A Behavioral Model of High-Voltage Traction Inverters for EMC Analysis

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***Abstract*— The electromagnetic interference delivered by high voltage power inverters driving traction electric motors in a real application can differ from those observed in the qualification tests, significantly. Wanting to address this issue in advance researchers and practitioners performs both circuit and electromagnetic simulations, but to do that a model of the source of switching noise, the power inverter in our case, is need. The behavioral models available in literature are based on the series and shunt impedance insertion method, which is not sufficiently accurate. Aiming to address this issue, this paper proposes a new approach to derive the behavioral model unknowns from a set of measurement results. The proposed solution was applied to an automotive high-voltage three-phase power inverter, and its effectiveness was validated comparing the switching currents resulted from the model prediction with those obtained from measurements.**

Keywords—Conducted emission, power inverter, behavioral modeling; switching noise, electromagnetic interference.

I. INTRODUCTION

Power inverters of electric vehicle (EV) or hybrid electric vehicle (HEV) driving traction motors are the primary source of switching noise, which is mostly propagated through the cables connecting the inverter to the electric motor (i_{SU}, i_{SV}, i_{SW}) and to the battery pack (i_{SP}, i_{SN}), as show in Fig. 1. The level of such currents can be high enough and spread over a wide frequency range to impair the operation of other on board electronic equipment, such as those needed for the navigation of the vehicle or for communication purposes. Wanting to mitigate such disturbances and to comply with the tight emission limits required by national and international standards [1], the input and output of the power inverter are heavily filtered, the inverter itself is enclosed in a metal case and shielded cables are used to connect it to the motor and to the battery pack. Such electromagnetic barriers allow the inverter to comply with the emission limits required at the module level [1]. However, the noise currents delivered by the power module in a real application, i.e., in a vehicle, can differ from those measured in test bench prescribed for its EMC qualification, significantly. As a consequence, the power module can still affect the operation of other electronic systems nearby and the compliance to system level specifications is not guaranteed. Wanting to perform system level simulations, several authors have proposed circuit models that allow one to predict the electromagnetic emissions delivered by power modules under different loading and operating conditions. Looking at the open literature, there can be found analytical models, i.e., equivalent circuits based on the real one that besides the models of the real components and the power transistors driving circuits comprise the wiring parasitic inductances and capacitances, as well as the parasitic coupling with the inverter metal case [2]. Such a kind of models lend themselves well to optimize the design of the EMI barriers, i.e., EMI filters and shields, but given their high complexity, they are time consuming, thus not suitable to perform analysis at the system level. In addition, OEMs disclose the information needed to build up such models unlikely, therefore the end users cannot build up such models by themselves.

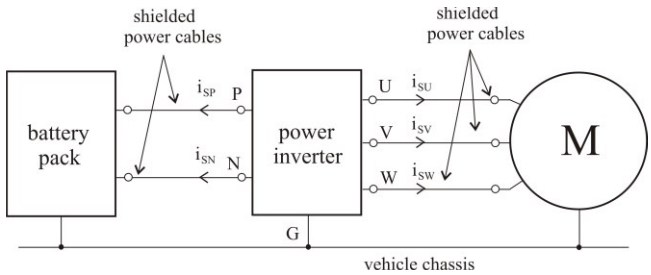

Fig. 1. Block diagram of traction power system comprising a battery pack, a three phase inverter and an electric motor. The noise currents generated by the power inverter at the input ports (i_{SP}, i_{SN}) and at the output ports (i_{SU}, i_{SV}, i_{SW}) are highlighted.

Wanting to overcome these limitations, some authors developed behavioral models, know also as black box models, which circuit topology usually refer to the extended Thevenin or Norton equivalent circuit. The unknows of such models are usually obtained from the measurement of voltages and/or currents at the device terminals as discussed in [3-6]. As far as the dc power supply terminals of the module (P, N, G in Fig. 1) are concerned, such methods require the direct or indirect measurement of switching noise currents flowing through the power supply cables for different loading conditions [3]. In practice, the power circuit to be modeled should be supplied by a noiseless dc source, which internal impedance is known and that should be changed during the tests. The currents delivered by the power module under a given number of loading conditions should be acquired thus used to obtain the unknows of the EMI model. The principle at the basis of such methods can be better explained referring to a two-terminal noise source like that shown in Fig. 2. In this case the purpose is to obtain the model unknowns V_S, Z_S. To do that, the two-terminal module is

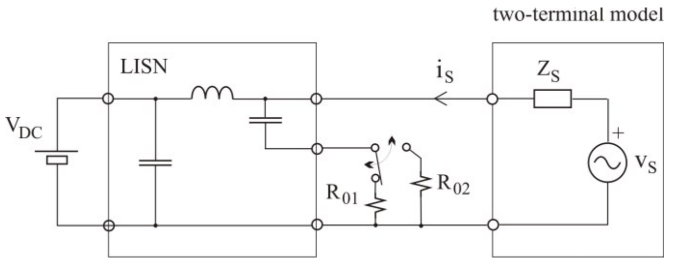

Fig. 2. Test setup used to obtain the model unknowns of a two-terminal power module.

powered by the dc source V_{DC} through a Line Impedance Stabilization Network (LISN) like that prescribed in [1]. It allows for changing the source impedance in the frequency band of interest. Therefore, the noise current flowing in the power supply line can be measured with the LISN output port loaded by R_{01} first, and by R_{02} afterwards, obtaining i_{S1} and i_{S2}, respectively. Then, the model unknows can be derived from the circuit inspection obtaining

$$v_S = (R_{01}i_{S1} - R_{02}i_{S2})\frac{i_{S1}}{i_{S2} - i_{S1}} + R_{01}i_{S1},$$

$$Z_S = \left(R_{01}\frac{i_{S1}}{i_{S2}} - R_{02}\right)\frac{i_{S2}}{i_{S2} - i_{S1}}.$$

Such an approach, which belongs to the Series and Shunt Impedance Insertion (SSII) modeling methods, is valid regardless of the number of terminals but it is worth noticing that the number of measurements needed to derive the model unknows increases significantly with the number of terminals [3]. Besides that, the methods based on SSII are not sufficiently accurate because the source of switching noise is not cyclical and the currents (i_{S1}, i_{S2}) are acquired in subsequent not triggered time frames.

This paper proposes a method to extract a behavioral model of switching power circuits that separates the measurement of the model impedances from that of the equivalent sources. The model is suitable to be used in EMC analysis carried out with circuit or electromagnetic simulators.

The paper is organized as follows. Section II shows the behavioral model developed in this work along with the method to obtain the model unknowns. Section III provides a description of the test benches used to measure the power supply noise currents generated by a three phase inverter driving a motor. In the same section the validation of the proposed approach is presented and some concluding remarks are drawn in Section IV.

II. Proposed Modeling Approach

This section provides a method to derive a behavioral model of a switching circuit like a power inverter driving an electric motor. The circuit topology is that resulting from the Thevenin's Theorem for a multi port network, whereas the model unknows are derived in two steps. The impedances, which parameters are time invariant, are obtained from a set of measurements carried out with the power module not switching, thus the current flowing through the power supply cables are measured in the time domain.

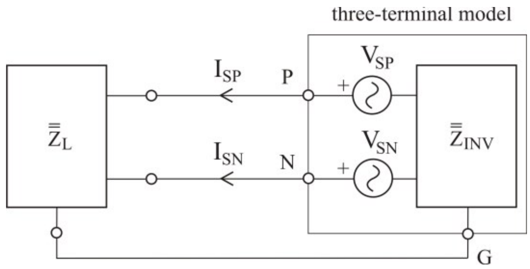

Fig. 3. Equivalent circuit comprising the that of the power inverter at the input port and that of the dc power supply $\bar{\bar{Z}}_L$.

A. Topology of the proposed EMI model

The switching currents delivered by a power inverter at the dc input lines can be modeled referring to the three-terminal Thevenin equivalent circuit shown in Fig. 3. It comprises of a passive two-port modeled by its 2x2 impedance matrix $\bar{\bar{Z}}_{INV}$ connected to the P and N equivalent voltage sources, V_{SP}, V_{SN}. The correctness of using linear time-invariant models for representing time varying power converters was proved in [7]. Furthermore, the circuit in Fig. 3 shows the two port network describing of the dc power supply network, which is modeled by the 2x2 impedance matrix $\bar{\bar{Z}}_L$.

B. Extraction of the model unknowns

The unknowns of the proposed models, i.e., the elements of $\bar{\bar{Z}}_{INV}$ and $\bar{V}_S = [V_{SP}, V_{SN}]$ can be derived from two distinct set of measurements. About $\bar{\bar{Z}}_{INV}$ it is obtained from the impedance matrices measured at the ports PG and NG of the circuit shown in Fig. 3, with the inverter connected and disconnected from the dc power supply network, obtaining $\bar{\bar{Z}}_T$ and $\bar{\bar{Z}}_L$, respectively. Based on that, it results

$$\bar{\bar{Y}}_{INV} = \bar{\bar{Y}}_T - \bar{\bar{Y}}_L \quad (1)$$

where $\bar{\bar{Y}}_T = \bar{\bar{Z}}_T^{-1}$, $\bar{\bar{Y}}_L = \bar{\bar{Z}}_L^{-1}$, $\bar{\bar{Y}}_{INV} = \bar{\bar{Z}}_{INV}^{-1}$.

Regarding $\bar{V}_S = [V_{SP}, V_{SN}]$, its elements are obtained from the simultaneous measurement of the noise currents flowing through the power supply cables, i.e., i_{SP}, i_{SN} reported in Fig. 3, during the normal operation of the power inverter.

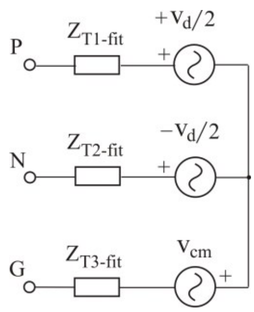

Fig. 4. T equivalent circuit of the power inverter at the dc input port.

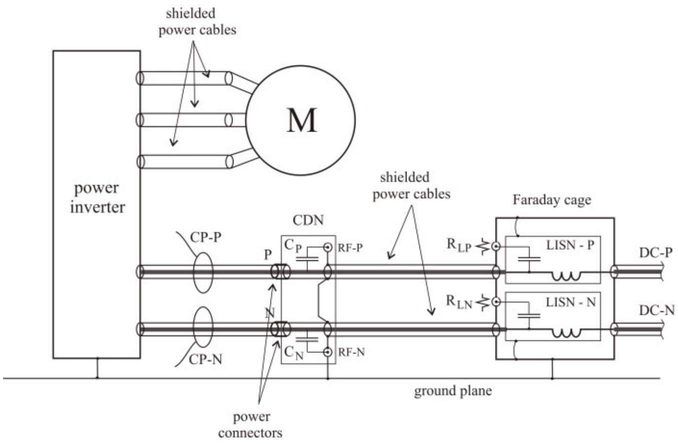

Fig. 5. Test bench assembled to perform the measure scattering parameter matrix needed to obtain $\bar{\bar{Z}}_L$, $\bar{\bar{Z}}_{INV}$ as well as the equivalent voltage sources V_{SP}, V_{SN}.

Based on that, the voltage source vector can be evaluated as

$$\bar{V}_S = \left(\bar{\bar{Z}}_{INV} + \bar{\bar{Z}}_L\right) \cdot \bar{I}_S \qquad (2)$$

where $\bar{I}_S = [I_{SP}, I_{SN}]$ is the Fast Fourier Transform (FFT) of the current vector $\bar{\imath}_S = [i_{SP}, i_{SN}]$.

A T equivalent circuit of the proposed model is given Fig. 4 where Z_{Ti-fit} are obtained by approximating

$$Z_{T1} = Z_{INV-11} - Z_{INV-12}$$
$$Z_{T2} = Z_{INV-22} - Z_{INV-21}$$
$$Z_{T3} = Z_{INV-12}$$

with the vector fitting approach [8], which is available in Matlab. About the equivalent voltage sources, they can be expressed in terms of common mode (V_{CM}) and differential mode (V_D) voltages as

$$V_{CM} = \frac{V_{SP} + V_{SN}}{2}$$

and

$$V_D = V_{SP} - V_{SN},$$

and the time domain voltage sources v_{CM} and v_D are derived applying the inverse FFT algorithm.

III. Test Bench, Measurements and Validation

The above presented model was extracted for a high voltage power inverter driving a three-phase Permanent Magnet Synchronous Motor, with the test setup sketched in Fig. 5. A picture of it is show in Fig. 6. Besides the power inverter and the electric motor, it comprised of two high voltage LISNs enclosed in a Faraday cage, that prescribed by [1] for test setups with shielded cables, and an ad-hoc coupling decoupling network (CDN) specifically designed within this project to perform the impedance measurements. The inverter was connected to a 340V dc power supply (not shown in Fig. 5) by means of shielded power cables through the two HV LISNs and the CDN. The impedances belonging to $\bar{\bar{Z}}_L$ were obtained from the measurement of the scattering parameters at the port RF-P and

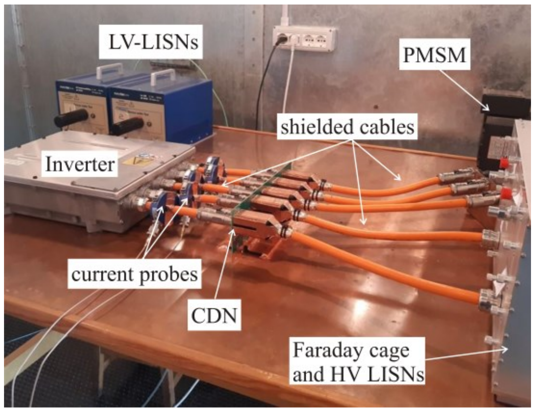

Fig. 6. Photograph of the test bench shown in Fig. 5.

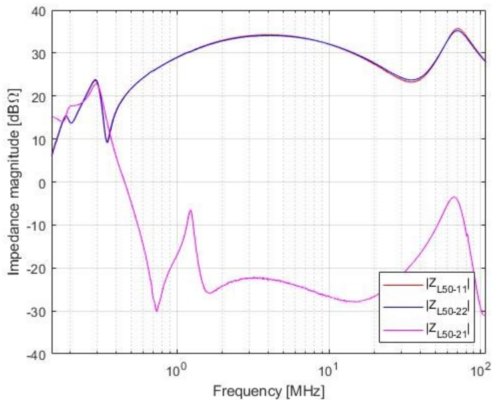

Fig. 7. Magnitude of the impedance composing $\bar{\bar{Z}}_L$ with the HV LISN loaded by $R_{LP} = R_{LN} = 50\Omega$.

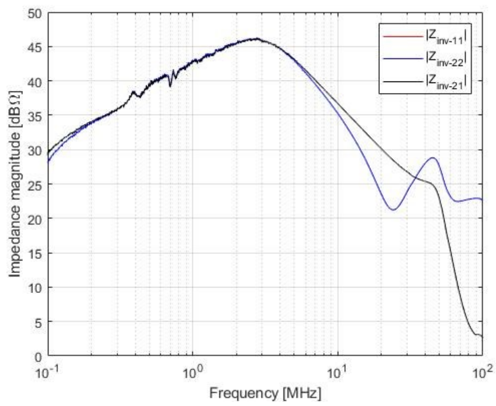

Fig. 8. Magnitude of the impedance composing $\bar{\bar{Z}}_{INV}$ resulted from (1).

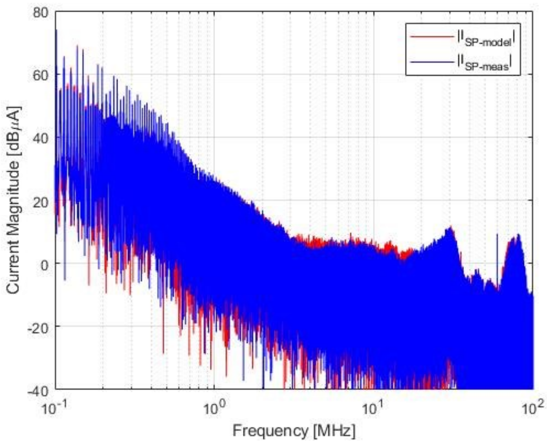

Fig. 9. Spectrum of the current flowing in the positive power cable P resulted from the model prediction (red) and from the measurement (blu) carried out with the HV LISN outputs loaded by $R_{LP} = R_{LN} = 8\Omega$.

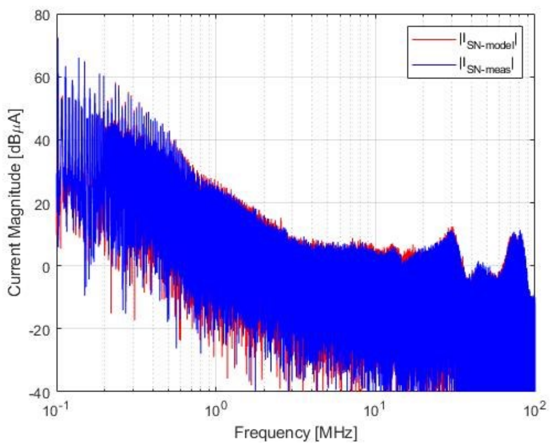

Fig. 10. Spectrum of the current flowing in the negative power cable N resulted from the model prediction (red) and from the measurement (blu) carried out with the HV LISN outputs loaded by $R_{LP} = R_{LN} = 8\Omega$.

RF-N of the CDN, with the inverter disconnected from the CDN (power connectors P and N) and the LISN output ports loaded by $R_{LP} = R_{LN} = 50\Omega$. The impedance magnitudes resulted from that measurement is show in Fig. 7. The measurement of the scattering parameters was repeated with the inverter connected to the CDN. It was supplied at the HV and LV ports and controlled to be in idle mode, i.e., with output phase UVW not switching. The magnitude of the impedance belonging to $\bar{\bar{Z}}_{INV}$, which resulted from (1), are show in Fig. 8. About these plots, it is worth mentioning that $|Z_{INV-11}|$ and $|Z_{INV-11}|$ are overlapped. Finally, the inverter was controlled to run the motor unloaded at $\omega_m = 1000$ rpm, and the switching currents flowing through the power supply cables were sensed with two equal current probes (see CP-P, CP-N in Fig. 5) and acquired by a digital oscilloscope simultaneously. Then, the equivalent voltage sources v_{CM} and v_D were evaluated using the approach presented in Section II.

The behavioral model of the power inverter was validated comparing the current spectra obtained from the simulation of a circuit comprising the inverter model and the high frequency model of the HV LISNs loaded by $R_{LP} = R_{LN} = 8\Omega$ with those resulted from the measurements. Figs. 9 and 10 show the magnitude of the simulated current spectra (red) and the measured one (blue), which differ by less than ±3dB in the frequency range 150kHz – 108MHz.

IV. Conclusions

A new method to derive the behavioral model of dc supplied power switching circuits has been proposed. It is based on the Thevenin equivalent circuit and on the evaluation of its unknowns from two sets of separate measurements; that needed to evaluate the impedance matrix and that needed to obtain the equivalent voltage sources. The proposed solution was used to model a high voltage three phase inverter driving an electric motor. The model was validated comparing the power supply current spectra resulted from the simulations with those resulted from the measurements under different loading conditions.

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