Digitally Implemented Average Current Mode Control in Discontinuous-Conduction-Mode PFC Rectifier

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Abstract — This paper proposes a digital average current mode control method in discontinuous conduction mode (DCM) power factor correction (PFC) rectifier. The proposed control technique does not estimate but directly senses the average value of the inductor current in each switching cycle. It is implemented by means of a conventional current sensing circuit and a microcontroller. The calculation burden of the microcontroller is the same with that of conventional two-loop controlled converter, because the additional calculation process is not required. The control method achieves lower total harmonic distortion (THD) and higher power factor (PF) than conventional one. Experimental results with 200-W prototype hardware verify the feasibility and performance of the proposed control method.

Keywords — Average current mode control, digital control, discontinuous conduction mode (DCM), power factor correction (PFC), variable-duty-cycle control.

I. INTRODUCTION

Harmonic current by various non-resistive electronic loads is known as the cause of problems such as power loss, noise, voltage distortion and reduced line utilization [1]. Especially, regulations on line current distortion in electronic devices pose a challenge for low power converters [2]. Active PFC units based on switch mode power supplies are considered as a general solution to minimize the harmonic current. They feature smaller size, higher efficiency and PF than passive PFC circuit [3].

DCM operation of active PFC rectifiers features low diode recovery loss, zero-current turn-on of main switch, and constant switching frequency. It is usually adopted in low to medium power range, i.e., under 200~300W, where the conduction losses in semiconductor devices are not dominant. As its conventional control law, open-loop control or constant duty cycle control is widely used due to its simplicity in the control scheme. The controller needs only a low-bandwidth voltage loop which dispenses with current sensing and control circuit [4]. The drawbacks of the open-loop control technique, however, are relatively low PF and high THD, especially when high line input voltage is applied [5]. Various researches have been done on the variable duty cycle control to overcome the aforementioned problems of open-loop control. One-cycle control method enables the variable duty cycle control with two resettable integrators and resistive sensing network of rectified input voltage [6]. Harmonic injection methods to achieve near-unity PF have been also reported though their analog control circuits are complicated [7]-[8]. Another variable frequency control scheme is implemented with input voltage feedforward [9]. Reference [10] digitally realized the control scheme in [9] by using multiplying and square-root operations.

On the other hand, sensing the average inductor current in DCM is not straightforward in digital control implementation. Generally, digital controllers employ one-sample-per-switching-cycle to sense the analog information. The sampling instant of the average inductor current depends not only on the on-time of switch as in continuous conduction mode (CCM) [11] but also on the on-time of diode. Sample correction method in CCM rectifier, which corrects the sampled inductor current by multiplying correction factor, has been proposed to minimize distortion at both ends of the half line period where the inductor current is temporarily in DCM [12].

An average current mode control in digitally controlled DCM PFC rectifier is suggested in this paper. The control method uses conventional sensing circuit and fixed sampling instant to achieve the variable duty cycle control. It requires same calculation burden with conventional CCM PFC rectifier and does not need any additional computation. The proposed control technique also features low THD and high PF.

The paper consists of following sections: In Section II, the mode analysis is given and the mathematical expression of properly controlled inductor current shape is derived. Section III illustrates experimental results with 200W prototype hardware such as oscillograms and measured THD data. The paper is summarized and concluded in Section IV.

Figure 1. Proposed current control circuit with boost rectifier.
II. PROPOSED CONTROL TECHNIQUE

A. Operational Mode Analysis

A circuit diagram of a boost rectifier with proposed control technique is shown in Fig. 1. Rectified input and output voltage sensing networks and gate driver are omitted for clarity. The rectifier operates in DCM with fixed frequency and fixed sampling instant. The current sensing circuit which consists of current sensing transformer \( T \), zener diode \( D_z \), sensing capacitor \( C_S \), and reset switch \( Q_1 \) is the same as that in conventional control [13]-[14]. The digital controller controls the gate signals of the two switches \( Q_1 \) and \( Q_2 \).

The operation of the current sensing circuit is different from the conventional one in two ways: First, the gate signal of \( Q_2 \) is not complementary to that of the main switch \( Q_1 \). The turn-on instant of \( Q_2 \) is delayed until the inductor current becomes zero thus the sensing capacitor voltage \( v_{C_S} \) remains constant. After \( v_{C_S} \) is sampled by the digital controller, \( Q_2 \) turns on to discharge \( C_S \). Second, the turn-off of \( Q_1 \) is controlled by a current loop in the average current mode control.

The operational waveforms are illustrated in Fig. 2. \( v_{gs1} \), \( i_L \), \( v_{C_s} \) and \( v_{gs2} \) are gate signal of \( Q_1 \), inductor current, sensing capacitor voltage, and gate signal of \( Q_2 \) respectively. Unit switching period is separated in three modes to explain the operation of the proposed current control. Time segments \( d_1T_S \) and \( d_2T_S \) represent the on-time of \( Q_1 \) and output diode \( D \) respectively. To simplify the analysis, semiconductor devices are assumed to be ideal. Rectified input voltage \( |v_{in}| \) and output voltage \( v_O \) are also approximated to be constant within a switching period.

- **Mode 1** \([t_0-t_1]\): At \( t_0 \), \( Q_1 \) turns on by the fixed frequency signal of the digital controller. \( i_L \) linearly increases with the slope \( \frac{|v_{in}|}{L} \), where \( L \) is inductance of boost inductor. Then the current sensing transformer \( T \) outputs the scaled inductor current through \( D_2 \) to \( C_S \). \( v_{C_S}(t) \) is shown in (1).

\[
v_{C_S}(t) = \frac{1}{nC_S} \int_{t_0}^{t} \frac{i_L(t)}{T} \, dt \tag{1}
\]

\( n \) represents the turn number of secondary winding of \( T \). \( v_{C_S}(t) \) increases in parabolic way from zero because the rectifier operates in DCM.

- **Mode 2** \([t_1-t_2]\): When \( Q_1 \) turns off at \( t_1 \) by the current loop, \( i_L \) decreases with the slope \( \frac{v_O - |v_{in}|}{L} \), \( v_{C_S}(t) \) still increases but the sign of the parabolic curve is inverted.

- **Mode 3** \([t_2-t_3]\): After \( t_2 \), \( i_L \) maintains zero and \( v_{C_S}(t) \) remains constant as in (2):

\[
v_{C_S}(t) = \frac{T_S}{nC_S} \left( \frac{1}{T_S} \int_{t_0}^{t} i_L(t) \, dt \right) = \frac{T_S}{nC_S} \left( \frac{1}{T_S} \int_{t_0}^{t} i_{avg}(t) \, dt \right)
\]

\[
= \frac{T_S}{nC_S} \times i_{avg}, \tag{2}
\]

where \( T_S \) is the switching period and the term in the braces is the average inductor current in a switching cycle or \( i_{avg} \). In this interval, \( v_{C_S}(t) \) represents the average inductor current because \( i_L \) in Mode 1 and Mode 2 is fully integrated. The digital controller then triggers the internal ADC to sample the \( v_{C_S}(t) \) to compare it with the inner loop current reference. After the sampling, \( Q_2 \) turns on.
to discharge \( C_i \) before the next turn-on signal of \( Q_1 \) occurs. Because the sampling instant is fixed in unit switching cycle, the digital controller does not spend any resource to determine when the sampling should occur in each switching cycle.

### B. Inductor Current Shaping

Generally, the inductor current waveform of average current mode controlled CCM boost rectifier resembles the sinusoidal of the input voltage to achieve high PF and low THD. However, unlike in CCM rectifier, the envelope of the inductor current in DCM rectifier does not look like a sinusoid. Fig. 3 compares the current shapes in general CCM and DCM PFC rectifier over half line period. The envelope of the inductor current of DCM PFC, \( i_{\text{avg}}(t) \), does not follow the sine wave though its average value per switching cycle, \( i_{\text{avg}}(t) \), tracks sine wave as in Fig. 3(b).

If the average inductor current per switching cycle is controlled to be proportional to the sinusoidal input voltage, \( i_{\text{avg}}(t) \) can be derived from the duty ratios \( d_1 \) and \( d_2 \) as in (3) and (4):

\[
d_1 = \frac{L_i(t)}{v_{in}(t)T_S}, \quad d_2 = \frac{L_i(t)}{(v_O - v_{in}(t))T_S}. \tag{3}, (4)
\]

\( i_{\text{avg}}(t) \) is derived from the geometry of the waveform shown in Fig. 2 with unknown constant \( X \) as shown in (5).

\[
i_{\text{avg}}(t) = \frac{1}{2}(d_1 + d_2) I_{pk} = X \sin \omega t \tag{5}
\]

\( v_{in}(t) \) is assumed to be the pure sine wave with peak value \( V_{pk} \) and period \( T_L = \frac{2\pi}{\omega L} \) which is expressed in (6).

\[
v_{in}(t) = V_{pk} \sin \omega t \tag{6}
\]

Substituting (3), (4) and (6) into (5) yields (7).

\[
i_{pk}(t) = 2 T_L \frac{P}{\eta L} \sqrt{1 - \frac{V_{pk} \sin \omega t}{v_O} \sin \omega t \frac{v_O}{\sin \omega t}} \tag{7}
\]

The unknown constant \( X \) can be derived from input power consideration. If \( P_{\text{in}}, P_0 \) and \( \eta \) are the input power, output power, and efficiency of the rectifier system, i.e., input filter and rectifier, (8) can be achieved.

\[
P_{\text{in}} = \frac{P_0}{\eta} = I_{rms} V_{rms} = \frac{1}{2} V_{pk} X \tag{8}
\]

\( I_{rms} \) and \( V_{rms} \) in (8) represent rms value of the input current \( i_{\text{avg}} \) and the input voltage \( v_{in} \) respectively. Rearranging and substituting (8) into (7) finally yields (9).

\[
i_{pk}(t) = 2 \frac{T_L P_0}{\eta L} \sqrt{1 - \frac{V_{pk} \sin \omega t}{v_O} \sin \omega t \frac{v_O}{\sin \omega t}} \tag{9}
\]

Fig. 4 illustrates the inductor current envelopes for various line input voltages and 200W load. The current envelope shows a “saddle” in the middle of the half line period. It is because the term in the second square root in (9) is not a constant but a function of \( \sin \omega t \). The performance improvement between the proposed and the conventional current control techniques is emphasized when high line voltage is applied because the saddle is deeper at higher line voltages.

### III. EXPERIMENTAL RESULTS

The proposed current control technique is verified by 200W laboratory prototype circuit. The circuit parameters are summarized in Table I.

The line input voltage to the prototype circuit is 60Hz 230VAC and the output voltage is controlled to be 400V/DC.

Fig. 5 compares current waveforms between conventional open-loop control and the proposed variable duty cycle control at full load with same power stage. The line input current \( i_{\text{avg}} \) is the filtered version of \( i_0 \) by simple second-order L-C filter as shown in Fig. 1. In Fig. 5(a),

![Figure 4: Inductor current envelopes for various line voltages and 200W load in average current mode DCM rectifier.](image)

<p>| TABLE I. CIRCUIT PARAMETERS OF LABORATORY PROTOTYPE |
|-------------------------------------|-------------|</p>
<table>
<thead>
<tr>
<th><strong>Device</strong></th>
<th><strong>Parameter</strong></th>
</tr>
</thead>
<tbody>
<tr>
<td>( L )</td>
<td>70( \mu )H</td>
</tr>
<tr>
<td>( C )</td>
<td>220( \mu )F/450V electrolytic</td>
</tr>
<tr>
<td>( Q_i )</td>
<td>STW20NM50</td>
</tr>
<tr>
<td>( D )</td>
<td>FSF10A60</td>
</tr>
<tr>
<td><strong>Current Sensing Circuit</strong></td>
<td></td>
</tr>
<tr>
<td>( T )</td>
<td>Turns ratio 1:50</td>
</tr>
<tr>
<td>( D_2 )</td>
<td>1N4744</td>
</tr>
<tr>
<td>( C_3 )</td>
<td>220( \mu )F/50V ceramic</td>
</tr>
<tr>
<td><strong>Digital Controller</strong></td>
<td></td>
</tr>
<tr>
<td>Part name</td>
<td>dsPIC33FJ16GS502</td>
</tr>
<tr>
<td>Switching/sampling frequency</td>
<td>65kHz</td>
</tr>
<tr>
<td>Number operation</td>
<td>Q15 method based on fractional data type</td>
</tr>
<tr>
<td>ADC resolution</td>
<td>10 bits</td>
</tr>
<tr>
<td>( K_{adc} )</td>
<td>0.3</td>
</tr>
<tr>
<td>( FM )</td>
<td>2.2245</td>
</tr>
</tbody>
</table>
in conventional control does not resemble sine wave and is distorted. However, in Fig. 5(b), saddles are observed in the inductor current envelope as explained in Fig. 4. The line input current of the proposed control scheme is much less distorted and looks like sinusoid.

Measured harmonic components of two control methods at 75% load are 31.9% (conventional) and 14.9% (proposed). The proposed control outperforms the conventional control by obtaining lower THD and higher PF.

IV. CONCLUSION

An average current control method for digital DCM PFC rectifier has been proposed. The control method achieves lower current distortion by employing a conventional sensing circuit method than conventional open-loop control method. Operational modes, current shaping, and rectifier design procedure have been analyzed and explained with mathematical expressions. The proposed control technique has been verified and compared with conventional one by experimental results of 200W a prototype system. Waveforms and measured data have proved that the proposed control with average current sensing and sampling technique achieves lower line current distortion.

V. ACKNOWLEDGEMENT

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REFERENCES


Figure 5. Current waveforms comparison between (a) conventional open-loop control and (b) proposed variable duty cycle control.