Assessment of Equivalent Noise Source Approach for EMI Simulations of Boost Converter

Vuttipon Tarateeraseth [1], Ivan A. Maio [2], Flavio G. Canavero [3]

Dipartimento di Elettronica, Politecnico di Torino
Corso Duca degli Abruzzi 24, 10129, Torino, Italy
[1]vuttipon.tarateeraseth@polito.it
[2]maio@polito.it
[3]flavio.canavero@polito.it

***Abstract*—This paper presents the analytical equations to model the equivalent noise source for conducted EMI simulation in the frequency domain. The continuous conduction mode boost converter is modeled. By means of a test case, it is shown that the equivalent noise source provides a reasonable estimate of the switching waveform and a good estimate of the EMI noise.**

I. Introduction

Nowadays, the use of power electronics is widespread, because there are significant benefits in both reduction of physical sizes and efficiency. However, power electronics increase the conducted electromagnetic interference (EMI). The conducted EMI is generated by the high frequencies inherent in fast-transition voltage and current waveforms [1]. Prior research has recommended several ways to predict the conducted EMI [5] - [7]. They can be categorized into two approaches: time domain and frequency domain approaches. The frequency domain approach is preferable because it requires shorten simulation time and has no convergence problem. However, equivalent noise sources are demanded to replace the switching elements of the circuits. In order to represent the switching elements, the waveform of the equivalent noise source has to be as similar as possible to the switching waveform. The error depends on the waveshape and parameters of the equivalent noise source.

In this paper, we apply the frequency domain method to the analysis of a boost converter circuit. The equivalent noise source required by the method is estimated by means of analytic relations. The accuracy of the approach is verified by comparison to time domain simulations and measurements.

II. Definition of the Test Case

The layout of boost converter involved in the comparison of this paper is shown in Fig. 1. This layout has been designed to validate the modeling of the noise path. The converter operates at $f_0 = 40$ kHz with 45% duty cycle; its input voltage is $V_{in} = 20$ Vdc and is applied through a Line Impedance Stabilization Network (LISN). The converter is composed of: load resistor (300 Ω), boost inductor ($L_{boost} = 470$ μH, $R_{lp} = 1.3$ Ω and $C_{lp} = 27$ pF), output capacitor ($L_{cp} =$ 25 nH, $C_o = 470$ μF, $ESR = 280$ mΩ), diode (BYW77P-220) and MOSFET (IRFP250N). The parasitic elements of the boost inductor and the output capacitor are extracted from their datasheets.

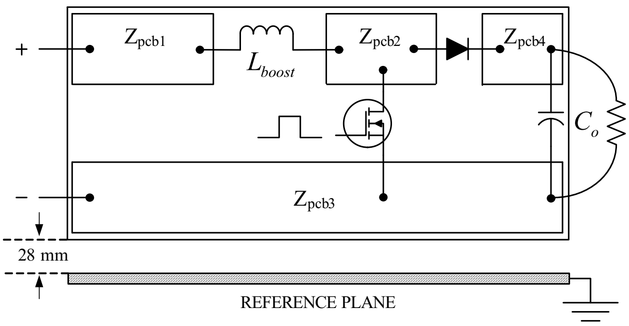


Fig. 1. Layout of boost converter

III. Modeling of the Noise Path

The modeling of Switched Mode Power Supply (SMPS) can be divided into three parts: the noise sources (switching elements: MOSFET and diode), the noise path (PCB traces and passive components) and the LISN. This section deals with the modeling of the noise path and the LISN for the test case.

The structure of the noise path is illustrated by the elemental schematic of Fig. 2, which shows the interconnection of the LISN model and of a switched mode circuit containing one switch only. The impedances of the elemental schematic represent the impedances of the connections of the switch terminals to the line terminal (L, Z_L), to the neutral terminal (N, Z_N) and to the reference terminal (Z_{Lc} and Z_{Nc}; the laboratory reference plane is defined as a perfect conductor). These impedances take into account all possible galvanic and parasitic connection between the switch and the LISN terminals. In our case, Z_L is composed of the parasitic resistance (R_{pcb}) and inductance (L_{pcb}) of PCB and boost inductor (L_{boost}), with its parasitic resistance (R_{lp}) and parasitic capacitance (C_{lp}). Z_N is composed of the parasitic resistance and inductance of PCB. According to the fact that $R_{lp} \gg \sum R_{pcb}$, $L_{boost} \gg \sum L_{pcb}$ and $Z_L \gg Z_N$, the R_{pcb} and L_{pcb} have a minor effect on both differential mode (DM) and common mode (CM) noise predictions and are ignored.

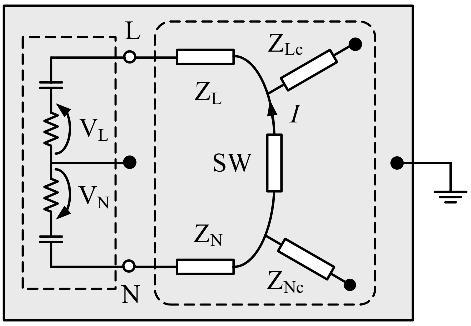


Fig. 2. Elemental schematic describing the conversion of the switch current into noise signal detected at the LISN

In contrast, Z_{Lc} and Z_{Nc}, which have a capacitive nature, play an important role to determine the common mode noise. The per-unit-length capacitance (pF/in) of straight traces over a reference plane at distance (h) can be analytically estimated by (1). The validity of this formula is $0.1 < \frac{w}{h} < 3.0$ and $1 < \epsilon_r < 15$ [3].

$$C_{pcb} = \frac{0.67(\epsilon_r + 1.41)}{ln\left(\frac{5.98h}{0.8w+t}\right)} \tag{1}$$

The traces of the PCB of the test case are 15 mm wide and 28 mm above the reference plane. The $\frac{w}{h}$ ratio is 0.5; that is in the validity range of Eq. (1). The passive components of the test circuit are modeled by the HF equivalent circuits [1]. The self resonant frequency of boost inductor (f_{rl}) is used to estimate the parallel parasitic capacitance which can be expressed as $C_{lp} = 1/[L_{boost}(2\pi f_{rl})^2]$. The LISN is modeled by its schematic (see Fig. 3). In order to implement this schematic in SPICE, $R_{dum} \geq 100M\Omega$ is added to prevent floating nodes by providing dc paths to ground [4].

IV. Frequency Domain Simulation Method

In order to carry out a frequency domain analysis of an SMPS the non-linear and time-varying elements (e.g. MOSFET and diode for a simple boost converter) of the circuit must be somehow replaced by linear elements. If the voltage or current waveforms of the unwanted elements are known, then the substitution theorem can be applied [2], making the circuit compatible with a frequency domain analysis. According to the theorem, the branch of a switching element (e.g. MOSFET) can be replaced by either one independent voltage or current source impressing waveforms identical to the original branch voltage and current, respectively. The problem is then how to obtain a reasonable estimate of the switch waveforms without carrying out measurement or a time-domain analysis. The voltage waveform across the switch depends on the operation mode of the converter. The waveforms of the discontinuous conduction mode (DCM) case are complicated and are not addressed in this paper. For the continuous conduction mode (CCM) case, the voltage across MOSFET or current passing though it is in the form of trapezoid. The choice between using trapezoidal voltage or current source as a noise source depends on the converter topology. For buck converters and three phase inverters [6], the trapezoidal current source is applied when analyzing the DM noise and the trapezoidal voltage source is utilized for the CM noise. For boost converter type, the main disturbance sources of CM and DM are represented by the voltage across the MOSFET switch [7]. Moreover, to simplify the analysis, the diode recovery effect is not taken into account. In order to estimate the parameters of the noise source, we exploit the analytic relations of [8] as discussed in the next subsection.

A. Equivalent Noise Source

The one-sided spectrum of a trapezoidal periodic waveform with $t_r \neq t_f$ is:

$$\begin{aligned} \mathcal{C}^{+}_{tra}(s) = \frac{2}{T}(\frac{V_{sw}}{s^2}) \Bigg[& \frac{1}{t_r}\left(1 - e^{-st_r}\right) \\ & -\frac{1}{t_f}\left(e^{-s(t_{off}+t_r)} - e^{-s(t_{off}+t_r+t_f)}\right)\Bigg] \end{aligned} \tag{2}$$

where

$$\begin{aligned} s &= jn\omega_0 \quad \{n = 1, 2, \ldots\} \\ t_r &= \text{Time taken for the voltage to rise to its off-state value during the turn-off transient [s]} \\ t_f &= \text{Time taken for the voltage to fall to its on-state value during the turn-on transient [s]} \\ t_{off} &= \text{Off state of the switch [s]} \\ T &= \frac{2\pi}{\omega_0} = \text{Switching period [s]} \end{aligned}$$

The parameter V_{sw} is the amplitude of the voltage waveform across the MOSFET. V_{sw} depends on the type of converter. In this paper,the CCM boost converter is investigated and V_{sw} is given by Eq. (3).

$$V_{sw} = V_{DS} = \frac{V_{in}}{1-D} \tag{3}$$

where

$$\begin{aligned} V_{in} &= \text{Input voltage [V]} \\ D &= \text{Duty cycle} \\ &= [t_{off} + t_r + t_f]/T \end{aligned}$$

According to [8], the t_r, t_f and t_{off} of voltage across MOSFET are estimated by using Eqs. (4) - (6), respectively.

$$t_r = \frac{Q_{gd\_d}(R_g + R_{g\_app})}{V_{gp}} \tag{4}$$

$$t_f = \frac{Q_{gd\_d}(R_g + R_{g\_app})}{V_{GS\_app} - V_{gp}} \tag{5}$$

$$\begin{aligned} t_{off} = \; & t_{g\_off} + (R_g + R_{g\_app})(C_{iss}^{@V_{ds}}) \\ & \cdot \ln\left(\frac{V_{gp}}{V_{GS(th)}\left[1 - \frac{V_{gp}}{V_{GS\_app}}\right]}\right) \end{aligned} \tag{6}$$

where

$$\begin{aligned}
V_{gp} &= \text{Gate plateau voltage [V]} \\
V_{GS(th)} &= \text{Gate threshold voltage [V]} \\
V_{GS_app} &= \text{Applying gate source voltage [V]} \\
Q_{gd_d} &= \text{Gate drain charge specified in datasheet [C]} \\
C_{iss}^{@V_{ds}} &= \text{Input capacitance at appropriate } V_{ds}\text{[F]} \\
R_g &= \text{Gate resistance internal of MOSFET}[\Omega] \\
R_{g_app} &= \text{Resistance applied to gate of MOSFET}[\Omega] \\
t_{g_off} &= \text{Off period of gate drive voltage [s]}
\end{aligned}$$

The numerical data of these equations are retrieved from MOSFET's datasheet, thereby obtaining the equivalent noise sources without measurement or simulation.

V. Output noise

The model of the test case that has been developed in section III is shown in Fig. 3. In order to check the accuracy of this model, the transient analysis is carried out by SPICE. The noise spectra are computed via Fast Fourier Transform (FFT) by using the following simulation settings. These settings guarantee the correct evaluation of the spectra in the 50 kHz bandwidth and 30 MHz sweep range.

According to the Nyquist criterion, the printing increment is about $T_{step} \leq 1/(2f_{max}) \leq 16.66$ ns to reach the upper frequency of conducted EMI (30 MHz). The maximum time step is about $T_{max} = T_{step}/2 \cong 8.33$ ns to prevent aliasing problem. To obtain 50 kHz bandwidth, the difference between T_{start} and T_{stop} is about $BW = 1/f_{BW} = 20$ μs [5].

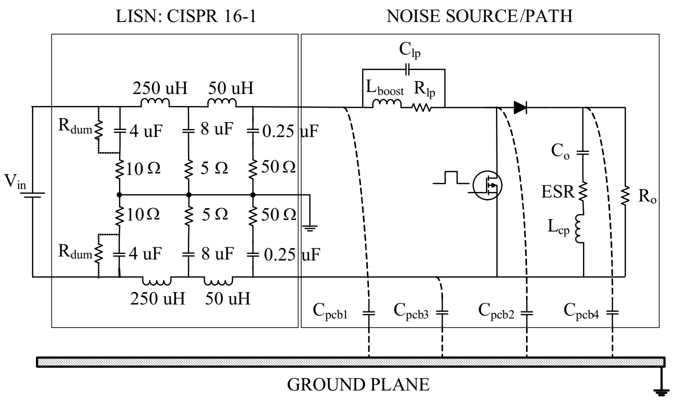


Fig. 3. Modeling for time domain simulation

The accuracy of the model is assessed by comparing its responses to the measurements of the actual voltage across the MOSFET and the voltages across LISN at line-ground (v_{LG}) and neutral-ground (v_{NG}) terminals. The v_{LG} and v_{NG} are measured by a digital oscilloscope with sampling rate $f_s =$ 250 MS/s and number of samples $N = 5002$; the bandwidth of measurement settings is $BW = f_s/N = 50$ kHz [9]. The measured DM and CM noise voltages are calculated by Eqs. (7) - (8); the Discrete Fourier Transforms (DFT) of DM and CM noise voltages are computed off-line by using MATLAB.

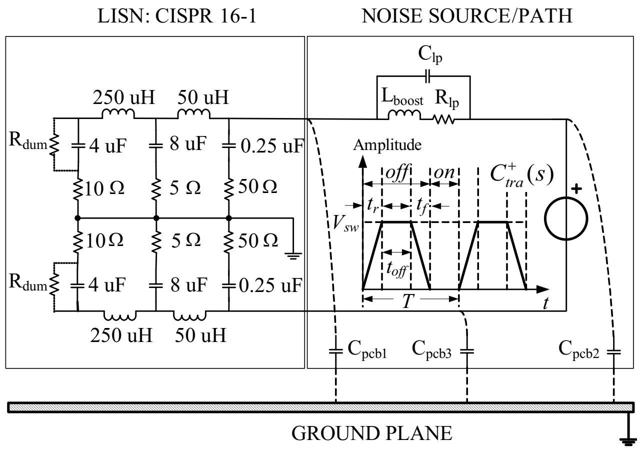


Fig. 4. Simplified equivalent circuit for frequency domain simulation

TABLE I
Comparison among the experiment, simulation and analytical calculation

	Measured Result	Simulated Result	Analytical Prediction[1]
t_r (ns)	76	76	81.6
t_f (ns)	44	54	40.8
t_{off} (μs)	12.86	12.91	12.89

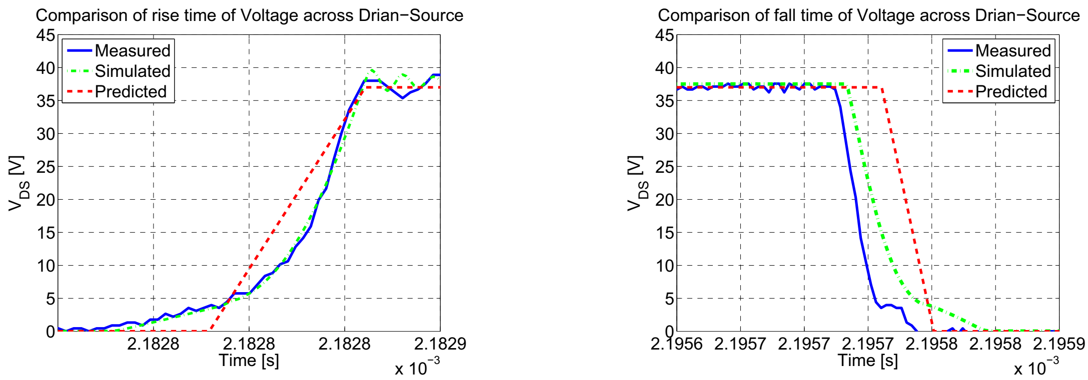
[1]IRF250P datasheet: Q_{gd_d}@$40V_{ds} \simeq 30$ nC, C_{iss}@$40V_{ds} \simeq 2200$ pF, $V_{GS\_app} = 15$ V, $R_{g\_app} = 10$ Ω, $R_g = 3.6$ Ω (from SPICE), $V_{GS(th)} = 4$ V, $V_{gp} \simeq 5$ V

$$v_{dm}(t) = \frac{v_{LG}(t) - v_{NG}(t)}{2} \tag{7}$$

$$v_{cm}(t) = \frac{v_{LG}(t) + v_{NG}(t)}{2} \tag{8}$$

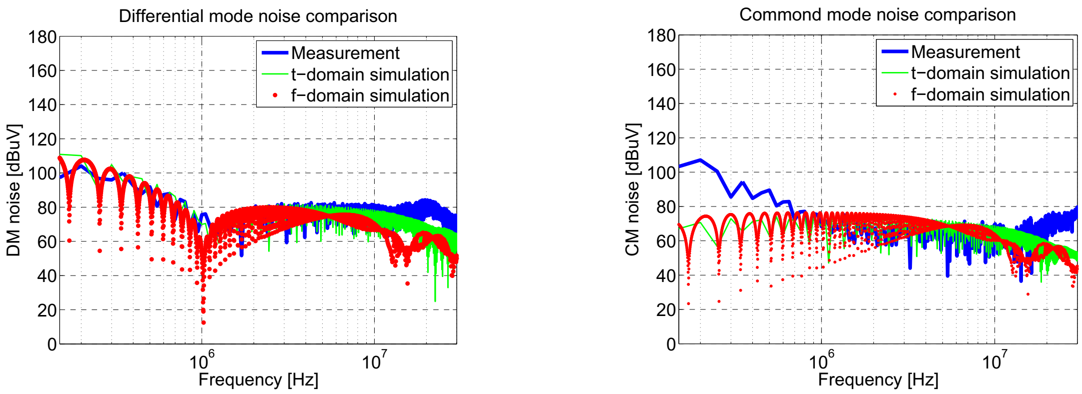
The equivalent noise source defined in section IV is verified by comparing it to the measured and simulated waveforms of the voltage across MOSFET terminals; i.e. time domain simulation and analytical prediction of V_{DS} are compared. As shown in Table I and Fig. 5, the estimated equivalent source is pretty close to the actual switch waveform. To validate the model in terms of conducted EMI, the DM and CM noise predictions in the frequency range 150 kHz - 30 MHz computed by time domain approach are compared to the measurement as shown in Fig 6. The model provides a good agreement for DM noise prediction, but not for CM noise prediction. For CM noise, the difference may be caused by lack of dominant common mode noise paths. Furthermore, the overestimation of the parasitic parameter of passive components might affect to the difference in high frequency of both DM and CM noise spectrum [6].

In order to estimate the error introduced by the frequency domain approach, the same model of the test case used in time domain approach is applied. However, the MOSFET is replaced by the equivalent noise source. As indicated by the voltage shift theorem [2], some components (diode, output capacitor, C_{pcb4}, load resistor) can be ignored as shown in Fig.



(a) Comparison of t_r of voltage across MOSFET (b) Comparison of t_f of voltage across MOSFET

Fig. 5. Comparison of voltage across MOSFET (Drain-Source)



(a) Differential mode noise comparison (b) Common mode noise comparison

Fig. 6. Conducted EMI comparison

4. For the SPICE implementation, the ac analysis in SPICE (sweeping frequency from 150 kHz - 30 MHz) is applied. The equivalent noise source is represented by a voltage-controlled voltage source ($E_{laplace}$) in Laplace form, following Eq. (2). $E_{laplace}$ is excited by ac voltage source (1 Vac). As illustrated in Fig. 6, it is clear that the time domain and frequency domain simulations are in a good agreement. Nevertheless, the frequency domain approach provides shorter simulation times and eliminates the convergence problems.

VI. Conclusions

This paper has shown that the equivalent noise source for conducted EMI simulation of a continuous conduction mode boost converter can be achieved by using the simple generic equations and MOSFET's datasheet. The accuracy of equivalent noise source is verified by comparing the analytical prediction of voltage across MOSFET to the time domain simulation and the measurements. The good approximation by using analytical approach is demonstrated. For conducted EMI prediction, it is shown that both simulation approaches are in a good agreement, yet the frequency domain approach offers shorter simulation times and is free from convergence problems. The equivalent noise source has been proven to be accurate and efficient enough for the conducted EMI prediction in single-switch DC-DC converters working in continuous conduction mode.

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