Design-Oriented Analysis and Performance Evaluation of Clamped-Current-Boost Input-Current Shaper for Universal-Input-Voltage Range

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Abstract—In cost-sensitive off-line applications, compliance with the existing line-current harmonic standards can be achieved by employing clamped-current-boost (CCB) input-current shapers (ICS's). In this paper, a thorough analysis of the CCB ICS is presented, and a complete design-oriented mathematical model is derived. The design equations are given in closed forms so that they can be easily computed by any standard mathematical software. The model is verified experimentally on a 100-W universal-input-voltage-range ICS.

Index Terms—Boost ac–dc converter, clamped-current control, line-current shaper, power factor correction.

I. INTRODUCTION

As it has been demonstrated in [1]–[4], various power-quality standards, such as IEC 1000-3-2 [5], can be met by line-current shapers that are substantially different from sinusoidal ones. A trapezoidal waveshaper is particularly attractive due to its potential to achieve high power factor (PF) with low peak-to-rms ratio (crest factor) and relatively low total harmonic distortion (THD) [1]. In [1]–[4], trapezoidal-waveshape approaches are exploited in implementing various simple low-cost boost input-current shapers (ICS's). These ICS's, also known as clamped-current-boost (CCB) ICS's [2]–[4], do not require dedicated ICS controllers. Instead, any conventional current-mode pulse-width-modulation (PWM) controller chip can be used to control the CCB ICS's.

The conceptual circuit diagram of the CCB ICS is shown in Fig. 1. In this circuit, the PWM modulator compares the difference between reference voltage $V_{\text{ref}}$ and slope-compensation ramp $v_R$ (used to stabilize the current loop for duty cycles >50%) with the voltage on sensing resistor $R_s$, which is proportional to inductor current $i_L$. It should be noted that in the CCB ICS, reference voltage $V_{\text{ref}}$ is not derived from the rectified line voltage, but it is equal to the output voltage of the error amplifier $V_e$. Since for ICS's the crossover frequency of the voltage-regulation loop is much lower than twice the line frequency (typically, four–eight times) [6], the output voltage of the error amplifier is practically constant during a half-line cycle. As a result, the peak inductor current is limited (clamped) by the constant reference $V_{\text{ref}}$, which makes the input current roughly trapezoidal. Note that in practical circuits, sensing of inductor current $i_L$ is implemented by sensing the switch current $i_Q$ since during on-time $i_L = i_Q$.

So far, a number of papers have been published dealing with the analysis and design of the CCB ICS's. In [2], detailed analysis of the CCB ICS is performed for the circuit operating with maximum duty cycle < 0.5 and without external ramp. A simplified analysis and design guidelines for the CCB ICS operating with maximum duty cycle close to 100% and external ramp are presented in [3]. Finally, in [4], the average-current control scheme for the CCB ICS is proposed.

The objective of this paper is to present a complete design-oriented analysis of the clamped-current control for CCB ICS's by extending the analysis given in [3] to include the effects of the boost inductance, switching frequency, and maximum duty cycle of the controller. The design equations that include all these effects are derived in closed forms so that they can easily be computed by any standard mathematical software, thus making it possible for design engineers to apply the given design equations to their own sets of design specifications. The presented mathematical model of and design procedure for the CCB ICS are experimentally verified on a universal input-voltage-range 100-W prototype converter.

II. ANALYSIS

In this section, the analysis of the CCB ICS shown in Fig. 1 is performed assuming the following.
1) Input voltage is a full-wave-rectified sine wave, i.e.,
\[ v_i = V_{im} \sin(\omega t) \]
where \( V_{im} \) is the amplitude and \( \omega \) is the angular frequency.
2) The dc-output-voltage \( V_o \) has a negligible ac ripple.
3) Switching frequency \( f_s \) is constant and much larger than line frequency \( f_L \), so that the input voltage can be considered constant during a switching cycle (quasi-static approach).
4) Reference voltage \( V_{ref} \) to the PWM modulator is constant during each half of a line cycle because the bandwidth of the output-voltage loop is much smaller than the rectified line frequency \( f_L \).
5) Phase shift of the line current caused by the input filter can be neglected.

In the CCB ICS in Fig. 1, the conduction of switch \( Q \) is initiated by the internal oscillator of the controller (not shown in Fig. 1). The switch is turned off either when sensed voltage \( R_1 \cdot i_L \) reaches the difference between reference voltage \( V_{ref} = V_e \) and slope-compensation-ramp voltage \( v_R \), i.e.,
\[ R_1 i_L = V_{ref} - v_R \]  
(1)
as shown in Fig. 2(a) or when the duty cycle of the switch reaches its preset maximum \( D_{\text{max}} \), as illustrated in Fig. 2(b).

Within a half-line cycle, the boost inductor can operate in both the discontinuous-conduction mode (DCM) and continuous-conduction mode (CCM). At lower instantaneous line voltages, the boost-inductor current \( i_L \) is discontinuous, while at higher line voltages \( i_L \) is continuous. Depending on which event terminates the conduction of \( Q \) in a switching cycle, two discontinuous and two continuous conduction modes of operation are possible. In this paper, the DCM and CCM, where \( Q \) turns off when the switch duty ratio reaches \( D_{\text{max}} \) [Fig. 2(b)], are denoted as DCM1 and CCM1, respectively. Similarly, the DCM and CCM in which sensed voltage \( R_1 i_L \) reaches the difference voltage \( V_{ref} - v_R \) [Fig. 2(a)] are denoted as DCM2 and CCM2, respectively. From the two CCM’s, only CCM2 needs to be considered since CCM1 usually encum-
slope-compensation-ramp current \(i_R\) are defined from (1) as

\[
i_L = \frac{V_{\text{ref}}}{R_i} - \frac{v_R}{R_i} = I_{\text{ref}} - i_R. \tag{2}
\]

To obtain the expressions for inductor current waveforms and boundary angles, the slope of the compensation ramp must be determined first.

### A. Slope-Compensation Ramp

To ensure the stability of the current loop, the slope of the compensation (external) ramp \(S_e\) should be at least 50% of the maximum down slope of the inductor current in CCM2 \(S_{f_{\text{max}}}^i\), i.e.,

\[
S_e^i = k_S \cdot S_{f_{\text{max}}}^i \quad k_S \geq 0.5. \tag{3}
\]

From Fig. 2(a) and (2), the slope of the compensation ramp is

\[
S_e^i = \frac{I_R}{D_{\text{max}} T_s} = \frac{I_{\text{RM}}}{T_s}. \tag{4}
\]

The inductor current in CCM2 has a maximum down slope at the DCM1-CCM2 boundary

\[
S_{f_{\text{max}}}^i = \frac{V_o - V_{\text{DLC2}}}{L} \tag{5}
\]

where the input voltage at the DCM1-CCM2 boundary is

\[
V_{\text{DLC2}} = (1 - D_{\text{max}}) \cdot V_o. \tag{6}
\]

Substituting (6) into (5) and using (3), the slope of the compensation ramp is determined as

\[
S_e^i = k_S \cdot \frac{D_{\text{max}} V_o}{L}. \tag{7}
\]

Finally, from (4) and (7), the amplitude of the compensation ramp \(I_{\text{RM}} = V_{\text{RM}}/R_i\) is obtained as

\[
I_{\text{RM}} = k_S \cdot \frac{D_{\text{max}} V_o}{L T_s}. \tag{8}
\]

where \(f_s = 1/T_s\).

### B. Discontinuous Conduction Mode

In DCM, the average inductor current, which is also the line current, is defined as

\[
i_{\text{L,avg,DCM}} = \frac{T_{\text{on}} + T_{\text{off}}}{2 T_s} \cdot i_{L_{\text{pk,DCM}}}. \tag{9}
\]

where \(T_{\text{off}}\) is the time which it takes for the inductor current to decrease from \(i_{L_{\text{pk,DCM}}}\) to zero and \(i_{L_{\text{pk,DCM}}}\) is the peak inductor current

\[
i_{L_{\text{pk,DCM}}} = \frac{V_{\text{im}} |\sin(\omega T_f)|}{L I_{\text{f}}} D_{\text{DCM}}. \tag{10}
\]

where \(D_{\text{DCM}} = T_{\text{on}}/T_s\). \(T_{\text{off}}\) is determined from the inductor flux balance as

\[
T_{\text{off}} = \frac{V_{\text{im}}}{V_o} \frac{|\sin(\omega T_f)|}{|\sin(\omega T_f)|} T_{\text{on}}. \tag{11}
\]

Substituting (10) and (11) into (9), the average inductor current in DCM is obtained as

\[
i_{L_{\text{avg,DCM}}} = \frac{1}{2L f_s} \cdot \frac{V_{\text{im}} |\sin(\omega T_f)|}{1 - \frac{V_{\text{im}} |\sin(\omega T_f)|}{V_o}} D_{\text{DCM}}^2. \tag{12}
\]

In DCM1, \(D_{\text{DCM}} = D_{\text{max}}\) and (12) becomes

\[
i_{L_{\text{avg,DCM1}}} = \frac{1}{2L f_s} \cdot \frac{V_{\text{im}} |\sin(\omega T_f)|}{1 - \frac{V_{\text{im}} |\sin(\omega T_f)|}{V_o}} D_{\text{max}}^2. \tag{13}
\]

It should be noted that at low line voltages \(V_{\text{im}} |\sin(\omega T_f)| \ll V_o\), the average inductor current in DCM1 is proportional to the line voltage.

The maximum value of the peak inductor current in DCM1, which determines the condition for MS1, is obtained by setting \(D_{\text{DCM}} = D_{\text{max}}\) and by substituting (6) for the instantaneous rectified line voltage \(V_{\text{im}} |\sin(\omega T_f)|\) in (10)

\[
i_{L_{\text{pk,DCM1, max}}} = \frac{D_{\text{max}} (1 - D_{\text{max}}) V_o}{L f_s}. \tag{14}
\]
In DCM2, the inductor current reaches the difference of the reference current and ramp current, i.e.,

\[ i_{L,DCM2} = I_{ref} - I_{RM}D_{DCM2}. \]  

(15)

From (10) and (15)

\[ D_{DCM2} = \frac{I_{ref}}{I_{RM} + \frac{V_{in}[\sin(\omega L t)]}{L_f s}} \]  

(16)

and substituting (16) into (12), the average inductor current in DCM2 is obtained as

\[ i_{L,\text{ave},DCM2} = \frac{1}{2L_f s} \cdot \left( \frac{I_{ref}}{I_{RM}} \right)^2 \cdot \frac{V_{in}[\sin(\omega L t)]}{V_o} \cdot \left( 1 - \frac{V_{in}[\sin(\omega L t)]}{V_o} \right) \cdot \left( 1 + \frac{V_{in}[\sin(\omega L t)]}{I_{RM}L_f s} \right)^2. \]  

(17)

Again, at low line voltages, \( V_{in}[\sin(\omega L t)] \ll V_o \) and \( V_{in}[\sin(\omega L t)] \ll I_{RM}L_f s = k_SD_{\text{max}}V_o \) [according to (8)], and the average inductor current in DCM2 is proportional to the line voltage.

C. Continuous Conduction Mode

In CCM2, the average inductor current is defined as

\[ i_{L,\text{ave},CCM2} = i_{L,\text{pk},CCM2} - \frac{\Delta i_{L,CCM2}}{2} \]  

(18)

where

\[ i_{L,\text{pk},CCM2} = I_{ref} - I_{RM}D_{CCM2} \]  

(19)

is the peak inductor current and \( \Delta i_{L,CCM2} \) is the peak-to-peak inductor-current ripple

\[ \Delta i_{L,CCM2} = \frac{V_{in}[\sin(\omega L t)]}{L_f s} \cdot D_{CCM2}. \]  

(20)

From the inductor flux balance, the switch duty ratio is determined as

\[ D_{CCM2} = 1 - \frac{V_{in}[\sin(\omega L t)]}{V_o}. \]  

(21)

Substituting (19)–(21) into (18), the average inductor current in CCM2 is obtained as

\[ i_{L,\text{ave},CCM2} = I_{ref} - I_{RM} + \left( \frac{I_{RM}}{V_o} - \frac{1}{2L_f s} \right) V_{in}[\sin(\omega L t)] + \frac{V_{in}^2}{2L_f s V_o} \sin^2(\omega L t). \]  

(22)

It can be seen from (22) that the average inductor current in CCM2 consists of three components: a constant-current component, a component proportional to the line voltage (desired component), and a component proportional to the square of the line voltage. The three CCM2 current components are shown in Fig. 4.

D. Boundary Angle Between Operation Modes

The DCM1-DCM2 boundary angle is obtained from (10) and (15) at \( D_{\text{DCM}} = D_{\text{max}} \)

\[ \theta_{DLD2} = a \sin \left[ \frac{L_f s}{V_{in}} \left( \frac{I_{ref} - I_{RM}}{D_{\text{max}} - I_{RM}} \right) \right] \]  

(25)

while the DCM1-CCM2 boundary angle is directly determined from (6)

\[ \theta_{DLC2} = a \sin \left( \frac{V_o}{V_{in}} \left( 1 - D_{\text{max}} \right) \right). \]  

(26)

Finally, the DCM2-CCM2 boundary angle \( \theta_{D2C2} \) is obtained by equating the peak inductor current from (19) with the peak-to-peak inductor-current ripple from (20), and using (21)

\[ I_{ref} = I_{RM} \left( 1 + \frac{V_{in}\sin(\theta_{D2C2})}{I_{RM}L_f s} \right) \left( 1 - \frac{V_{in}\sin(\theta_{D2C2})}{V_o} \right). \]  

(27)
E. Reference Current

The reference current in (17), (22), (25), and (27) can be determined from the input-output power balance

\[ P_i = \frac{2}{\pi} V_{in} \left( \int_{0}^{\theta_{DD}} i_{L,ave,DCM1}(\theta) \sin(\theta) \, d\theta + \int_{\theta_{DD}}^{\theta_{DC}} i_{L,ave,DCM2}(\theta) \sin(\theta) \, d\theta + \int_{\theta_{DC}}^{\pi/2} i_{L,ave,CCM2}(\theta) \sin(\theta) \, d\theta \right) = \frac{P_o}{\eta} \]  

(28)

where \( \theta = \omega_L t \) and \( \eta \) is the efficiency of the ICS power stage. Equation (28) encompasses all three MS's from Table I. The definition of boundary angles \( \theta_{DD} \) and \( \theta_{DC} \) is given in Table I.

The input power during CCM2 can be expressed in a closed form. By substituting (22) into (28), after integration it follows:

\[ P_{i,CCM2} = \frac{2}{\pi} V_{in}(I_{ref} - I_{RM}) \cos(\theta_{DC}) + \frac{1}{\pi} V_{in}^2 \left( \frac{I_{RM}}{V_o} - \frac{1}{2Lf_a} \right) \cdot \left( \frac{\pi}{2} - \theta_{DC} + \sin(2\theta_{DC}) \right) + \frac{1}{3\pi} \frac{V_{in}^3}{L_f a V_o} \cos(\theta_{DC})(2 + \sin^2(\theta_{DC})). \]  

(29)

The input power during DCM1 and DCM2 can be obtained by numerical integration of the following:

\[ P_{i,DCM1} = \frac{1}{\pi} \frac{V_{in}^2 D_{max}}{L_f a} \int_{0}^{\theta_{DD}} \frac{\sin^2(\theta)}{1 - \frac{V_{in} \sin(\theta)}{V_o}} \, d\theta \]  

(30)

and

\[ P_{i,DCM2} = \frac{1}{\pi} \frac{V_{in}^2}{L_f a} \left( \frac{I_{ref}}{I_{RM}} \right)^2 \cdot \left( \int_{\theta_{DD}}^{\theta_{DC}} \frac{\sin(\theta)}{1 - \frac{V_{in} \sin(\theta)}{V_o}} \, d\theta + \frac{1}{L_f a} \frac{V_{in} \sin(\theta)}{I_{RM} L_f a} \right)^2 \]  

(31)

In MS1, the boundary angles are equal, \( \theta_{DD} = \theta_{DC} = \theta_{DLC2} \), and, as can be seen from (26), independent of the reference current. Hence, by substituting \( \theta_{DLC2} \) from (26) into (29) and (30), \( I_{ref} \) can be directly determined from the power balance (28).

In MS3, the boundary angles are \( \theta_{DD} = 0 \) and \( \theta_{DC} = \theta_{DLC2} \), defined by (27). After substituting \( I_{ref} \) from (27) into (29) and (31), from the power balance (28) the boundary angle \( \theta_{DLC2} \) can be found first. Then, \( I_{ref} \) can be calculated from (27).

Finally, in MS2, the boundary angles are \( \theta_{DD} = \theta_{DLD2} \) (25) and \( \theta_{DC} = \theta_{DLC2} \) (27). Again, after replacing \( I_{ref} \) from (27) into (25) and (29)–(31), from the power balance (28) the boundary angle \( \theta_{DLC2} \) can be found first, and then \( I_{ref} \) can be calculated from (27).

The whole procedure described above can be easily implemented by using any standard mathematical software (e.g., Mathcad). In a particular design, \( I_{ref} \) is calculated for all three MS's. The actual value of \( I_{ref} \), i.e., the actual MS, is the one which satisfies one MS condition from Table I.

F. Input-Current Harmonics

The input current contains only odd harmonics whose rms value can be determined by using Fourier analysis

\[ I_{i,k} = \frac{2\sqrt{2}}{\pi} \left( \int_{0}^{\theta_{DD}} i_{L,ave,DCM1}(\theta) \sin(k\theta) \, d\theta + \int_{\theta_{DD}}^{\theta_{DC}} i_{L,ave,DCM2}(\theta) \sin(k\theta) \, d\theta + \int_{\theta_{DC}}^{\pi/2} i_{L,ave,CCM2}(\theta) \sin(k\theta) \, d\theta \right). \]  

(32)

The THD is obtained by summing up the squares of the first \( N \) (e.g., \( N = 19 \)) odd harmonics

\[ \text{THD} = \sqrt{\frac{\sum_{k=3}^{2N+1} P_{i,k}^2}{I_{i,1}}}, \quad k = 3, 5, \ldots, (2N+1). \]  

(33)
Fig. 6. Line-current waveform versus $k_S$ at $V_{\text{rms}} = 90$ V in (a) MS1, (b) MS2, and (c) MS3.

III. DESIGN

In this section, the derived mathematical model is used in the design of an experimental 100-W/385-V dc CCB ICS for the universal input-voltage range ($V_{\text{rms}} = 90$–265 V).

According to the line-current components, determined by (13), (17), and (22), the design variables are the product of boost inductance $L$ and switching frequency $f_s$, maximum duty cycle $D_{\text{max}}$, and the height of the ramp current, $I_{RM}$. The design of $L f_s$ and $D_{\text{max}}$ is determined by the specifications of the power stage and performed as for a conventional boost ICS circuit [6]. Therefore, the design of the CCB ICS in this paper is focused on the control circuit, particularly on the ramp current height $I_{RM}$ determined by (8). The key design parameter is the normalized slope of the ramp current $k_S$.

For higher line voltages, i.e., $V_{\text{rms}} > 180$ V, the operation corresponds to MS3. In MS3, the quality of the line current is inversely proportional to $k_S$, i.e., with increasing $k_S$, PF decreases and THD increases. As an example, line-current waveforms obtained in Mathcad for three different values of $k_S$ at $V_{\text{rms}} = 230$ V ($L = 0.5$ mH, $f_s = 100$ kHz, $D_{\text{max}} = 0.9$, and $\eta = 0.9$) are shown in Fig. 5—the corresponding values of PF and THD (calculated for the first
19 odd harmonics) are given in Table II.

At lower line voltages, all three MS’s are possible. For example, at minimum line voltage \( V_{\text{min}} = 90 \) V, with increasing \( k_S \) the operation changes from MS1 \( \rightarrow \) MS2 \( \rightarrow \) MS3, as presented in Table III and Fig. 6. In MS2, similarly as in MS3, the quality of the line current is inversely proportional to \( k_S \), i.e., with increasing \( k_S \), PF decreases and THD increases. However, in MS1, the quality of the line current varies proportionally to \( k_S \), i.e., with increasing \( k_S \), PF increases and THD decreases. It was found to be a good design compromise to select \( k_S \) so that at minimum line voltage, operation in MS1 close to the boundary with MS2 is achieved. It follows from Table III that \( k_S = 1 \).

Line-current waveforms obtained in Mathcad with \( L = 0.5 \) mH, \( f_s = 100 \) kHz, \( D_{\text{max}} = 0.9 \), \( \eta = 0.9 \), and \( k_S = 1 \) for five rms line voltages are shown in Fig. 7. The circuit operates in MS1 at \( V_{\text{rms}} = 90 \) V, in MS2 at \( V_{\text{rms}} = 100 \) V and \( V_{\text{rms}} = 120 \) V, and in MS3 at \( V_{\text{rms}} = 230 \) V and \( V_{\text{rms}} = 265 \) V. Corresponding values of PF and THD are given in Table IV. The values of \( I_{\text{ref}} \) and boundary angles \( \theta_{DD} \) and \( \theta_{DC} \) are also included in Table IV. In Figs. 5–7, ideal sinusoidal line-current waveforms are also shown.

It should be noted that the line-current waveform and PF at higher line voltages can be improved by employing nonlinear feedforward control [3].

### Table III

MS, PF, and THD versus \( k_S \) at \( V_{\text{rms}} = 90 \) V

<table>
<thead>
<tr>
<th>( k_S )</th>
<th>MS</th>
<th>PF</th>
<th>THD [%]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.5</td>
<td>1</td>
<td>0.98</td>
<td>20.4</td>
</tr>
<tr>
<td>0.75</td>
<td>1</td>
<td>0.993</td>
<td>11.6</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>0.999</td>
<td>4.3</td>
</tr>
<tr>
<td>1.25</td>
<td>2</td>
<td>0.997</td>
<td>8.2</td>
</tr>
<tr>
<td>1.5</td>
<td>2</td>
<td>0.99</td>
<td>14.4</td>
</tr>
<tr>
<td>1.75</td>
<td>2</td>
<td>0.983</td>
<td>18.6</td>
</tr>
<tr>
<td>2</td>
<td>3</td>
<td>0.977</td>
<td>21.9</td>
</tr>
<tr>
<td>2.5</td>
<td>3</td>
<td>0.965</td>
<td>27.0</td>
</tr>
</tbody>
</table>

The experimental circuit diagram is shown in Fig. 8. The control circuit is based on the conventional current-mode PWM controller 3842 [7]. Design of resistors \( R_2, R_9, R_{10}, \) and \( R_{11} \) which are related to the external ramp is given in [8].

### IV. Experimental Results

Experimental line-current waveforms at full resistive load (\( I_{\text{LOAD}} = 0.26 \) A) for four rms line voltages are presented in Fig. 9(a)–(c) and (e). The experimental waveforms are in good agreement with the theoretical waveforms, except for...
In order to improve the quality of the line-current waveform at higher rms line voltages, i.e., to increase PF and decrease THD, feedforward control [3] is employed at $V_l(t) > 225$ V. The nonlinear feedforward control is implemented with three 75-V Zener diodes in series, as shown in Fig. 8. Instead of high-voltage Zener diodes, low-voltage Zener diodes can be used with an additional resistive voltage divider. Notice that at lower rms line voltages ($V_{\text{rms}} = 100$ V and $V_{\text{rms}} = 120$ V), the feedforward control has no effect, i.e., the Zener diode $D_5$ in Fig. 8 is never conducting. As can be seen from Fig. 9(d) and (f), the feedforward limits the peak current and makes

<table>
<thead>
<tr>
<th>$V_{\text{rms}}$ [V]</th>
<th>MS</th>
<th>$I_{\text{ef}}$ [A]</th>
<th>$\theta_{\text{do}}$ [°]</th>
<th>$\theta_{\text{dc}}$ [°]</th>
<th>PF</th>
<th>THD [%]</th>
</tr>
</thead>
<tbody>
<tr>
<td>90</td>
<td>1</td>
<td>7.22</td>
<td>18</td>
<td>18</td>
<td>0.999</td>
<td>4.3</td>
</tr>
<tr>
<td>100</td>
<td>2</td>
<td>6.93</td>
<td>16</td>
<td>16</td>
<td>0.998</td>
<td>5.7</td>
</tr>
<tr>
<td>120</td>
<td>2</td>
<td>6.27</td>
<td>1</td>
<td>52</td>
<td>0.995</td>
<td>9.8</td>
</tr>
<tr>
<td>230</td>
<td>3</td>
<td>2.66</td>
<td>0</td>
<td>71</td>
<td>0.942</td>
<td>35.7</td>
</tr>
<tr>
<td>265</td>
<td>3</td>
<td>1.49</td>
<td>0</td>
<td>66</td>
<td>0.825</td>
<td>68.2</td>
</tr>
</tbody>
</table>
the current waveform wider, resulting in higher PF and lower THD. The measured PF and THD for five rms line voltages are presented in Table V. The measured line-current harmonics at $V_{\text{rms}} = 100$ V and $V_{\text{rms}} = 230$ V are shown in Fig. 10. The harmonic limits for Class D from the IEC 1000-3-2 standard [5] are also given in Fig. 10. The limits for $V_{\text{rms}} = 100$ V are obtained by multiplying the $V_{\text{rms}} = 230$ V limits with $230/100 = 2.3$. As can be seen from Fig. 10, the measured harmonics are well below the limits. It should be noted that the IEC 1000-3-2 specifications are satisfied even without the implementation of the feedforward control.

At higher rms line voltages, the line current contains some irregularities which can be explained with the circuit behavior in DCM. Namely, after the boost diode turns off, the MOSFET drain-source capacitance and the boost diode capacitance oscillate with the boost inductance, as shown in the inductor-current waveforms in Fig. 11. The amount of charge transferred back into the filter capacitance $C_1$ varies depending on the duration of DCM. For example, in Fig. 11(c), the DCM oscillation consists of three half sinusoids, which means that more charge is transferred into $C_1$ than out of $C_1$, i.e., extra charge is available from $C_1$ to supply the load and less charge has to be drawn from the line. Hence, the line current slightly decreases around interval $T_{2n}$. In DCM around interval $T_3$, there are two full-sinusoidal oscillations, and no extra charge is accumulated in $C_1$, resulting in no distortion of the line current. Around interval $T_1$, the circuit operates in CCM. The irregularities in the line-current waveform at higher line voltages can be made smaller by increasing the filter capacitance $C_1$.

![Fig. 10. Experimental line-current harmonics at full resistive load at (a) $V_{\text{rms}} = 100$ V and (b) $V_{\text{rms}} = 230$ V with and without feedforward control (FFC).](image)

<table>
<thead>
<tr>
<th>$V_{\text{rms}}$ [V]</th>
<th>w/o FFC</th>
<th>w/ FFC</th>
</tr>
</thead>
<tbody>
<tr>
<td>90</td>
<td>0.994</td>
<td>0.994</td>
</tr>
<tr>
<td>100</td>
<td>0.996</td>
<td>0.996</td>
</tr>
<tr>
<td>120</td>
<td>0.979</td>
<td>0.979</td>
</tr>
<tr>
<td>230</td>
<td>0.876</td>
<td>0.954</td>
</tr>
<tr>
<td>265</td>
<td>0.706</td>
<td>0.851</td>
</tr>
</tbody>
</table>
V. SUMMARY

A design-oriented analysis of the CCB ICS that includes the effects of the boost inductance, switching frequency, and maximum duty cycle of the controller is presented. The derived design equations are given in closed forms so that they can easily be computed by any standard mathematical software. The presented mathematical model of and proposed design procedure for the CCB ICS are experimentally verified on a universal-input-voltage-range 100-W converter. It is demonstrated that the CCB ICS meets the IEC 1000-3-2 standards.

REFERENCES


Laszlo Huber (M’86), for a photograph and biography, see this issue, p. 486.

Milan M. Jovanović (S’86–M’89–SM’89), for a photograph and biography, see this issue, p. 486.