition over its full length. It was shown that the main parameter leading to frequency dependent p.u.l. parameters RLCG/m(f) is the dielectric permittivity εr(f) of the insulation material. In the considered frequency range, slight variations in cross section or skin effect had a minor impact. Due to the frequency dependency it is generally not possible to generate a simple but nonetheless accurate and broadband TL model based on 2D extraction of the RLCG/m coupling matrix at only one single frequency point. However, even though frequency dependency of the isolation material
was not taken into account for the proposed simplified LCTL modeling approach, simulation results of similar accuracy as those obtained from 3D MTL or FEM simulation could be achieved in case of a harness of multiple TWP. That is because the spread of measurement results from a realistic harness is even bigger than the variation introduced by neglecting material properties in simulation.

REFERENCES


