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Corrective frequency compensation for parasitics in boost power converter with sensorless current mode control

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ABSTRACT

For boost power converter with Sensorless Current Mode (SCM) control, the inductor current is acquired by current estimator instead of current sensing. Without consideration of parasitics, the estimator is low-accurate, which affects system Small Signal Model (SSM) and degrades the control performance. In this paper, the issue is studied in frequency domain, and solved by a Corrective Frequency Compensation for Parasitics (CFCP) strategy. First, transfer functions for power converter, current estimator and current mode controller are given for frequency-domain analysis. Second, without consideration of parasitics, conventional proportional integral compensation is used to improve the system stability. Furthermore, with consideration of parasitics, converter main pole is replaced by two poles while a zero emerges at the origin. In order to cancel out the influence, the proposed CFCP strategy adopts a second-order transfer function to correct zero/pole variation induced by parasitics. Finally, steady state performance and transient response of the converter are improved, which are verified by simulations and experiments.

1. Introduction

As a classical step-up converter, boost converter is widely used in wind power generation, High-Voltage Direct-Current (HVDC) and photovoltaic systems [1–3]. In order to improve the converter performance, different Current Mode (CM) control strategies are proposed owing to their potential to achieve high bandwidth, transient performance and simple compensation [4–8]. With accurate control for inductor current, the inductor delay is removed from the loop, which reduces system order. Conventionally, current sensing is required for CM control, which can be realized by multiple methods, such as sampling resistor, mirroring circuit and Hall effect sensor [9–11]. Although sampling resistor is very simple, it is non-isolative and low-accurate. Mirroring circuits are widely used in integrated circuits, but it is sensitive to EMI and the accuracy is also low. An accurate and isolative method for current sampling is using Hall effect sensor. However, Hall current sensors are relatively expensive, and the additional current sampling module degrades system stability. Furthermore, all aforementioned methods bring delay and noise to the system, and increase overall power consumption, size and cost of converter.

In order to solve the issues mentioned above, SCM control techniques are used, which realize current mode control without current sensing [12–16]. Conventionally, a SCM controller has simpler control scheme and lower cost than conventional CM controller, while achieves better performance than Voltage Mode (VM) controller. Among different SCM controllers, analog ripple based or V^2 controllers acquire inductor current information through the output voltage ripple [17–20]. Theses controllers are widely used in industrial applications, and can achieve a high control loop bandwidth. However, stability of V^2 control relies on a large ESR in output capacitor, which increases the power consumption and output voltage ripple. Besides, application of the control strategy is highly limited to buck converter. For power converter with Pulse Width Modulation (PWM), the inductor current can be estimated through output voltage, line voltage and duty cycle [21,22]. Since current sensor is replaced by current estimator, converter cost and size are saved. However, the transient and steady state performances can be influenced by a low accurate current estimator.

Furthermore, circuit parasitics are considered in SCM control to improve accuracy of current estimation and converter performance. With consideration of parasitics and thermal effects, a self-tuning estimation method is derived, which is based on the well-known RC-filter principle [23]. However, a current sink must be used to carry out the self-calibration process. Without additional sink, comprehensive and self-corrective compensation strategies are proposed for digital predictive SCM control [24,25]. Both parasitics and output voltage ripple are considered in calculating the inductor current slopes, which

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improve accuracy and convergence of current estimation, and eliminate output voltage steady state error. However, the calculations are complex, which increase cost and implementation difficulty of digital controller. Furthermore, parasitics are studied in frequency domain to optimize performance of converter with CM control [26,27]. However, the studies do not suit SCM control, since parasitics have different influence to CM and SCM controlled converters.

In order to eliminate influence of circuit parasitics to SCM control, the CFCP strategy is proposed to regain converter frequency characteristic. Based on boost converter, transfer function from reference current to output voltage is derived without and with consideration of parasitics, respectively. Without parasitics, the transfer function contains a RC main pole and a Right-Half-Plane (RHP) zero. Therefore, conventional PI compensator can be used to improve the system stability. With consideration of parasitics, the RC main pole is replaced by two poles, while a new zero emerges at the origin. To cancel out the impacts, the CFCP strategy adopts a second-order transfer function to correct zero/pole variation induced by parasitics. Since zero at the origin is eliminated, steady state performance of the converter is improved. The study improves stability of power converters under SCM control, and it benefits high performance power applications where current sensors are unavailable.

This paper is organized as follows. Based on boost converter, SCM control principles, small signal analysis and PI compensation strategy are given in Section 2. In Section 3, parasitics are considered in small signal analysis, while the CFCP strategy is derived to cancel out influence of parasitics. Section 4 verifies converter stability by open-loop bode plots. Furthermore, the robustness is proved along different loads, line voltages and output voltages. Finally, the proposed strategy is verified by experimental results in Section 5, and a brief conclusion is given in Section 6.

2. SSM for boost converter with SCM control

Boost converter scheme with SCM controller is given in Fig. 1. The outer loop adopts a PI compensator, which outputs the reference current i_{ref} for inner current loop. In order to carry out current mode control without current sensing, an estimated current i_{est} is acquired by the current estimator, which replaces conventional current sensors. The inner current loop adopts a SCM controller to calculate duty cycle d, which eliminates error between i_{ref} and i_{est}.

SSM for boost converter with SCM controller is shown in Fig. 2, where $G_{vd}(s)$ is transfer function of boost converter, $H_{PI}(s)$ is transfer function of PI compensator, $H_{dv}(s)$ and $H_{di}(s)$ are linearized transfer functions of SCM controller. The model is used for stability analysis and PI compensator design.In the following, current estimation and control strategies are given to achieve SCM control. Based on SCM controller and boost converter scheme, $\{G_{vd}(s)$,$H_{dv}(s)$, $H_{di}(s)$ $\}$ are derived.

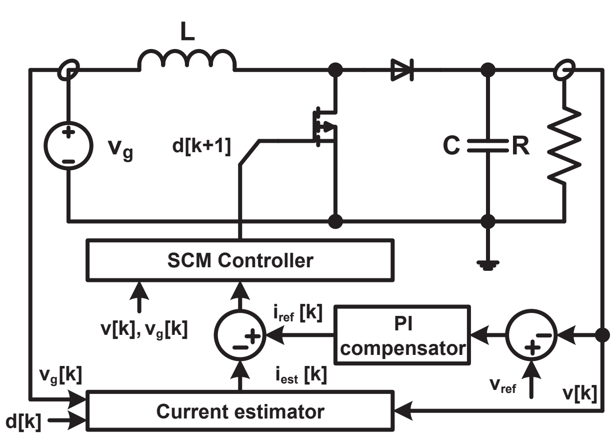

Fig. 1. Boost converter with SCM controller.

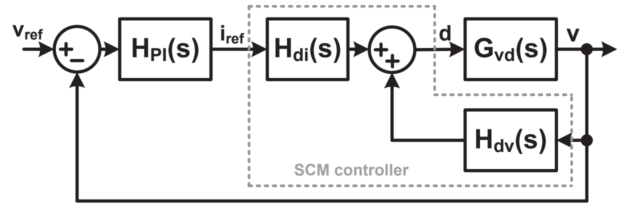

Fig. 2. SSM of the converter with SCM controller.

Furthermore, transfer function from reference current to output voltage is acquired, and it is used for PI compensator design. Finally, system cross-over frequency and phase margin are optimized.

2.1. Current estimation and SCM control strategies

Current estimation and SCM control strategies are based on inductor current formula, which is influenced by output voltage v, line voltage v_g and duty cycle d. In each switching cycle, the state-averaged voltage on the inductor is $v_g[k]-v[k](1-d[k])$, which determines the increment of inductor current value, as shown in (1).

$$i[k+1] = i[k] + \frac{T}{L}[v_g[k]-v[k](1-d[k])], \tag{1}$$

where k denotes the switching cycle. According to (1), the inductor current can be estimated by

$$i_{est}[k] = \sum_{i=1}^{k-1} \frac{T}{L}[v_g[i]-v[i](1-d[i])]. \tag{2}$$

Besides, duty cycle is derived from (1), as shown as follows

$$d[k] = \frac{(i[k+1]-i[k])L/T + v[k]-v_g[k]}{v[k]}. \tag{3}$$

Supposing $i_{est}[k]$ equals $i[k]$ and is regulated to $i_{ref}[k]$ in one switching cycle, i.e. $i[k+1] = i_{est}[k+1] = i_{ref}[k]$, the duty cycle can be calculated by

$$d[k] = \frac{(i_{ref}[k]-i_{est}[k])L/T + v[k]-v_g[k]}{v[k]}. \tag{4}$$

(4) is used for the SCM controller, which eliminates current error in one switching cycle.

2.2. Transfer function from reference current to output voltage

Transfer function from reference current to output voltage is derived from $\{G_{vd}(s), H_{dv}(s), H_{di}(s)\}$. First, $G_{vd}(s)$ is acquired by formulas of output voltage v and inductor current i. Without consideration of parasitics, state averaged voltage on the inductor in one switching cycle is $v_g-v(1-d)$, while the output current equals $(1-d)i$. Therefore, v and i are given by

$$\begin{cases} v = \frac{R}{1+sRC}(1-d)i \\ i = \frac{1}{sL}[v_g-v(1-d)] \end{cases}. \tag{5}$$

Differential function of (5) is given by

$$\begin{cases} \hat{v} = \frac{R}{1+sRC}[(1-d)\hat{i} - i\hat{d}] \\ sL\hat{i} = -(1-d)\hat{v} + v\hat{d} \end{cases}. \tag{6}$$

Eliminating $\hat{i}$, transfer function from duty cycle to output voltage is given by

$$G_{vd}(s) = \frac{\hat{v}}{\hat{d}} = \frac{v}{(1-d)R}\frac{-sL + (1-d)^2R}{s^2LC + sL/R + (1-d)^2}. \tag{7}$$

Furthermore, $H_{dv}(s)$ and $H_{di}(s)$ are acquired through (2) and (4). They are expressed in Laplace domain as