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Energy-based modelling and comparison of experimental and real-time simulated Boost converter with parasitics

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ABSTRACT This paper presents an energy-based modeling approach for a boost converter including parasitic effects. The model is derived using the Lagrangian formulation with Rayleigh dissipation, allowing the inductor DC resistance, capacitor ESR, switch resistance, diode resistance, and diode forward voltage drop to be incorporated in a systematic way. From the averaged representation, the non-ideal static gain, the small-signal linearized model, and the duty-cycle-to-output-voltage and duty-cycle-to-inductor-current transfer functions are obtained. The model is validated using a 25 kHz, 12 V-to-24 V experimental prototype and a switched real-time HIL simulation. Frequency-domain results show that the ideal model overestimates the resonant peak and does not reproduce the damping observed experimentally, whereas the parasitic-aware model provides a closer approximation of the measured gain and phase responses. Time-domain tests under duty-cycle and load steps further confirm that the inclusion of parasitic elements is essential to reproduce the mean voltage and current dynamics of the practical converter. Although algebraically equivalent to conventional state-space averaging under the same assumptions, the energy-based formulation provides a structured separation between stored energy, dissipative effects, and external sources, facilitating the derivation and extension of parasitic-aware converter models.

KEYWORDS Boost converter, Energy-based modelling, Real-time simulation.

I. INTRODUCTION

Power electronics design increasingly depends on accurate modelling approaches to predict converter operation under practical conditions. Although conventional state-space averaging [1] remains widely adopted for Boost converter analysis [2], parasitic resistances, semiconductor voltage drops, and other non-ideal effects can also be incorporated into its switching-state equations and averaged representation [3]. Recent works indicate that energy-based techniques, such as the Lagrangian formulation, allow parasitic effects to be incorporated in a systematic manner while preserving the physical meaning of the model [4], [5], especially in efficiency-oriented applications such as renewable energy systems [6] and electric vehicle power-trains [7]. The relevance of parasitic-aware modeling has further increased with the use of wide-bandgap devices, for which frequency-dependent conduction effects may not be adequately represented by simplified loss models [8].

Despite these developments, three important gaps remain in power electronics education and industrial practice. First, most textbooks [9] still present converter models under ideal assumptions, which limits the preparation of students to

deal with practical non-idealities. Second, although real-time simulation platforms such as Typhoon HIL [10] support fast prototyping, their accuracy strongly depends on adequate parasitic parametrization, a step often neglected in practical implementations [11]. Third, experimental comparisons involving energy-based models and hardware prototypes have been reported [12], but combined validation of static, frequency-domain, and time-domain behavior under different operating transients remains limited. This is particularly relevant for control design, since neglected parasitic elements may modify the resonant and damping characteristics of the converter and consequently affect control-oriented predictions [13].

The Lagrangian formulation is not intended to produce converter dynamics that cannot be obtained from Kirchhoff's laws or conventional state-space averaging. Under the same assumptions and parameter values, both approaches must lead to equivalent models. The distinction lies in the derivation procedure: while the conventional approach is generally organized around the circuit equations of each switching interval, the energy-based formulation separates stored energy, resistive dissipation, and external sources into

distinct mathematical terms. This distinction offers limited simplification for the second-order Boost converter, whose conventional equations are already relatively simple, but provides a systematic and reusable structure that can become more relevant for converters containing a larger number of energy-storage elements and parasitic branches.

This work extends the energy-based modelling methodology previously applied by the authors to a Buck converter in a conference paper [14]. In contrast to that preliminary study, the present work investigates a Boost converter and provides a more comprehensive analysis, including additional parasitic elements, detailed small-signal modelling, a normalized analysis of the influence of each parasitic parameter on the static gain, natural frequency, resonant peak, and damping ratio, parameter characterization, and dynamic validation under multiple operating transients.

This work addresses these gaps through four main contributions: (1) a complete Lagrangian formulation of the Boost converter, explicitly including the inductor DC resistance (DCR) r_L , capacitor equivalent series resistance (ESR) r_C , switch resistance r_{sw} , diode resistance r_D , and diode forward voltage drop V_D ; (2) analytical derivation of the non-ideal static gain, small-signal model, and duty-cycle-to-output-voltage and duty-cycle-to-inductor-current transfer functions; (3) a normalized parameter-sensitivity analysis addressing the static gain, natural frequency, resonant peak, and damping ratio; and (4) experimental validation using a 25 kHz, 12 V-to-24 V prototype, including comparisons with a Typhoon HIL switch-level simulation and the averaged analytical models under multiple test conditions.

The paper is organized as follows: Section II derives the energy-based model with parasitics, Section III describes the experimental and validation procedure and presents the comparative results for all test cases, including the frequency-domain, time-domain, and parameter-sensitivity analyses, and Section IV summarizes the conclusions of the work.

II. BOOST CONVERTER MODELING

The Boost converter is DC-DC non-isolated single switched converter. Its schematics is shown in Fig. 1, already taking into account parasitics effects.

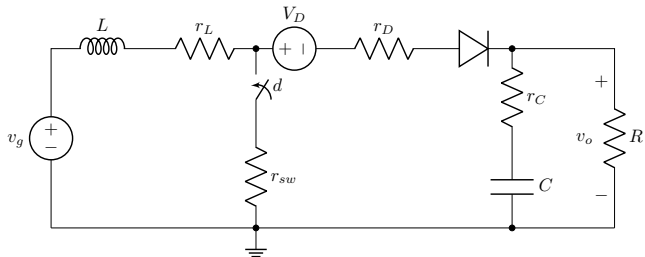


FIGURE 1. Schematics of the boost converter with parasitic effects.

The boost converter analysed in this paper operates in Continuous Conduction Mode (CCM). In this mode, the

inductor current never reaches zero during the switching cycle, which results in two distinct phases: when switch is on and when it is off. During these intervals, the switch and the diode operate in a complementary manner, ensuring continuous energy transfer to the load. In this mode of operation its ideal static voltage gain is described by the following equation:

$$\frac{V_o}{V_{in}} = \frac{1}{1-d} \quad (1)$$

with d being the duty-cycle of the switch.

The dynamical model is obtained using the Lagrange Energy Method, using generic charges q_L and q_C for the inductor and capacitor, respectively. Given $i_L = \dot{q}_L$ and $q_C = Cv_C$, it follows that

$$\dot{q}_C = i_C = C\dot{v}_C. \quad (2)$$

For the boost converter, the current delivered to the output stage depends on the switching state. Defining $u = 1$ for the closed-switch interval and $u = 0$ for the open-switch interval, the output current is given by

$$i_R = (1-u)i_L - i_C. \quad (3)$$

Therefore, the output voltage can be written as

$$v_o = R[(1-u)i_L - i_C]. \quad (4)$$

The kinetic energy $\mathcal{T}$ is related to the inductor current $i_L = \dot{q}_L$, whilst the potential energy $\mathcal{U}$ is related to the electric charge q_C . Since resistive parasitic elements are considered, the Rayleigh dissipation term $\mathcal{D}$ must also be included. For the boost converter of Fig. 1, the energy terms are given by

$$\begin{aligned} \mathcal{T} &= \frac{1}{2}L\dot{q}_L^2, & \mathcal{U} &= \frac{1}{2C}q_C^2, \\ \mathcal{D} &= \frac{1}{2}[r_L + ur_{sw} + (1-u)r_D]\dot{q}_L^2 + \frac{1}{2}r_C\dot{q}_C^2 \\ &\quad + \frac{1}{2}R[(1-u)\dot{q}_L - \dot{q}_C]^2. \end{aligned} \quad (5)$$

The diode forward voltage drop is included as a generalized force, since it is not a quadratic dissipative term. Thus, the external forces are given by

$$\mathcal{F}_L = v_g - (1-u)V_D, \quad \mathcal{F}_C = 0. \quad (6)$$

The Lagrange Energy Method states that $\mathcal{L} = \mathcal{T} - \mathcal{U}$, with its differential equation given by

$$\frac{d}{dt} \left(\frac{\partial \mathcal{L}}{\partial \dot{q}_k} \right) - \frac{\partial \mathcal{L}}{\partial q_k} + \frac{\partial \mathcal{D}}{\partial \dot{q}_k} = \mathcal{F}_k. \quad (7)$$

The subindex k might be regarded as L for the inductor or C for the capacitor. Solving Eq. (7) results in

$$\begin{aligned} L\dot{i}_L + [r_L + ur_{sw} + (1-u)r_D]i_L \\ + (1-u)R[(1-u)i_L - i_C] \\ = v_g - (1-u)V_D, \end{aligned} \quad (8)$$

$$v_C - R[(1-u)i_L - i_C] + r_C i_C = 0. \quad (9)$$

From Eq. (9), the capacitor current can be written as

$$i_C = \frac{(1-u)Ri_L - v_C}{R + r_C}. \quad (10)$$

For compactness, the following auxiliary parameters are defined:

$$\rho = R + r_C, \quad \alpha = \frac{R}{\rho}, \quad \beta = \frac{Rr_C}{\rho}. \quad (11)$$

Remarking that $v_o = v_C + r_C i_C$, and substituting Eq. (10), it is possible to state that

$$v_o = \alpha v_C + (1-u)\beta i_L. \quad (12)$$

Considering i_L and v_C as states and v_o as output, the boost converter with parasitics can be represented in state-space form. Since $u \in \{0, 1\}$, one has $(1-u)^2 = (1-u)$. Therefore, substituting Eq. (12) into Eq. (8), the inductor-current dynamics becomes

$$\dot{i}_L = -\frac{r_u}{L}i_L - \frac{(1-u)\alpha}{L}v_C + \frac{v_g - (1-u)V_D}{L}, \quad (13)$$

where

$$r_u = r_L + ur_{sw} + (1-u)(r_D + \beta). \quad (14)$$

Similarly, from Eq. (10), the capacitor-voltage dynamics is obtained as

$$\dot{v}_C = \frac{(1-u)\alpha}{C}i_L - \frac{1}{\rho C}v_C. \quad (15)$$

The averaged model is obtained by considering $u = d$, where d is the duty ratio. Thus,

$$r_d = r_L + dr_{sw} + (1-d)(r_D + \beta). \quad (16)$$

Finally, the boost converter with parasitic effects can be represented as

$$\begin{bmatrix} \dot{i}_L \\ \dot{v}_C \end{bmatrix} = \begin{bmatrix} -\frac{r_d}{L} & -\frac{(1-d)\alpha}{L} \\ \frac{(1-d)\alpha}{C} & -\frac{1}{\rho C} \end{bmatrix} \begin{bmatrix} i_L \\ v_C \end{bmatrix} + \begin{bmatrix} \frac{v_g - (1-d)V_D}{L} \\ 0 \end{bmatrix} \quad (17)$$

The output equation is given by

$$v_o = [(1-d)\beta \quad \alpha] \begin{bmatrix} i_L \\ v_C \end{bmatrix}. \quad (18)$$

Therefore,

$$v_o = (1-d)\beta i_L + \alpha v_C. \quad (19)$$

The model in (17)–(19) is algebraically equivalent to that obtained through conventional state-space averaging. In fact, evaluating (13)–(15) for $u = 1$ and $u = 0$ produces the conventional state equations for the on and off intervals, respectively, and their duty-weighted average directly recovers (17). The distinction therefore lies in the derivation procedure rather than in the resulting converter dynamics. While the conventional approach typically introduces the parasitic elements separately into the circuit equations of each switching interval, the present formulation groups stored energy in $\mathcal{T}$ and $\mathcal{U}$, resistive effects in $\mathcal{D}$, and the

diode voltage drop in the generalized force. This separation provides a systematic basis for extending the formulation to converters with additional storage elements and parasitic branches.

A. Static gain with parasitic effects

The static conversion ratio of the boost converter can be obtained directly from the equilibrium conditions of the averaged model. At steady state, considering a constant duty ratio d , the state derivatives are zero. Therefore, from Eq. (17),

$$0 = v_g - (1-d)V_D - (1-d)\alpha v_C - [r_L + dr_{sw} + (1-d)(r_D + \beta)]i_L, \quad (20)$$

$$0 = -\frac{1}{\rho C}v_C + (1-d)\frac{\alpha}{C}i_L. \quad (21)$$

From Eq. (21), the capacitor voltage is

$$v_C = (1-d)\alpha \rho i_L. \quad (22)$$

Since $\alpha = R/\rho$, it follows that

$$v_C = (1-d)Ri_L. \quad (23)$$

Using the output-voltage relation

$$v_o = \alpha v_C + (1-d)\beta i_L, \quad (24)$$

and substituting Eq. (23), one obtains

$$v_o = (1-d)Ri_L. \quad (25)$$

Substituting Eq. (23) into Eq. (20), the steady-state inductor current is given by

$$i_L = \frac{v_g - (1-d)V_D}{r_L + dr_{sw} + (1-d)(r_D + \beta) + (1-d)^2\alpha R}. \quad (26)$$

Therefore, from Eq. (25), the non-ideal static voltage gain is

$$M_{\text{Boost},p}(d) = \frac{v_o}{v_g} = \frac{(1-d)R}{\Omega_{\text{Boost}}} \left[1 - (1-d)\frac{V_D}{v_g} \right], \quad (27)$$

where

$$\Omega_{\text{Boost}} = r_L + dr_{sw} + (1-d)(r_D + \beta) + (1-d)^2\alpha R. \quad (28)$$

In the ideal case, by neglecting the parasitic resistances and the diode voltage drop, namely

$$r_L = r_{sw} = r_D = r_C = 0, \quad V_D = 0, \quad (29)$$

one has $\alpha = 1$, $\beta = 0$, and

$$\Omega_{\text{Boost}} = (1-d)^2 R. \quad (30)$$

Thus, Eq. (27) reduces to the classical CCM static gain of the ideal boost converter:

$$M_{\text{Boost},i}(d) = \frac{v_o}{v_g} = \frac{1}{1-d}. \quad (31)$$

Fig. 2 compares the ideal and non-ideal static gains as functions of the duty ratio. The ideal model predicts an unbounded gain as d approaches unity. In contrast, the non-ideal model presents a finite maximum gain and then decreases for very high duty ratios, due to the increasing effect of conduction losses and the reduced effective energy transfer to the output stage.

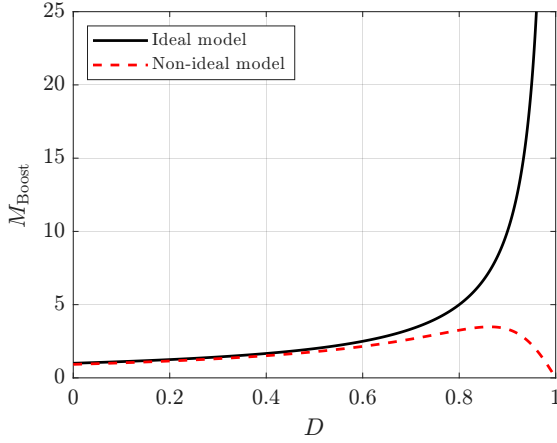


FIGURE 2. Ideal and non-ideal static gain of the boost converter.

B. Small-signal linearization

The averaged model in Eqs. (17)–(18) is nonlinear with respect to the duty ratio, since the state matrix and the output equation depend explicitly on d . Therefore, for frequency-domain analysis and control-oriented interpretation, the model is linearized around a steady-state operating point

$$\bar{x} = \begin{bmatrix} \bar{i}_L \\ \bar{v}_C \end{bmatrix}, \quad d = \bar{d}. \quad (32)$$

Considering small perturbations around this point,

$$i_L = \bar{i}_L + \hat{i}_L, \quad v_C = \bar{v}_C + \hat{v}_C, \quad d = \bar{d} + \hat{d}, \quad (33)$$

and neglecting second-order terms, the small-signal model is written as

$$\dot{\hat{x}} = A\hat{x} + B_d\hat{d}, \quad \hat{v}_o = C_v\hat{x} + D_d\hat{d}, \quad (34)$$

where

$$\hat{x} = \begin{bmatrix} \hat{i}_L \\ \hat{v}_C \end{bmatrix}. \quad (35)$$

Defining

$$\bar{r}_d = r_L + \bar{d}r_{sw} + (1 - \bar{d})(r_D + \beta), \quad (36)$$

and

$$\Gamma = V_D + \alpha\bar{v}_C - (r_{sw} - r_D - \beta)\bar{i}_L, \quad (37)$$

the linearized matrices are given by

$$A = \begin{bmatrix} -\frac{\bar{r}_d}{L} & -\frac{(1 - \bar{d})\alpha}{L} \\ \frac{(1 - \bar{d})\alpha}{C} & -\frac{1}{\rho C} \end{bmatrix}, \quad (38)$$

$$B_d = \begin{bmatrix} \frac{\Gamma}{L} \\ -\frac{\alpha\bar{i}_L}{C} \end{bmatrix}, \quad (39)$$

and

$$C_v = [(1 - \bar{d})\beta \quad \alpha], \quad D_d = -\beta\bar{i}_L. \quad (40)$$

The term Γ represents the effective small-signal gain from duty-cycle perturbations to the inductor-current dynamics. It includes the diode voltage drop, the capacitor voltage contribution, and the variation of the equivalent series resistance associated with the switching state. In addition, the direct feedthrough term D_d appears due to the capacitor ESR contribution in the output-voltage equation.

C. Transfer function representation

The transfer functions from the duty-cycle perturbation $\hat{d}$ to the inductor current $\hat{i}_L$ and output voltage $\hat{v}_o$ are obtained from the linearized model in Eq. (34). For compactness, let

$$a_{11} = -\frac{\bar{r}_d}{L}, \quad a_{12} = -\frac{(1 - \bar{d})\alpha}{L}, \quad (41)$$

$$a_{21} = \frac{(1 - \bar{d})\alpha}{C}, \quad a_{22} = -\frac{1}{\rho C}. \quad (42)$$

The characteristic polynomial of the linearized converter is

$$\Delta(s) = \det(sI - A) = (s - a_{11})(s - a_{22}) - a_{12}a_{21}. \quad (43)$$

Equivalently,

$$\Delta(s) = \left(s + \frac{\bar{r}_d}{L}\right) \left(s + \frac{1}{\rho C}\right) + \frac{(1 - \bar{d})^2\alpha^2}{LC}. \quad (44)$$

The duty-cycle-to-inductor-current transfer function is

$$G_{i_L d}(s) = \frac{\hat{i}_L(s)}{\hat{d}(s)} = \frac{\frac{\Gamma}{L} \left(s + \frac{1}{\rho C}\right) + \frac{(1 - \bar{d})\alpha^2\bar{i}_L}{LC}}{\Delta(s)}. \quad (45)$$

The duty-cycle-to-output-voltage transfer function is

$$G_{v_o d}(s) = \frac{\hat{v}_o(s)}{\hat{d}(s)} = C_v(sI - A)^{-1}B_d + D_d. \quad (46)$$

Substituting Eqs. (38)–(40), one obtains

$$G_{v_o d}(s) = -\beta\bar{i}_L + \frac{N_{v_o d}(s)}{\Delta(s)}, \quad (47)$$

where

$$N_{v_o d}(s) = \left[(1 - \bar{d})\beta \left(s + \frac{1}{\rho C}\right) + \frac{(1 - \bar{d})\alpha^2}{C} \right] \frac{\Gamma}{L} - \left[\alpha \left(s + \frac{\bar{r}_d}{L}\right) - \frac{(1 - \bar{d})^2\alpha\beta}{L} \right] \frac{\alpha\bar{i}_L}{C}. \quad (48)$$

Equations (45) and (47) are the analytical small-signal transfer functions used to represent the duty-cycle-to-current and duty-cycle-to-output-voltage dynamics of the boost converter. Since the parasitic elements are retained in α , β , ρ , $\bar{r}_d$, and Γ , their effects appear explicitly in the pole locations, damping, steady-state gain, and zero dynamics.

III. EXPERIMENTAL METHODOLOGY

The experimental validation was organized to compare the proposed parasitic-aware model with the switched HIL simulation and the experimental prototype. Three complementary tests were considered: frequency-response identification, duty-cycle step response, and load step response. The frequency-response test evaluates the small-signal transfer

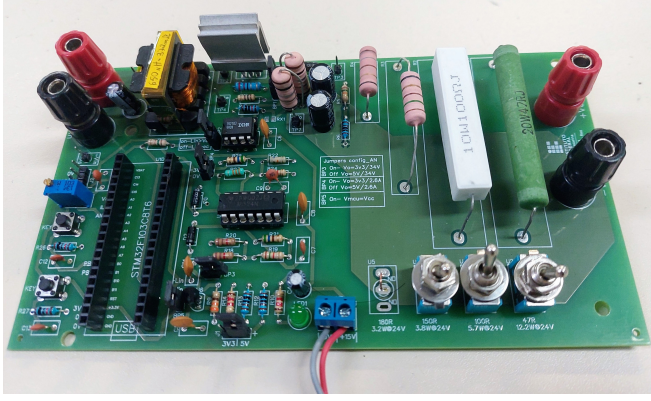


FIGURE 3. Boost converter prototype.

functions derived in the previous section, while the two time-domain tests assess the large-signal transient behavior under operating-point changes.

The comparisons are performed for the following cases:

- 1) duty-cycle-to-output-voltage and duty-cycle-to-inductor-current Bode diagrams;
- 2) 5% step change in duty cycle, $d = 0.50 \leftrightarrow 0.55$;
- 3) 45% load step between $R = 41.8\Omega \leftrightarrow 23\Omega$.

The experimental prototype specification is showed on Tab. 1.

TABLE 1. Prototype Specification

Parameter	Variable	Value
Output Power	P_o	24 W
Input Voltage	V_{in}	12 V
Output Voltage	V_o	24 V
Switching Frequency	f_s	25 kHz
Average Duty Cycle	D	0.50
Current Ripple	ΔI_{in}	20%
Input Inductance	L	650 μ H
Inductor DCR	r_L	372 m Ω
Output Capacitor	C_o	200 μ F
Capacitor ESR	r_C	260 m Ω
Switch On-Resistance	r_{sw}	40 m Ω
Diode On-Resistance	r_D	959 m Ω
Diode Forward Voltage	V_D	0.68 V

The experimental setup is based on an educational laboratory prototype available to the authors and was selected to validate the proposed modelling procedure under controlled and reproducible conditions, rather than to represent the full range of practical Boost-converter ratings.

A. Frequency response analysis

Fig. 4 and Fig. 5 compare the duty-cycle-to-output-voltage and duty-cycle-to-inductor-current responses obtained experimentally, through HIL simulation, and from the analytical

models. The first point to note is the difference between the ideal and non-ideal models. The ideal model predicts a higher resonant peak and a sharper phase transition, especially in the voltage response. This is expected, since the ideal formulation does not include the damping introduced by the inductor DCR, capacitor ESR, switch resistance, diode resistance, and diode voltage drop. When these effects are included, the analytical response approaches the HIL and experimental curves more closely. To quantify this agreement, a normalized model-fit index was calculated as

$$\text{Fit} = 100 \left(1 - \frac{\|y - \hat{y}\|}{\|y - \text{mean}(y)\|} \right), \quad (49)$$

where y denotes the experimental frequency-response data and $\hat{y}$ represents either the switched HIL simulation or the averaged parasitic-aware model. For the duty-cycle-to-output-voltage response, the magnitude fit was 92.97% for the HIL simulation and 93.77% for the averaged model, while the corresponding phase fits were 91.92% and 92.68%, respectively. For the duty-cycle-to-inductor-current response, the magnitude fits were 68.45% and 62.78%, and the phase fits were 71.85% and 71.38%, respectively. These results quantitatively confirm the close agreement observed for the voltage dynamics, while also showing that the current response presents larger discrepancies, particularly in magnitude.

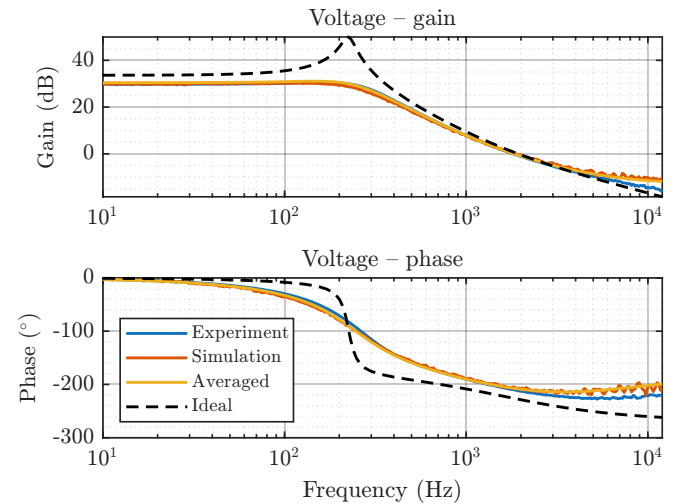


FIGURE 4. Voltage Bode Plot

A second point, more subtle, is the remaining gain difference between the experimental response and the parasitic-aware models. This difference should not be interpreted only as an error in the static conversion ratio. The Bode diagrams do not represent the static gain V_o/V_g . They represent the incremental gains obtained after linearization around a given operating point. In other words, at low frequency,

$$G_{v_o d}(0) \approx \left. \frac{\partial V_o}{\partial d} \right|_{\bar{d}}, \quad G_{i_L d}(0) \approx \left. \frac{\partial I_L}{\partial d} \right|_{\bar{d}}. \quad (50)$$

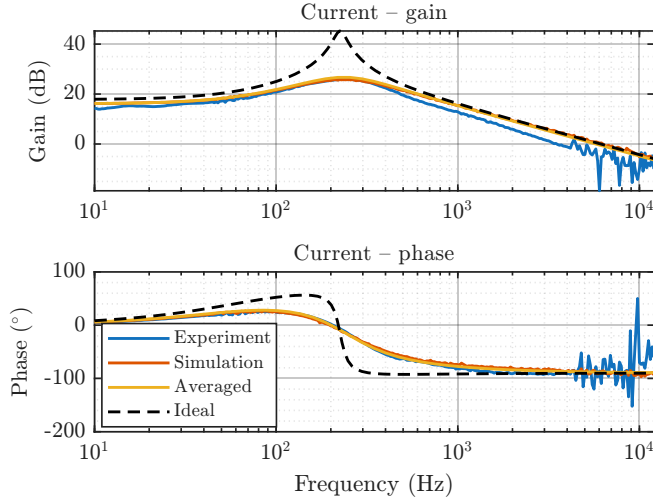


FIGURE 5. Current Bode Plot

Therefore, the measured gain is associated with the local slope of the equilibrium curve, not with the equilibrium value itself.

This distinction is important for the boost converter because its equilibrium curves are strongly nonlinear with respect to the duty ratio. For the ideal converter,

$$M_i(d) = \frac{V_o}{V_g} = \frac{1}{1-d}. \quad (51)$$

Thus,

$$V_o(d) = \frac{V_g}{1-d}, \quad (52)$$

and the low-frequency duty-cycle-to-output-voltage gain is given by the local derivative

$$\frac{\partial V_o}{\partial d} = \frac{V_g}{(1-d)^2}. \quad (53)$$

Hence, the voltage Bode diagram is related to the slope of $V_o(d)$, rather than to the value of $V_o(d)$. For example, at $d = 0.5$, the ideal converter has $V_o = 2V_g$, but the incremental gain is $4V_g$. With $V_g = 12$ V, this gives approximately 48 V per unit variation of duty ratio, or 33.6 dB, which is consistent with the low-frequency level observed in Fig. 4.

The same reasoning applies to the inductor current, but with even higher sensitivity. In the ideal case,

$$I_L(d) = \frac{V_g}{R(1-d)^2}. \quad (54)$$

Therefore,

$$\frac{\partial I_L}{\partial d} = \frac{2V_g}{R(1-d)^3}. \quad (55)$$

This cubic dependence on $1/(1-d)$ makes the current gain particularly sensitive to small differences in the actual operating point. This helps explain why the current response in Fig. 5 presents a larger spread than the voltage response, mainly at higher frequencies, where switching ripple, sensor

noise, and acquisition bandwidth also become more relevant. This higher sensitivity is also consistent with the lower fit indices obtained for the current response compared with the voltage response.

The sensitivity can be made explicit by using normalized sensitivity functions. For the ideal static gain,

$$S_{M_i}^d = \frac{d}{M_i} \frac{\partial M_i}{\partial d} = \frac{d}{1-d}. \quad (56)$$

However, for the low-frequency duty-cycle-to-output-voltage gain,

$$G_{v_{od,i}}(0) \propto \frac{1}{(1-d)^2}, \quad (57)$$

which leads to

$$S_{G_{v_{od,i}}(0)}^d = \frac{d}{G_{v_{od,i}}(0)} \frac{\partial G_{v_{od,i}}(0)}{\partial d} = \frac{2d}{1-d}. \quad (58)$$

For the duty-cycle-to-inductor-current gain,

$$G_{i_{Ld,i}}(0) \propto \frac{1}{(1-d)^3}, \quad (59)$$

and therefore

$$S_{G_{i_{Ld,i}}(0)}^d = \frac{3d}{1-d}. \quad (60)$$

At the nominal operating point $d = 0.5$, these sensitivities are

$$S_{M_i}^d = 1, \quad S_{G_{v_{od,i}}(0)}^d = 2, \quad S_{G_{i_{Ld,i}}(0)}^d = 3. \quad (61)$$

Thus, the small-signal gains observed in the Bode diagrams are more sensitive to the duty ratio than the static conversion ratio itself.

In the non-ideal model, the same interpretation remains valid, but the equilibrium curve is modified by the parasitic elements. The static gain is

$$M_p(d) = \frac{(1-d)R}{\Omega_{\text{Boost}}} \left[1 - (1-d) \frac{V_D}{V_g} \right], \quad (62)$$

with

$$\Omega_{\text{Boost}} = r_L + dr_{sw} + (1-d)(r_D + \beta) + (1-d)^2 \alpha R. \quad (63)$$

Consequently, the low-frequency gains are associated with the local slopes of this non-ideal equilibrium curve. Since Ω_{Boost} depends on the load, ESR, DCR, semiconductor resistances, and diode voltage drop, small deviations in these parameters change the slope around the operating point and, therefore, change the Bode gain.

This is the main reason why a residual gain mismatch may remain even when the parasitic-aware model captures the overall frequency response correctly. The model and the experiment may have similar steady-state voltages and currents, but a small difference in effective duty ratio, load value, sensor gain, dead time, or parasitic parameter changes the local derivative used by the linearized model. Since the boost converter has high duty-ratio sensitivity, especially in the current dynamics, these small differences become visible as gain offsets in Fig. 4 and Fig. 5. Therefore, the

agreement obtained with the parasitic-aware model should be evaluated mainly by its ability to reproduce the damping, resonance region, phase trend, and overall dynamic shape, while recognizing that exact low-frequency gain matching requires very accurate identification of the effective operating point.

B. Time response analysis

Fig. 6 shows the output-voltage and inductor-current responses for positive and negative duty-cycle steps. The experimental response is compared with the switched HIL simulation, the averaged parasitic-aware model, and the ideal averaged model. The first observation is that the ideal model predicts a much larger voltage and current excursion than the prototype. It also produces an underdamped oscillatory response that is not observed with the same intensity in the experimental waveform. This behavior is consistent with the frequency-domain results: by neglecting the parasitic resistances and semiconductor voltage drops, the ideal model overestimates the conversion gain and removes an important part of the physical damping of the converter.

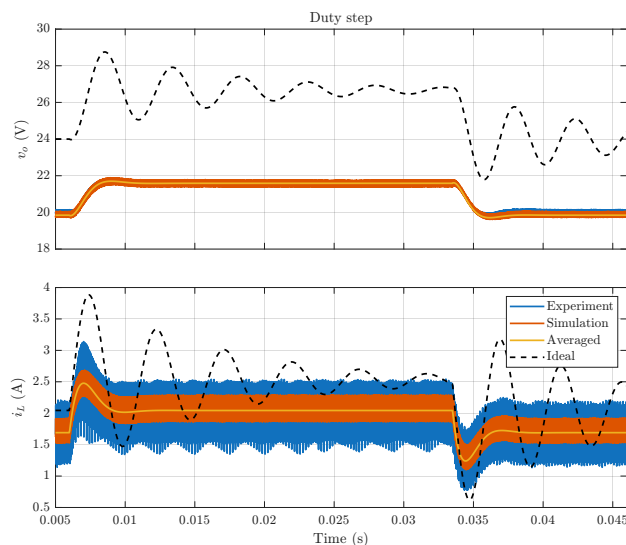


FIGURE 6. Duty-cycle step response.

The parasitic-aware averaged model gives a much more realistic prediction of the mean output voltage and mean inductor current. It does not reproduce the switching ripple, as expected from an averaged representation, but it follows the main transient envelope and the final operating levels more closely than the ideal model. The switched HIL simulation adds the ripple component and remains close to the experimental response, especially in the output voltage. In the inductor current, the experimental waveform presents larger ripple and slightly different peak values, mainly after the positive duty-cycle step. This difference is compatible with practical effects such as sensor bandwidth, PWM and driver delays, dead time, semiconductor nonlinearities, and

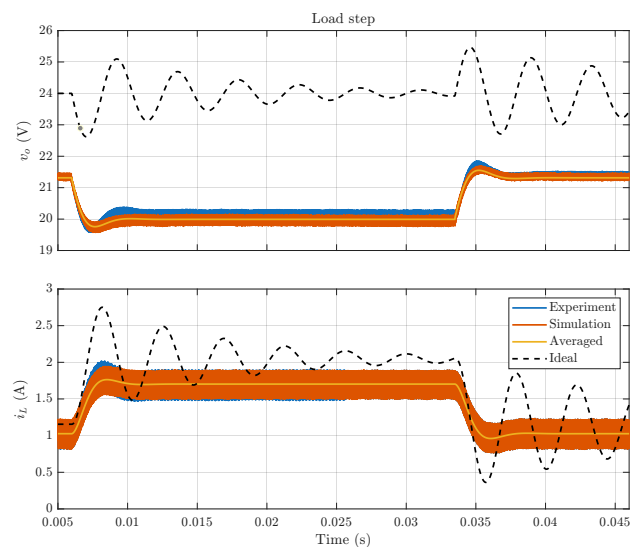


FIGURE 7. Load step response.

uncertainty in the effective values of the inductor resistance and capacitor ESR.

The duty-cycle step therefore highlights two different aspects of the validation. First, the ideal model is useful as a reference, but it is not adequate for quantitative prediction of the tested prototype, since it overestimates both the operating level and the oscillatory behaviour. Second, the parasitic-aware model captures the average dynamics with good accuracy, while the HIL simulation better represents the switched nature of the experimental converter. This separation between averaged dynamics and switching ripple is important, because the proposed energy-based model is intended to describe the dominant converter dynamics without requiring a switch-level representation.

Fig. 7 presents the response to load perturbations. When the load demand increases, the output voltage drops temporarily and the inductor current rises to supply the additional power. When the load demand decreases, the opposite behavior is observed: the output voltage presents an overshoot and the inductor current moves to a lower steady-state value. The experimental, HIL, and averaged parasitic-aware responses follow the same physical trend in both transitions.

The ideal model again deviates significantly from the practical response. Its voltage waveform remains centered around a higher operating level and exhibits larger oscillations after the load transition. This occurs because the ideal converter assumes lossless energy transfer and does not include the dissipative terms that reduce the steady-state voltage and damp the transient response. As a result, the ideal response may reproduce the qualitative direction of the transient, but it does not provide a reliable estimate of the actual voltage level, current level, or damping observed in the prototype.

The parasitic-aware averaged model provides a closer description of the mean transient behaviour during the load step. The switched HIL simulation follows the same average

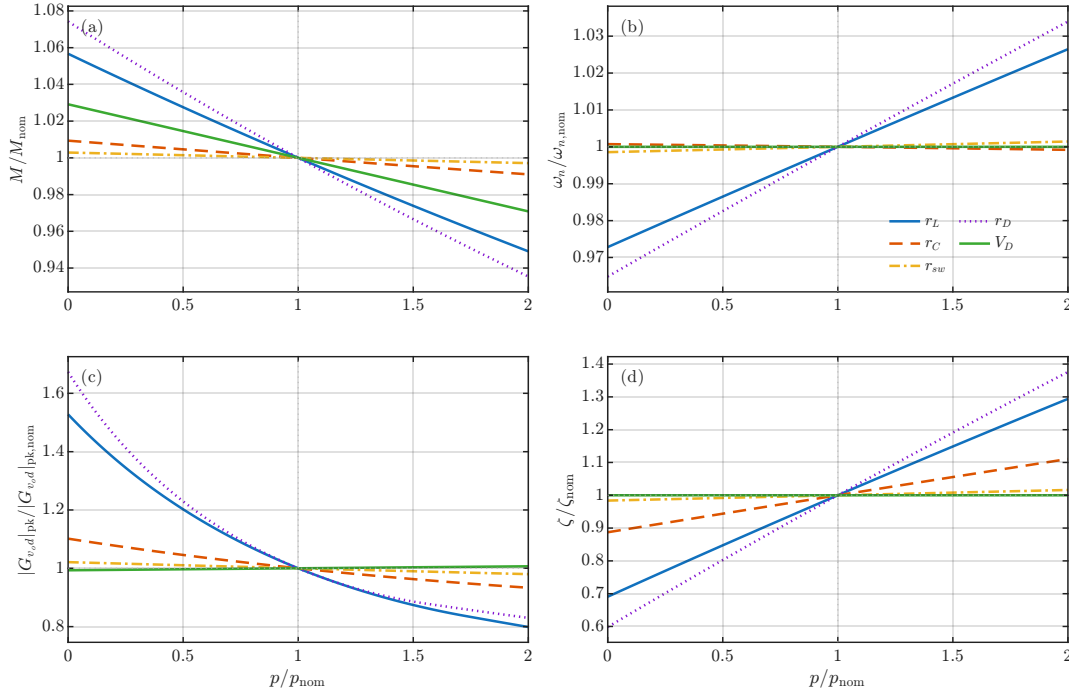


FIGURE 8. Normalized influence of the parasitic parameters on: (a) static gain, (b) natural frequency, (c) peak magnitude of G_{vod} , and (d) damping ratio.

tendency and also includes the high-frequency ripple present in the experimental data. The remaining differences between HIL and experiment are more visible in the current waveform and can be attributed to load-switching dynamics, sensor noise, acquisition bandwidth, component tolerances, and parameter variations with current and temperature. Even with these differences, the comparison confirms that including parasitic effects is essential to reproduce the dominant time-domain behavior of the practical boost converter.

C. Parameter sensitivity analysis

To quantify the influence of each parasitic element, an one-at-a-time parametric analysis is performed. Each parameter $p \in \{r_L, r_C, r_{sw}, r_D, V_D\}$ is varied from zero to twice its nominal value, while the remaining parameters are kept constant. Both the parameters and the resulting metrics are normalized with respect to their nominal values. The characteristic polynomial in (44) can be written as $\Delta(s) = s^2 + a_1s + a_0$, yielding $\omega_n = \sqrt{a_0}$ and $\zeta = a_1/(2\omega_n)$, whereas the peak magnitude is obtained directly from $G_{vod}(j\omega)$.

As shown in Fig. 8, r_D and r_L have the strongest influence at the investigated operating point. Increasing either resistance reduces the static gain and resonant peak while increasing the natural frequency and damping ratio. The capacitor ESR has a secondary effect, particularly on the peak magnitude and damping, whereas r_{sw} produces comparatively small variations. The diode forward voltage mainly affects the static gain and has negligible influence on ω_n and ζ , since it does not directly appear in the state matrix. Therefore, the converter dynamics are predominantly determined by the

parasitic resistances in the main current path. The relevance of the parasitic elements is determined not only by the absolute converter power but also by their values relative to the operating conditions. From (27), the resistive effects can be interpreted through ratios such as r_L/R , r_{sw}/R , r_D/R , and r_C/R , whereas the relative influence of the diode forward voltage is associated with V_D/V_g . Consequently, increasing the operating voltage reduces the relative contribution of an approximately constant diode forward-voltage drop, while resistive parasitics may remain significant depending on the load current and effective load resistance. Similarly, although the switching frequency does not explicitly appear in the averaged formulation, substantial changes in f_s may modify ripple and frequency-dependent semiconductor and magnetic losses, requiring the corresponding parasitic parameters to be reassessed.

IV. CONCLUSION

This paper presented an energy-based modeling approach for a boost converter including parasitic effects. The Lagrangian formulation led to a compact averaged model, from which the static gain, small-signal linearization, and duty-cycle-to-output-voltage and duty-cycle-to-inductor-current transfer functions were obtained. The experimental validation showed that the ideal model overestimates the voltage and current gains, produces excessive resonant peaks, and predicts less damping than observed in the practical converter.

By including the inductor DCR, capacitor ESR, switch resistance, diode resistance, and diode voltage drop, the proposed model provided a closer representation of both frequency-domain and time-domain responses. The Bode di-

agrams showed that the parasitic-aware model better captures the damping and phase behavior around the resonant region, while the time-domain tests confirmed that it reproduces the mean transient response more accurately than the ideal formulation. The switched HIL simulation further reproduced the ripple component observed experimentally.

The results confirm that parasitic elements are relevant not only for efficiency estimation, but also for dynamic modeling and control-oriented analysis. The energy-based formulation does not yield dynamics that could not be obtained through conventional state-space averaging; rather, it provides a structured separation between stored energy, dissipation, and external sources. Although this advantage is modest for the second-order Boost converter, the resulting framework provides a consistent basis for extending parasitic-aware modeling to converters with a larger number of energy-storage elements.

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AUTHOR'S CONTRIBUTIONS

R.J.PIONTKEWICZ: Conceptualization, Data Curation, Formal Analysis, Funding Acquisition, Investigation, Methodology, Project Administration, Resources, Software, Supervision, Validation, Visualization, Writing – Original Draft, Writing – Review & Editing. **R.TRENTINI:** Conceptualization, Data Curation, Formal Analysis, Funding Acquisition, Investigation, Methodology, Project Administration, Resources, Software, Supervision, Validation, Visualization, Writing – Original Draft, Writing – Review & Editing. **M.E.BAITTINGER:** Data Curation, Formal Analysis, Methodology. **L.V.BONALUME:** Data Curation, Formal Analysis, Investigation, Methodology.

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DATA AVAILABILITY

The data used in this research is available in the body of the document.

REFERENCES

- [1] R. D. Middlebrook, S. Cuk, "A general unified approach to modelling switching-converter power stages", in *1976 IEEE power electronics specialists conference*, pp. 18–34, IEEE, 1976, doi:10.1109/PESC.1976.7072895.
- [2] M. B. Vendruscolo, A. Toebe, E. Mattos, L. Michels, A. M. S. S. Andrade, "Modeling and Control of a Bipolar DC-DC Boost Converter for Bipolar DC Microgrids", *Eletrônica de Potência*, vol. 31, p. e202612, Feb. 2026, doi:10.18618/REP.e202612.
- [3] R. W. Erickson, D. Maksimovic, *Fundamentals of power electronics*, Springer Science & Business Media, 2007, doi:10.1007/b100747.
- [4] R. Ortega, A. Van Der Schaft, B. Maschke, G. Escobar, "Interconnection and damping assignment passivity-based control of port-controlled Hamiltonian systems", *Automatica*, vol. 38, no. 4, pp. 585–596, 2002, doi:10.1016/S0005-1098(01)00278-3.

- [5] M. U. Munieeb, S. S. Sofi, F. Bashir, M. U. Hassan, S. A. Nahvi, "Some contributions to Lagrangian modelling of power converters", *Franklin Open*, vol. 8, p. 100125, 2024, doi:10.1016/j.fraope.2024.100125.
- [6] F. Blaabjerg, F. Iov, R. Teodorescu, Z. Chen, "Power electronics in renewable energy systems", in *2006 12th International Power Electronics and Motion Control Conference*, pp. 1–17, IEEE, 2006, doi:10.1109/EPEPEMC.2006.4778368.
- [7] R. Stadlmayr, K. Schlacher, "An energy-based control strategy for DC/DC power converters", in *2009 European Control Conference (ECC)*, pp. 3967–3972, IEEE, 2009, doi:10.23919/ECC.2009.7075020.
- [8] K. Wang, B. Li, H. Li, X. Yang, A. Qiu, "Characterization and modeling of frequency-dependent on-resistance for GaN devices at high frequencies", *IEEE Transactions on Power Electronics*, vol. 35, no. 5, pp. 4925–4933, 2019, doi:10.1109/TPEL.2019.2947075.
- [9] D. W. Hart, D. W. Hart, *Power electronics*, vol. 166, McGraw-Hill New York, 2011.
- [10] J. R. López Gutiérrez, P. Ponce, A. Molina, "Real-time power electronics laboratory to strengthen distance learning engineering education on smart grids and microgrids", *Future Internet*, vol. 13, no. 9, p. 237, 2021, doi:10.3390/fi13090237.
- [11] M. İnci, Y. Altun, "Investigation on the Effect of Parasitic Elements on PID Control of DC-DC Buck Converter", *Düzce Üniversitesi Bilim ve Teknoloji Dergisi*, vol. 12, no. 2, pp. 696–708, 2024.
- [12] S. Choi, M. Saeedifard, R. Shenoy, "A Modern Education Power Electronics Laboratory to Enhance Hands-on Active Learning", in *2011 ASEE Annual Conference & Exposition*, pp. 22–67, 2011, doi:10.18260/1-2-17349.
- [13] B. Lehman, R. M. Bass, "Extensions of averaging theory for power electronic systems", *IEEE Transactions on Power electronics*, vol. 11, no. 4, pp. 542–553, 1996, doi:10.1109/63.506119.
- [14] R. J. Piontkewicz, R. Trentini, "Energy-Based Modelling and Comparison of Experimental and Real-Time Simulated Buck Converter with Parasitics", in *2025 Brazilian Power Electronics Conference (COBEP)*, pp. 1–5, 2025, doi:10.1109/COBEP66423.2025.11231266.

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