Adaptive Digital Predictive Peak Current Control Algorithm for Buck Converters

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Abstract

Digital current control techniques are an attractive option for DC-DC converters. In this paper, a digital predictive peak current control algorithm is presented for buck converters that allows the inductor current to track the reference current in two switching cycles. This control algorithm predicts the inductor current in a future period by sampling the input voltage, output voltage and inductor current of the current period, which overcomes the problem of hardware periodic delay. Under the premise of ensuring the stability of the system, the response speed is greatly improved. A real-time parameter identification method is also proposed to obtain the precision coefficient of the control algorithm when the inductance is changed. The combination of the two algorithms achieves adaptive tracking of the peak inductor current. The performance of the proposed algorithms is verified using simulations and experimental results. In addition, its performance is compared with that of a conventional proportional-integral (PI) algorithm.

Key words: Adaptive control, Buck converter, Parameter identification, Peak current control, Predictive algorithm

I. INTRODUCTION

In the past three decades, the control of switch mode power supplies has been predominantly implemented using analog components [1]-[5]. However, with the emergence of powerful and low-cost digital signal processors (DSPs), digital control is becoming an increasingly attractive alternative to analog options [6]-[8]. Digital control offers the potential advantages of lower sensitivity to parameter variations, programmability and the possibility to improve performance using more advanced control schemes. As a result, microprocessors and DSP-based digital control have become widespread in the motor control and power electronic converter fields [9]-[12].

However, one of the major drawbacks of digital controllers is the inherent time delay of the control-loop structure. This delay results from both the zero-order-hold (ZOH) and the computational time delay required for analog-to-digital conversion (ADC), computations and pulse-width modulation (PWM) generation. In buck converters, all of these delays are usually equal to a complete sampling period, namely one switching cycle if it is equal to the switching period, since the duty ratio can only be updated at the beginning of each switching cycle. Such delays significantly degrade the control loop performance, which reduces the controller bandwidth.

In order to improve the performance of digital controllers, many predictive current control techniques have been proposed to compensate for time delays [13]-[18]. In [19], the authors put forward a predictive digital interpolation current control for DC-DC converters. This control is based on the inductor current trajectory and uses two sampled values of the current to generate the duty ratio in the same switching period without requiring knowledge of the input voltage, output voltage and inductance. However, the control precision of this method is not high. The method proposed in [20] considers disturbances of the inductor current but not those of the reference current. Therefore, when the reference current is disturbed, the inductor current cannot achieve stability in two switching cycles. In [21], the author put forward an improved current control algorithm. However, the requirements for the hardware devices are high, which is not conducive to real applications. An approximate predictive technique using linear extrapolation has been proposed for buck converters in [22]. However, the duty ratio can only be updated once in two switching cycles, which reduces the transient performance of the controller. Two predictive digital current control schemes were introduced in

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but they had drawbacks and limitations.

In this paper, a new adaptive predictive peak current control algorithm for buck converters is proposed. The basic concept of this algorithm is to calculate the value of the duty ratio based on measured samples of the current and voltage signals so that the inductor current can track the reference current within two switching cycles. The algorithm is derived from an analysis of the inductor current waveform of a buck converter operating in the continuous conduction mode (CCM). This algorithm is both simple and easy to implement on digital chips. The response speed is greatly improved when compared to a conventional PI control algorithm. In addition, a real-time parameter identification method is introduced to solve the problem of small-range fluctuations of the inductance during long-term operation. Simulations and experimental results demonstrate the performance of the proposed control algorithm.

II. CURRENT FEEDBACK LOOP CONTROL DESIGN

The proposed algorithm is developed using a basic buck converter. Therefore, the results are based on the buck topology. Fig. 1 shows the topology of a digital controller that employs an outer voltage loop and an inner current loop. The system uses an ADC to obtain digital values of the analog variables for processing in a DSP to generate the corresponding digital values of the reference current and duty ratio. In the last step, DPWM action is performed to control the state of the converter switch by means of a leading-edge modulation.

A. Small-Signal Model of the Buck Converter

The state-space averaging method proposed by Middlebrook and Cuk [24], and Slobodan and Middlebrook [25] in 1976 is used in this study. This method bridges the gap that was believed to exist between the state-space technique and the averaging technique for modeling power stages. In recent years, it has become a very popular modeling method for both switching converters and resonant circuits. It has been especially used for modeling power stages. A small-signal model of a buck converter based on this method is shown in Fig. 2.

Where \( V_{in} \) and \( D \) are the input voltage and duty ratio in the steady state. In addition, \( L, C \) and \( R \) are the inductance, capacitance and load, respectively. \( \dot{v}_{in}(s), \dot{v}(s), \dot{d}(s) \) and \( \dot{i}(s) \) are small-signal values of the corresponding variables.

The corresponding transfer functions are obtained as follows. The control-to-output voltage transfer function, the control-to-inductor current transfer function, and the output impedance transfer functions are defined as:

\[
G_a(s) = \frac{V_{in}}{LCs^2 + \frac{L}{R}s + 1} \quad (1)
\]

\[
Z_v(s) = \frac{R}{RCs + 1} \quad (2)
\]

Fig. 1. Digital controller for a DC-DC converter.

Fig. 2. Small-signal model of a buck converter.

Fig. 3. Buck topology in the CCM. (a) Switch is on. (b) Switch is off.

B. Proposed Current Control Algorithm

Assuming a continuous mode operation for the buck converter topology, it is possible to differentiate the two states that correspond to the position of the main switch, as shown in Fig. 3. Then the slope of the inductor current can be obtained for each operating state. Their equations are expressed as:

\[
S_a = \frac{V_{in} - V_s}{L} \quad (4)
\]

\[
S_d = \frac{V_s}{L} \quad (5)
\]
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In the actual hardware equipment, there is a certain delay between the sampling and the update of the duty ratio. Therefore, if the delay is ignored, the system loses stability.

In order to solve this hardware delay problem, the inductor current of the next cycle is predicted in advance. Thus, it can track the reference current at the end of the cycle. The restrictions for the sampling value are continuously relaxed and it is assumed that the input and output voltages remain constant in the two switching cycles.

\[
\begin{align*}
  v_{in}(t) &= v_{in(n-1)} \\
  v_{o}(t) &= v_{o(n-1)} (n-1 \leq t \leq n+1) \\
  i_{i}(t) &= i_{i(n-1)}
\end{align*}
\]  

The predicted peak inductor current can be deduced at the end of two switching cycles using the following expression:

\[
\begin{align*}
  i_{i(n+1)} &= i_{i(n)} - S_{d(n-1)} \cdot d_{n-1} \cdot T_s + S_{u(n-1)} \cdot d_{n-1} \cdot T_s \\
  & \quad - S_{d(n)} \cdot d_{n} \cdot T_s + S_{u(n)} \cdot d_{n} \cdot T_s
\end{align*}
\]

Substituting (9) into (4) and (5) results in:

\[
\begin{align*}
  S_{d(n-1)} &= S_{d(n)} = \frac{v_{in(n)}}{L} \\
  S_{u(n-1)} &= S_{u(n)} = \frac{v_{in(n)} - v_{o(n)}}{L}
\end{align*}
\]

By substituting (11), (12), (9), and (6) in (10), the predictive peak current equation is obtained as:

\[
\begin{align*}
  i_{i(n-1)} &= i_{i(n-1)} - \frac{v_{in(n)}}{L} (1-d_{n-1}) T_s + \frac{v_{in(n)} - v_{o(n)}}{L} d_{n-1} T_s \\
  & \quad - \frac{v_{o(n)}}{L} (1-d_{n}) T_s + \frac{v_{in(n)} - v_{o(n)}}{L} d_{n} T_s
\end{align*}
\]

By simplifying (13), the expression of the discrete-time domain of the duty ratio is derived as:

\[
\begin{align*}
  d_n &= \frac{L}{v_{in(n)} T_s} (i_{i(n-1)} - i_{i(n-1)} + \frac{2v_{in(n)}}{v_{in(n)}} - d_{n-1}
\end{align*}
\]

The difference in (14) is transformed into the z-domain which yields:

\[
\begin{align*}
  d_n (1+z^{-1}) &= \frac{L}{v_{in(n)} T_s} z^{-1} (i_{i(n)} - i_{i(n)}) + \frac{2v_{in(n)}}{v_{in(n)}} \\
  d_n &= \frac{L}{v_{in(n)} T_s} (z+1)^{-1} (i_{i(n)} - i_{i(n)}) + \frac{2v_{in(n)}}{v_{in(n)}} (1+z^{-1})^{-1}
\end{align*}
\]

To perform a small-signal analysis, it is necessary to add the disturbance to the steady state and then assume:

\[
\begin{align*}
  \left\{ \begin{array}{l}
    d_{n} = D + \tilde{d}_{n} \\
    v_{in(n)} = V_{in(n)} + \tilde{v}_{in(n)} \\
    v_{o(n)} = V_{o(n)} + \tilde{v}_{o(n)} \\
    i_{i(n)} = I_{i(n)} + \tilde{i}_{i(n)} \\
    i_{d(n)} = I_{d(n)} + \tilde{i}_{d(n)}
  \end{array} \right.
\]
where the variables with the hat symbols represent the small-signal disturbance values and the variables in capital letters represent the steady-state values of the corresponding variables. Substituting (17) into (16), yields:

\[
D + \hat{d}_s = \frac{L}{V_{in(n)}}(z+1)^{-1}(I_{in(n)} + \hat{i}_{in(n)}) - (I_{in(n)} + \hat{i}_{in(n)}) + 2(V_{in(n)} + \hat{v}_{in(n)})/(1 + z^{-1})^{-1}
\]

(18)

At this point, the small-signal approximation is performed, wherein the disturbance from the steady-state values is negligible when compared to the steady-state values.

Then by using approximations, it is possible to neglect all of the nonlinear terms, such as the second-order terms in (18), and reobtain a linear system, which includes the steady-state duty ratio modulation \(D\). After separating the steady-state (DC) and dynamic disturbance (AC) parts of (18), the following results are arrived at for the proposed algorithm.

The steady-state (DC) duty ratio equation is:

\[
D = \frac{L}{V_{in(n)}T_s}(z+1)^{-1}(I_{in(n)} - I_{in(n)}) + 2V_{in(n)}/(1 + z^{-1})^{-1}
\]

(20)

The dynamic disturbance (AC) small-signal duty ratio equation is:

\[
\hat{d}_s = \frac{L}{V_{in(n)}T_s}(z+1)^{-1}(\hat{i}_{in(n)} - \hat{i}_{in(n)}) + 2V_{in(n)}/(1 + z^{-1})^{-1}\hat{v}_{in(n)} - \frac{D}{V_{in(n)}}\hat{v}_{in(n)}
\]

(21)

Replacing the coefficient of the perturbation variable yields:

\[
\hat{d}_s = A(z) \cdot (\hat{i}_{in(n)} - \hat{i}_{in(n)}) + B(z) \cdot \hat{v}_{in(n)} + C_0 \cdot \hat{v}_{in(n)}
\]

(22)

where:

\[
A(z) = \frac{L}{V_{in(n)}T_s}(z+1)^{-1}
\]

\[
B(z) = \frac{2}{V_{in(n)}}/(1 + z^{-1})^{-1}
\]

\[
C_0 = -\frac{D}{V_{in(n)}}
\]

Using (22), a block diagram of the current loop is obtained as shown in Fig. 5.

In order to facilitate the analysis, the s-domain transfer function is derived by using a bilinear transformation method. First, the inner feedback voltage loop is obtained:

\[
T_r = \frac{G_{d}(s)}{1 - G_{d}(s) \cdot B(s)}
\]

(24)

Second, the reference current-to-inductor current and the input voltage-to-inductor current transfer functions are deduced.

\[
\frac{\hat{i}_e(s)}{\hat{v}_a(s)} \bigg|_{v_a=0} = T_r = \frac{G_{d}(s) \cdot A(s)}{1 - G_{d}(s) \cdot B(s) + G_{d}(s) \cdot A(s)}
\]

(25)

\[
\frac{\hat{i}_e(s)}{\hat{v}_a(s)} \bigg|_{v_a=0} = \frac{G_{d}(s) \cdot C_{\circ}}{1 - G_{d}(s) \cdot B(s) + G_{d}(s) \cdot A(s)}
\]

(26)

C. Stability Analysis

The stability properties of the predictive peak current control under leading edge modulation can be determined by examining the waveforms in Fig. 4 and using (14). There is an inductor current with a disturbance of \(\Delta i\) or a reference current with a disturbance of \(\Delta i_e\) at the beginning of the \((n-1)\)th switching cycle. The duty ratio \(d_{n-1}\) cannot be updated until the next switching cycle due to the periodic delay of the hardware system. Therefore, the duty ratio \(d_{n-1}\) remains the steady duty ratio \(D_{st}\) prior to the disturbance.

When the circuit is in the steady state, the following equation applies:

\[
D_{st} = \frac{v}{v_{in}}
\]

(27)

Therefore, (14) can be simplified to (28).

\[
d_n = \frac{L}{v_{in(n)}T_s} (i_{s(n-1)} - i_{s(n-1)}) + D_{st}
\]

(28)

Equation (29) is equivalent to (28).

\[
d_n = k \cdot \Delta i + D_{st} \quad \text{or} \quad d_n = k \cdot \Delta i_e + D_{st}
\]

(29)

where:

\[
\Delta i = i_{s(n-1)} - i_{s(n-1)} \quad \text{or} \quad \Delta i_e = i_{e(n-1)} - i_{e(n-1)}
\]

\[
k = \frac{L}{v_{in(n)}T_s}
\]

The duty ratio \(d_n\) is a linear function of the disturbance and it is directly proportional to the disturbance of the reference current and negatively proportional to the disturbance of the inductor current. After the duty ratio \(d_{n-1}\) has been computed using (28), it is necessary to eliminate the current disturbance and to ensure that the peak inductor current reaches the reference current by the end of the \((n)\)th switching cycle.
Finally, the circuit returns to the steady state. The duty ratio $d_{n+1}$ is derived using (28).

$$d_{n+1} = \frac{L}{v_{in(n+1)}}(i_s(n) - i_s(n)) + 2\frac{v_{in(n+1)}}{v_{in(n+1)}} - d_s$$  \hspace{1cm} (30)

Since the duty cycle update has a periodic delay, it is defined as:

$$\begin{cases} i_s(n-1) = i_s(n) \\ i_s(n-1) = i_s(n) \end{cases}$$  \hspace{1cm} (31)

Using (28), the changes in the input and output voltages during two adjacent switching cycles are ignored. Therefore, the duty ratio $d_{n+1}$ is obtained as follows:

$$d_{n+1} = D_s$$  \hspace{1cm} (32)

It can be seen from the theoretical analysis that after the adjustment of the two switching cycles, the inductor current makes the duty ratio $d_{n+1}$ equal to the duty ratio of the steady state. Therefore, the circuit reaches the steady state again.

**D. Small-Signal Analysis of Current Loop**

A buck converter, as shown in the schematic in Fig. 1, is built to verify the control strategy. The specifications of the converter are shown in Table I. The sampling frequency of the ADC is equal to the switching frequency.

In order to evaluate the predictive current controller performance, Bode diagrams of the predictive current control algorithm are drawn and shown in Fig. 6 and Fig. 7.

The Bode diagrams are similar for the open-loop frequency responses, i.e., $G_{ii}$ open-loop and $G_{iv}$ open-loop. In fact, $G_{ii}$ open-loop is equivalent to adding a negative proportional constant to the $G_{iv}$ open-loop, which makes the input voltage to produce a larger amplitude attenuation and a larger phase lag. The periodic delay of the hardware devices is equivalent to adding an open-loop zero-point to the right half of the $s$-plane, which changes the digital control into a non-minimum phase system and reduces the stability of the system. However, the proposed predictive control algorithm is used to compensate for the phase delay caused by the open-loop zero-point, which improves the stability of the system. The cut-off frequency of the amplitude frequency response caused by the reference current disturbance is 1.65 kHz, which is consistent with engineering requirements ($1/20f_c<f<1/5f_s$). This ensures the response speed of the system and inhibits high-frequency interference such as the switching frequency. In addition, the cut-off frequency of the amplitude frequency response caused by the input voltage disturbance is 27 Hz. This reduces the high-frequency interference from the input voltage.

In the closed-loop frequency responses, all of the roots of the closed-loop characteristic equation are distributed in the left half of the $s$-plane. The bandwidth of the $G_{ii}$ closed-loop is 2.2 kHz, which ensures the dynamic response speed of the system. However, the amplitude frequency response of the $G_{iv}$ closed-loop is a negative value, which indicates that the disturbance caused by the input voltage is inhibited after passing the current loop. Therefore, the input voltage disturbance is greatly attenuated by the action of the current loop. The step response of the $G_{iv}$ closed-loop is shown in Fig. 8 and the input signal is tracked at 100us (in two switching cycles). The

<table>
<thead>
<tr>
<th>TABLE I</th>
<th>BUCK CONVERTER PARAMETERS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage ($v_{in}$)</td>
<td>60 V</td>
</tr>
<tr>
<td>Output voltage ($v_{o}$)</td>
<td>20–30 V</td>
</tr>
<tr>
<td>Inductance ($L$)</td>
<td>100 μH</td>
</tr>
<tr>
<td>Capacitance ($C$)</td>
<td>480 μF</td>
</tr>
<tr>
<td>Load ($R$)</td>
<td>3 Ω</td>
</tr>
<tr>
<td>Switching frequency ($f_s$)</td>
<td>20 kHz</td>
</tr>
</tbody>
</table>
The step response of the $G_{ii\_closed-loop}$ is shown in Fig. 9 and the input disturbance is suppressed.

Fig. 10 shows an experimental platform for testing the closed-loop frequency response of the buck circuit, and Fig. 11 is the experimental schematic circuit for measuring the closed-loop frequency response.

As can be seen from Fig. 11, the 87521A T/R test set module of an Agilent 4395A network analyzer injects a sinusoidal disturbance signal with a fixed amplitude and a continuously varying frequency into the control loop. The disturbance signal is superimposed with the sampled signal of the inductor current. Then it is passed through the digital controller and the power-stage circuit. Finally, it is fed back to the $B$ port of network analyzer.

The parameters of the disturbance signal are as follows:
- Start frequency to end frequency: 10 Hz to 10 kHz.
- Scanning bandwidth: 10 Hz.
- Signal amplitude: 224 mV.
- Power source: 0 dB.

The experimental test results are consistent with the theoretical proof of the stability analysis in part C. The amplitude-frequency response for the bandwidth of the system is about 2 kHz, which is basically consistent with the software simulation waveform in section III of this paper.

**E. Parameter Identification**

During long-term operation of the system, the inductance fluctuates due to increases in temperature, aging of the magnetic core, etc. Therefore, the coefficient of the control algorithm has to be changed in real time. Otherwise, the tracking performance of the inductor current declines and the system becomes unstable. Thus, a simple and practical real-time parameter identification method is proposed.

Replacing the reference current $i_{c(n+1)}$ with the sampling current $i_{d(n+1)}$ in (8) yields:

$$i_{s(n+1)} = i_{s(n)} - S_{d(n)} \cdot d_s T_s + S_{a(n)} \cdot d_a T_s$$

(33)

Substituting (11) and (12) into (33) yields:

$$i_{s(n+1)} = i_{s(n)} - \frac{V_{o(n)}}{L} \cdot (1 - d_s) Y_c + \frac{V_{o(n)} - V_{a(n)}}{L} \cdot d_a T_s$$

(34)

The formula is simplified as follows:

$$k_0 = \frac{V_{a(n)} \cdot d_a - V_{o(n)}}{i_{d(n+1)} - i_{s(n)}}$$

(35)

where, $k_0 = \frac{L}{Tr}$

By identifying the parameter $k_0$, the error caused by the direct identification of a very small inductance is solved.
Then the parameter \( k_0 \) is directly substituted into (23). Thus, the new coefficient \( A(z) \) of the control algorithm is immediately updated.

When the system is in the steady state, the numerator and the denominator both approach zero and it is not possible to determine the parameter \( k_0 \) at this time. The inductor current and duty ratio can only be changed when the system is disturbed and only the parameter \( k_0 \) can be solved at this time.

\[
\text{Abs}(i_{s(n+1)} - i_{s(n)}) \geq \varepsilon
\]  

(36)

where the parameter \( \varepsilon \) is a threshold designed by the user.

In the simulations and experiments in this study, the threshold of the current difference is set to 1 A. If an increase in the accuracy of the parameter \( k_0 \) identification is desired, the average of multiple inductor currents obtained from multiple switching cycles can be used.

### III. SOFTWARE SIMULATION RESULTS

A computer simulation of the predictive peak current algorithm-controlled buck converter operating in the CCM is performed with the circuit parameters shown in Table I. A predictive control algorithm (10) and a non-predictive control algorithm (8) are used to obtain inductor current waveforms as shown in Fig. 12.

When the delay of the digital control system is not taken into consideration, the inductor current (based on the non-predictive algorithm) produces oscillations. Therefore, in order to realize stable operation of the inductor current, the inductor current must be predicted one switching cycle in advance.

It can be seen from Fig. 13 and Fig. 14 that the reference current can be tracked completely for two switching cycles, regardless of whether the reference current increases or decreases. In fact, the delay of the first cycle is caused by hardware device sampling, algorithm operation, etc. The delay cannot be eliminated and can only be minimized. Therefore, the algorithm achieves tracking in only two switching cycles.

In Fig. 15, a positive inductor current disturbance with an amplitude of 3 A is superimposed on the inductor current at 0.005 seconds. After two switching cycles, namely at 0.0051 seconds, the inductor current achieves tracking of the reference current again. Thus, the system stabilizes at this time. Similarly, a negative inductor current disturbance with an amplitude of -3 A is added to the inductor current at 0.0052 seconds. Subsequently, the reference current is tracked in the second switching cycle after a delay of one switching cycle. Eventually, the system returns to the stable state.

The parameter \( k_0 \) can be identified only when the inductor current disturbance is greater than the threshold. The larger the inductor current disturbance, the more accurate the identification result becomes. Fig. 16 shows the identification result of an inductor current disturbance of 3 A. The parameter \( k_0 \) has a value of 2 in the stable state. This value is substituted into (23). Thus, the control coefficient is updated.

In Fig. 17, the reference current decreases from 15 A to 12 A at 0.017 seconds. After two switching cycles, it is tracked by the inductor current at 0.0171 seconds. At this time, the inductor current disturbance satisfies the threshold of the parameter \( k_0 \) identification, and the result for the parameter \( k_0 \)
is 2 \( (L = 100 \mu H, T_s = 50 \mu s) \). At 0.0172 seconds, an increase in the inductance \( (L = 150 \mu H) \) causes a decrease in the inductor current. However, the inductor current disturbance does not reach the threshold of the parameter \( k_0 \) identification at this time. Therefore, the control coefficient cannot be updated in real time and the reference current cannot be tracked within two switching cycles. Thus, the controller performance decreases.

The reference current increases from 12 A to 15 A at 0.0175 seconds. Because the control coefficient has not been updated, a current tracking error occurs at 0.0176 seconds. However, the inductor current disturbance satisfies the threshold of the parameter \( k_0 \) identification at this time, and the identification result is 3 \( (L = 150 \mu H, T_s = 50 \mu s) \). The control coefficient is updated at this time. Therefore, the reference current can be accurately tracked after two switching cycles at 0.0177 seconds. The predictive peak current control with self-adaptive parameters has been achieved.

**IV. EXPERIMENTAL RESULTS**

The performance of the proposed predictive peak current control scheme has also been experimentally investigated. The scheme was implemented on a TMS320F28335 TI DSP chip, which has an on-board 12-bit ADC and a DPWM with a 16-bit counter. The parameters of the buck converter are listed in Table I.

In Figs. 18, 19 and 20, inductor current and output voltage tracking waveforms of the conventional PI control method are shown as the reference current increases. It is assumed that the integral parameter \( k_i \) is unchanged and that the proportional parameter \( k_p \) is gradually increasing.

When the parameter \( k_p \) is 0.001, the stability of the inductor current waveform is optimal. However, the adjustment time for the inductor current and output voltage is longer. In Fig. 19, the parameter \( k_p \) is 0.002. In the case of a mutation of the reference current, the inductor current produces oscillations, which decreases the stability of the system. This occurs even though the adjustment time of the inductor current and the output voltage have been reduced when compared to the case shown in Fig. 18. In Fig. 20, the parameter \( k_p \) is increased to 0.003. In this case, the inductor current causes larger oscillations than in the case shown in Fig. 19. Although the adjustment time has been further shortened, the inductor current does not reach its stable state.

Therefore, it is difficult for a conventional PI control method to simultaneously take a rapid transient response and stability into account. In addition, the selection of the parameters requires previous engineering experience for continuous testing.

Fig. 21 and Fig. 22 show the transient responses of the inductor current waveform using the proposed predictive control when the reference current increases and decreases, respectively. The transient response of the inductor current has been greatly improved over that of the conventional PI control method without weakening its stability.

Fig. 23 and Fig. 24 are obtained by amplifying the waveforms 40 times. It can be seen that the reference current...
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is accurately tracked by the inductor current in two switching cycles, i.e., 100 μs. Most importantly, the inductor current does not oscillate.

Fig. 25 and Fig. 26 show transient response experimental waveforms of the inductor current when the inductor current increases. It can be seen that the inductor current returns to the stable state of 15 A after two switching cycles regardless of whether the inductor current increases or decreases.

Fig. 25. Transient tracking experimental waveform of the inductor current when the inductor current increases (from 15 A to 17 A).
Fig. 26. Transient tracking experimental waveform of the inductor current when the inductor current decreases (from 15 A to 13 A).

Fig. 27. Identification of the experimental waveform of the parameter $k_0$ at the time of an inductor current disturbance.

Fig. 28. Partial magnification (20,000 times) of area 1.

Fig. 29. Partial magnification (20,000 times) of area 2.

V. ADVANTAGES AND DISADVANTAGES

Under the premise of circuit stability, the predictive current control method proposed in this paper achieves rapid tracking of the inductor current within two switching cycles, which greatly improves the transient response speed of the current. However, this current control method also has some disadvantages.

1. The controller parameters are sensitive to drifts of the inductance. If the inductance cannot be accurately identified, the transient tracking performance of the inductor current is degraded.

2. When the current disturbance is large, the duty ratio of the adjacent two switching cycles changes greatly, which is likely to result in damage to the switching device.

3. Since data acquisition, processing and calculations are required for each switching cycle, the hardware requirements for the ADC and the controller are high when the switching frequency is high.

VI. CONCLUSIONS

In this paper, a predictive peak current control algorithm for the digital operation of buck converters has been presented.

One of the major limitations of digital control is its limited performance due to computational time delay and sampling delay. In this study, it has been shown that the proposed control scheme overcomes the delay problem and improves the stability of the system. By predicting the inductor current when the current is disturbed, the reference current is tracked...
in only one switching cycle after the delay period ends.

The design procedure of the controller is simple and load independent. Furthermore, a simple and practical method for the inductance identification is proposed to solve the problem of decreased control performance when the inductance changes. Experimental results are obtained and compared with results obtained using a conventional PI control strategy. Under the premise of system stability, the response speed of the inductor current is improved.

Although this paper focused on a buck converter application, similar concepts can be extended to other converter topologies.

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