Fixed-Frequency Boundary Control of Buck Converter With Second-Order Switching Surface

Wai-To Yan, Member, IEEE, Carl Ngai-Man Ho, Member, IEEE, Henry Shu-Hung Chung, Senior Member, IEEE, and Keith T. K. Au

Abstract—This paper presents a fixed-frequency boundary control of buck converters. The method is based on integrating the concept of variable hysteresis into the boundary control technique with second-order switching surface. The switching frequency is maintained constant over a wide range of supply voltages and output loads. The method is based on using a frequency-to-voltage converter and comparing its output voltage with a reference voltage to control the width of the hysteresis in the boundary controller. It also combines the advantages of the boundary control that the converter can reach the steady state in two switching actions after large-signal disturbances. The basic operating principles, stability analysis, and design procedures will be given. The proposed control method has been successfully applied to control a 140 W, 24 V/12 V buck converter. The steady-state characteristics, including the switching frequency and output voltage ripple, at different input voltages and output loads with and without the proposed control method have been compared. The system responses under large-signal supply voltage and load disturbances will be discussed.

Index Terms—Boundary control, constant switching frequency, geometric control method, second-order switching surface, state trajectory control.

I. INTRODUCTION

BOUNDARY control techniques with linear switching surfaces, such as hysteresis control and sliding-mode control [1]–[10], or nonlinear switching surfaces [11]–[14] have been proposed to be alternatives to pulsewidth-modulated control strategies in dc/dc switching regulators. It addresses the complete operation of a converter and does not differentiate startup, transient, and steady-state periods [1], [2]. Based on studying the trajectory families on the state plane, a switching surface is defined to determine the states of the switches. An ideal switching surface can achieve global stability, good large-signal operation, and fast dynamics. The methods with linear switching surface are based on guiding the trajectory from any initial state on the state plane to the target operating point while the one with nonlinear/curved switching surface can be made close to the ideal switching surface. Instead of guiding the trajectory, the switching surface is derived by estimating the trajectory movement after a switching action. This results in a high state trajectory velocity along the switching surface and can accelerate the trajectory moving toward the target operating point within two switching actions after input and output disturbances. A comparative study of the boundary controls using first- and second-order switching surfaces has been discussed in [12], [13].

However, due to the dependence of the output voltage ripple on the relative magnitudes of the input and output voltages, and load power, the boundary control is largely operated at a variable switching frequency. This will make the design of the filter components difficult in some applications. Recently some digital techniques have extended the concept of [11]–[13] for digital control, such as the time-optimal control in [17], [18] and the capacitor charge balance technique in [19].

For reducing the structure complexity, an attempt of an analog-based fixed-frequency boundary control having the same fast response characteristics in [11]–[13] is proposed in this paper. In keeping the operating frequency constant in the boundary control, there are many techniques in the literature. They can be categorized into the hysteresis comparator methods [20], variable width hysteresis comparator methods [21]–[25], and injection of a fixed-frequency external signal methods [26]–[28]. Among them, the variable width hysteresis comparator method has been proved to be robust and reliable [23]. This paper presents a control method that integrates the concept of the variable width hysteresis in boundary control with second-order switching surface. Instead of using the phase-locked loop (PLL) in [23], the proposed method is based on a frequency-to-voltage conversion to deal with the limited lock-in range in PLL [24], [25]. It does not require external frequency generator, but a reference dc voltage for controlling the switching frequency. It extends the advantages of the boundary control with the second-order switching surface that the converter can reach the steady state in two switching actions after large-signal input or output disturbances. Thus, it combines the advantages of operating at a fixed switching frequency in the steady state and having fast response in the transient period.

The basic operating principles, system stability analysis, and design procedures will be given. The method has been successfully applied to a buck converter. The steady-state behaviors, including the switching frequency and output voltage ripples, and large-signal dynamics under input and output disturbances have been studied. Theoretical predictions are verified with experimental results of a 140 W, 24 V/12 V prototype.
II. OPERATING PRINCIPLES

Fig. 1 shows the system block diagram. It consists of four major parts, including the main power conversion stage (PCS), the second-order boundary controller (SBC) [11]–[13], the frequency-to-voltage converter (FVC), and the error amplifier (EA). Fig. 2 shows the implementation of the whole system. FVC firstly converts the gate signal $v_g$ for PCS into a dc voltage $v_{FVC}$, which will then be compared with a reference voltage $V_f,\text{ref}$ by EA. The output of EA, $\Delta$, is used to control the hysteresis band. SBC inside will generate upper and lower bands together with $\Delta$ to determine the switching times of the main switch $S$ in PCS. Thus, the function of SBC is used to regulate the output voltage and the earlier mentioned four parts form a feedback loop for regulating the switching frequency. Fig. 3 shows the block diagram of the small-signal frequency control loop. Any variation in the switching frequency $\delta f_s$ will give rise to a voltage variation in $\delta v_f$ by the FVC. Then, as the variation of the reference frequency (and hence $v_{f,\text{ref}}$) is zero, the error signal $\delta e$ to the EA is the same as $\delta v_f$. $\delta e$ will be amplified by EA to give a signal variation in the hysteresis band of $\delta \Delta$ to the PCS and SBC, so that the switching frequency variation is corrected. This completes the small-signal control loop for regulating the switching frequency.

In the following discussions, the operations of the SBC will be given and the switching frequency will be derived. The small-signal models of the four parts and thus the overall control loop in Fig. 3 will then be formulated.

A. Operation of SBC

The PCS has two operating modes, namely Mode 1 and Mode 2, in one switching cycle. The key time-domain waveforms are shown in Fig. 4. Assume that the components are ideal in the following derivations. The following state-space equations can be derived for each mode. In Mode 1 ($t_1 \leq t < t_3$), $S$ is on and $D$ is off.

$$\frac{di_L}{dt} = \frac{1}{L}[v_i(t) - v_C(t)] \tag{1}$$

$$\frac{dv_C}{dt} = \frac{1}{C}[i_L(t) - \frac{v_C(t)}{R}] \tag{2}$$

where $L$ and $C$ are the output filter inductor and capacitor, respectively, $v_i$ and $i_L$ are the input voltage and inductor current, respectively, $v_C$ is the capacitor voltage (output voltage) and $R$ is the load resistance.

Thus, the inductor current increases linearly in this mode.
Assume that $i_C$ varies linearly in Mode 1. Thus, by substituting (5) into (1)

$$\frac{di_L}{dt} = -\frac{1}{L} v_C(t) \quad \text{(3)}$$

$$\frac{dv_C}{dt} = \frac{1}{C} \left[ i_L(t) - v_C(t) \right] \quad \text{(4)}$$

Thus, the inductor current decreases linearly in this mode.

Two sets of state trajectories, namely Mode-1 and Mode-2 trajectories shown in Fig. 5, corresponding to each mode can be derived as follows. The values of $L$ and $C$ given in Table I are used to derive Fig. 5. The Mode-1 trajectories are drawn with solid lines while the Mode-2 trajectories are drawn with dotted lines.

As the load variation is small, $v_C$ is relatively constant over a switching cycle. Thus

$$i_L = i_C + i_o$$

$$\Rightarrow \frac{di_L}{dt} \approx \frac{di_C}{dt} \quad \text{(5)}$$

Hence, the capacitor current ripple is the same as the inductor current ripple. Equation (5) is commonly used to derive the input-to-output conversion characteristics. Such assumption is valid because the variation in the capacitor voltage is much smaller than the average capacitor voltage. Thus, the variation in the load current ($i_o$) is very small.

TABLE I COMPONENT VALUES OF THE PROTOTYPE

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$S$</td>
<td>IRFZ44V</td>
<td>$D$</td>
<td>MUR460</td>
</tr>
<tr>
<td>$L$</td>
<td>35\mu H</td>
<td>$C$</td>
<td>300\mu F</td>
</tr>
<tr>
<td>$K_{vc}$</td>
<td>15600</td>
<td>$R_c$</td>
<td>1/4167</td>
</tr>
<tr>
<td>$R_{lp}$</td>
<td>510k\Omega</td>
<td>$C_{lp}$</td>
<td>1F</td>
</tr>
<tr>
<td>$R_c$</td>
<td>510\Omega</td>
<td>$C_c$</td>
<td>2.2\mu F</td>
</tr>
</tbody>
</table>

In Mode 2 ($t_3 < t < t_5$), $S$ is Off and $D$ is On.

Thus, the inductor current decreases linearly in this mode.

In Mode 2 ($t_5 < t < t_3$), $S$ is Off and $D$ is On.

Thus, the inductor current decreases linearly in this mode.

In Mode 2 ($t_3 < t < t_5$), $S$ is Off and $D$ is On.

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while the latter one is derived by using the natural response of the converter. For the sake of simplicity in the practical implementation, the values of \( K_1 \) and \( K_2 \) are constant and are defined at the nominal values of \( v_i \) and \( v_o \). References [12] and [13] have done a comprehensive analysis on studying the effects of the converter behaviors with nonoptimal values of \( K_1 \) and \( K_2 \). The results show that the system dynamics will only take more than two switching actions before reaching the steady state.

### B. Switching Frequency and Output Voltage Ripple

By substituting (5) into (1) and (3), it can be shown that

\[
t_3 - t_1 = L \left[ \frac{i_C(t_2) - i_C(t_1)}{v_i - v_{ref}} \right] \quad (16)
\]

and

\[
t_5 - t_3 = -L \left[ \frac{i_C(t_3) - i_C(t_2)}{v_i - v_{ref}} \right]. \quad (17)
\]

Equations (12) and (15) will then give

\[
i_C(t_5) = i_C(t_1) = -\sqrt{\frac{2k\Delta}{K_1}} \quad (18)
\]

and

\[
i_C(t_3) = \sqrt{\frac{2(1-k)\Delta}{K_2}} \quad (19)
\]

where \( k = \frac{v_{ref}}{v_i} \).

Detailed proofs of (18) and (19) are given in the appendix. Thus, by substituting (18) and (19) into (16) and (17), the switching frequency \( f_S \) is equal to

\[
f_S = T_S^{-1} = (t_5 - t_1)^{-1} = H K \Delta^{0.5} \quad (20)
\]

where \( H = \frac{1}{L} \frac{v_{ref}(t_5 - t_1)}{v_i} \) and \( K = \sqrt{\frac{2K_1 K_2}{2(1-k)K_1}} \).

As illustrated in Fig. 4, the output voltage ripple equals \( 2\Delta \).

### C. Small-Signal Modeling

A small-signal stability analysis is used to study and design the feedback loop of the frequency control loop, because its feedback mechanism is considered as a simple linear control system. Based on (20), the small-signal frequency-to-ripple transfer function \( G_{SBC} \) of SBC is

\[
G_{SBC}(s) = \frac{\delta f_S(s)}{\delta \Delta(s)} = K_{SBC} = -\frac{1}{2} K_T H K \Delta^{-1.5} \quad (21)
\]

where \( \Delta \) is the steady-state value of the hysteresis band at the considered operating point. \( K_T \) is the total gain including the gain of the transducers.

The transfer function of SBC is assumed to be constant because the crossover frequency of the feedback control loop shown in Fig. 3 is much lower than the switching frequency. For the prototype that will be discussed in Section IV, the crossover frequency of the feedback control loop is chosen to be 17 Hz and the switching frequency is 25 kHz. Moreover, the dynamic response of SBC is very fast. The validity of the assumption has been confirmed by the experimental measurements and theoretical predictions in Section IV.

The transfer function \( G_{FVC} \) of the FVC can be considered as a constant \( K_{FVC} \) because the bandwidth is wide as compared with the switching frequency of the converter.

The transfer function \( G_{LP}(s) \) of the low-pass filter at the output of FVC is

\[
G_{LP}(s) = \frac{\delta v_f(s)}{\delta v_{FVC}(s)} = \frac{\omega_{LP}}{s + \omega_{LP}} \quad (22)
\]

where \( \omega_{LP} = \frac{1}{\tau_{LP} C_{LP}} \) is the cutoff frequency of the low-pass filter.

A noninverting EA shown in Fig. 2 is used. Its transfer function \( G_c \) is

\[
G_c(s) = \frac{\delta \Delta(s)}{\delta \varepsilon(s)} = -\frac{s + \omega_c}{s} \quad (23)
\]

where \( \omega_c = \frac{1}{\tau_c C_c} \) is the cutoff frequency of the EA.

Thus, as shown in Fig. 3, the loop gain \( T_{OL} \) of the frequency control loop is

\[
T_{OL}(s) = G_{SBC}(s) G_{FVC}(s) G_{LP}(s) G_c(s) = K_{SBC} K_{FVC} \frac{\omega_{LP}(s + \omega_c)}{s(s + \omega_{LP})} \quad (24)
\]

The stability can be ensured if \( |T_{OL}(j\omega_c)| < 1 \) and \( \angle T_{OL}(j\omega_c) \) is 180° at the designed crossover frequency \( \omega_c \). Since the introduced feedback controller is of integral error type, theoretically, there is no steady-state error between the reference switching frequency and actual switching frequency.

### III. SIMPLIFIED DESIGN PROCEDURES

The system is designed by considering the following specifications:

1. Minimum values of \( L \) and \( C \) in the power stage are designed by the formulas

\[
L > \frac{V_o}{\Delta I_L f_S} \left( 1 - \frac{V_o}{V_{in, max}} \right) \quad (25)
\]

and

\[
C > \frac{\Delta I_L}{8f_S \Delta V_o}. \quad (26)
\]

Detailed derivations of (25) and (26) can be found in [29].

2. Based on (7) and (9), the values of \( K_1 \) and \( K_2 \) are chosen to be

\[
K_1 = \frac{L}{2C [V_{in} - v_{ref}]} \quad (27)
\]

\[
K_2 = \frac{L}{2C v_{ref}} \quad (28)
\]

where \( V_{in} = \frac{1}{2} (V_{in, min} + V_{in, max}) \).

3. The values of \( H \) and \( K \) can be obtained readily from (20).
4) The values of $R_{LP}, C_{LP}, R_C,$ and $C_C$ are chosen so that $\omega_{LP} = 2 \omega_C = \alpha(2\pi f_S)$, where $\alpha \ll 1$. Thus, by using (22) and (23)

$$\frac{1}{R_{LP}C_{LP}} = 2\alpha\pi f_S$$

(29)

and

$$\frac{1}{R_C C_C} = \alpha\pi f_S.$$  

(30)

### IV. EXPERIMENTAL VERIFICATIONS

A 140 W buck converter has been designed and tested and the specifications are given as follows:

- **a)** input voltage, $v_i$: 20–30 V;
- **b)** output voltage, $v_o$: 12 V;
- **c)** maximum output voltage ripple, $2\Delta$: 160 mV;
- **d)** maximum inductor current ripple: 7 A;
- **e)** switching frequency: 25 kHz;

Based on the design procedures given in Section III, the values of the components used in the prototype are given in Table I. The output load is a resistor bank. The FVC used is TC9400 with the bandwidth up to 100 kHz. The circuit implementation of the system is shown in Fig. 6.

Figs. 7 and 8 show the waveforms of the output voltage and the gate signal when the load resistance is changed from 1.1 $\Omega$ (11 A, 132 W) to 4.8 $\Omega$ (2.5 A, 30 W), and vice versa, respectively. The transient periods last about 42 $\mu$s and 84 $\mu$s, respectively, and the converter settles in two switching actions. The steady-state switching period is kept at about 40 $\mu$s before and after the two load disturbances. Fig. 9(a) and (b) show the dynamic behaviors of the output $v_f$ in Fig. 2 and the hysteresis band $\Delta$ during the two load disturbances given in Figs. 7 and 8. The settling times in both cases are approximately equal to 1.95 ms.
Figs. 10 and 11 show the waveforms when the input voltage is suddenly changed from 20 V to 30 V, and vice versa, respectively. The transient periods last about 40 µs and 50 µs, respectively. Again, the converter settles in two switching actions and the steady-state switching period is also kept at about 40 µs before and after the two input disturbances. The input voltage is introduced with a high percentage of ripples. Apart from studying the dynamic response, it can be observed that the output voltage can be regulated tightly at the steady state without being affected by the input voltage ripple.

The steady-state characteristics of the converter with and without proposed control method have been studied. Without the proposed control method, Fig. 12(a) shows the experimental results of the switching frequency and output voltage ripple when the input voltage is 24 V. The output power is varied from 15 W to 140 W. The switching frequency has a percentage variation of 57%, varying between 14.5 kHz and 22.8 kHz. Moreover, when the output power ranges between 15 W and 40 W, the converter is operated in discontinuous conduction mode. The converter is then in continuous conduction mode (CCM) when the output power is above 40 W.

With the proposed control method, Fig. 13(a) shows the experimental results of the switching frequency and output voltage ripple when the input voltage is 24 V. With the same output load range, the switching frequency has a percentage change of about 3%, varying between 24.5 kHz and 25.3 kHz. The converter is operated in CCM over the operating range. With or without the proposed control method, when the output power increases, the output voltage ripple decreases because the duty cycle of the main switch is increased.

Without the proposed control method, Fig. 12(b) shows the experimental results of the switching frequency and output voltage ripple without the proposed control method when the output power is 70 W. When the input voltage varies between 20 V and
30 V, the switching frequency has a percentage change of about 20%, varying between 20.8 kHz and 24.9 kHz.

Fig. 13(b) shows the corresponding experimental results with the proposed control method. When the input voltage varies between 20 V and 30 V, the switching frequency has a percentage change of about 2.4%, varying between 24.6 kHz and 25.2 kHz. With or without the proposed control method, when the input voltage increases, the output voltage ripple increases because the duty cycle of the main switch is decreased. In the earlier mentioned measurements, although the switching frequency has slight deviations from the reference frequency of 25 kHz, there is no frequency jittering throughout the operations.

Fig. 14(a) and (b) show the theoretical and experimental gain and phase curves of the control loop shown in Fig. 3, (a) Gain curve. (b) Phase curve.
respectively. The experimental results are in close agreement with the theoretical results, which are obtained by using the values shown in Table I. The crossover frequency is at about 17 Hz and the phase margin is more than 30°. The loop gain characteristics show that the controller is in stable operation. In this prototype, a low crossover frequency is chosen to eliminate fast variation of the switching frequency if the supply or load disturbances are pulsating and repetitive. Experimental results show that such low crossover frequency does not affect the fast dynamic response of the SBC. Nevertheless, the bandwidth of the control loop can be increased by changing the time constant of the EA shown in Fig. 2.

For the sake of simplicity in the earlier analysis, the equivalent series resistance (ESR) of the output capacitor is neglected. As studied theoretically and experimentally in [12] and [13], the effects of the ESR do not affect the converter’s performance significantly with the boundary control using second-order switching surface. In the experimental prototype, two different sensors, including sensing resistor and Hall-effect current sensor, have been used to sense the capacitor current. Both sensing methods do not cause adverse effect on the converter’s performance.

V. CONCLUSION

An improved boundary control technique with second-order switching surface for buck converters has been presented. It exhibits two key features. First, the technique combines the advantage of SBC that the converter can reach the steady state in two switching actions after large-signal disturbances. Second, the switching frequency can be kept at a relatively constant value and the implementation of the frequency control loop only requires simple circuitry. A 140 W prototype has been tested. Overall there is good agreement between the theoretical predictions and experimental measurement results. As the proposed control is a combination of the hysteresis band regulation and nonlinear switching surface, it can be extended to other switching surfaces such as the switching surfaces proposed in [14] and [15]. Further work will be dedicated to study the converter operating in DCM.

APPENDIX

A. Proofs of (18) and (19)

As shown in (29)

\[ \Delta = \frac{v_{ref}(1 - k)}{16 LC f_S^2}. \]  

(A.1)

Based on (1) and (5), for \( t_1 \leq t < t_3 \)

\[ v_C(t) = v_{ref} - \Delta + \frac{v_i(1 - k)}{2LC} (t - t_2)^2. \]  

(A.2)

Thus

\[ t_1 - t_2 = t_2 - t_3 = -\frac{k}{2f_S}. \]  

(A.3)

By substituting (A.3) into (A.2)

\[ v_C(t_1) = v_C(t_3) = v_{ref} + \frac{v_{ref} (1 - k)}{8LC f_S^2} \left( -\frac{1}{2} \right) \]

\[ = v_{ref} + 2 \left( -\frac{1}{2} \right) \Delta. \]  

(A.4)

Equations (18) and (19) can be obtained by putting (A.4) into (12) and (15), respectively.

REFERENCES

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