

# Direct AC/DC Rectifier With Mitigated Low-Frequency Ripple Through Inductor-Current Waveform Control

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**Abstract**—In a rectification system with unity power factor, the input power consists of a dc and a double-line frequency power component. Traditionally, an electrolytic capacitor (E-Cap) is used to buffer the double-line frequency power such that the dc output presents a small voltage ripple. The use of E-Cap significantly limits the lifetime of the rectifier system. In this paper, a differential ac/dc rectifier based on the use of an inductor-current waveform control methodology is proposed such that a single-stage direct ac/dc rectification without the need of an E-Cap for buffering the double-line frequency power, and a front-stage diode rectifier circuit can be achieved. The feasibility of the proposal has been practically confirmed in an experimental prototype.

**Index Terms**—E-capless, LED, low-frequency ripple, rectifier, single-stage, waveform control.

## I. INTRODUCTION

UNITY power factor (PF), high efficiency, high reliability, compact size, and low cost are the typical desired features of conventional ac/dc rectification systems. Traditionally, electrolytic capacitors (E-caps) of large capacitance are used to buffer the double mains frequency ripple current or power in the intermediate stage between the ac input and the dc output. The reduction of such ripple current or power is particularly important in lighting and battery-charging applications because the low-frequency current of 100 or 120 Hz could cause flickering in lighting systems and is detrimental to the storage properties of the batteries. Due to the relatively short lifetime of E-caps, many research efforts have been devoted to develop ac–dc power converter topologies without using E-caps [1]–[17]. Both active [1]–[15] and passive [16], [17] approaches have been reported in the recent literature.

For the active approach, an auxiliary circuit (typically comprising a bidirectional ac-dc switching circuit, an inductor, and a capacitor) [1]–[11] can be added to conventional diode rectifier

or full bridge inverter to buffer the ripple power. The energy stored in a capacitor is  $0.5 CV_C^2$ , where  $C$  is the capacitance and  $V_C$  is the capacitor voltage. In order to buffer the ripple power with reduced capacitor size, the general principles include:

- 1) charging and discharging a relatively small nonelectrolytic capacitor to a relatively high capacitor voltage in a separate ripple port [1], [3] in order to absorb the ripple power as shown in Fig. 1;
- 2) decoupling the ac power by linking the auxiliary circuit to either the ac input port [2], [4] or the dc ground [5], [11] or another inverter leg [6], or the output dc port [9];
- 3) integrating the auxiliary circuit into a standard switched mode converter [7] or switched mode rectifier [8];
- 4) shaping the input ac current in order to reduce the peak-to-average current ratio [12]–[14] and simultaneously meet the IEC harmonics requirements at a reduced PF;
- 5) storing the ripple power in an inductor instead of a capacitor [15].

For the passive approach that does not use any fully controlled active switches such as power MOSFETs, a large inductor is usually used to limit the input power into the diode bridge from the AC grid and thus the output current and power [16], [17]. A small valley-fill circuit [16] or more simply a small capacitor [17] can be used to reduce the ripple power of the dc port. Generally, some output current ripple is allowed because it has been pointed out that the luminous flux variation arising from the power fluctuation in an LED system can be reduced by proper thermal design of the heatsink [18]. Such passive LED systems are particularly suitable for outdoor applications and have reached commercialization stage for LED street-lighting systems.

In this paper, a direct (electrolytic-capacitor-free) ac/dc rectifier with mitigated low-frequency ripple through the use of an inductor-current waveform control is proposed. This method is different from the capacitor voltage waveform control reported in [19]. The proposed system allows the use of differential dc/dc converter topologies [20] to directly rectify power from the ac source to the load (i.e., a single-stage ac/dc rectification) without the need of the front-stage diode rectifier circuit and the PFC converter. The proposed rectifier is capable of automatically achieving PF correction without the need of a large input filter. The inductor-current waveform control mitigates the double-line frequency current without the need for a large capacitance in the system, thereby eliminating the need for an E-Cap.

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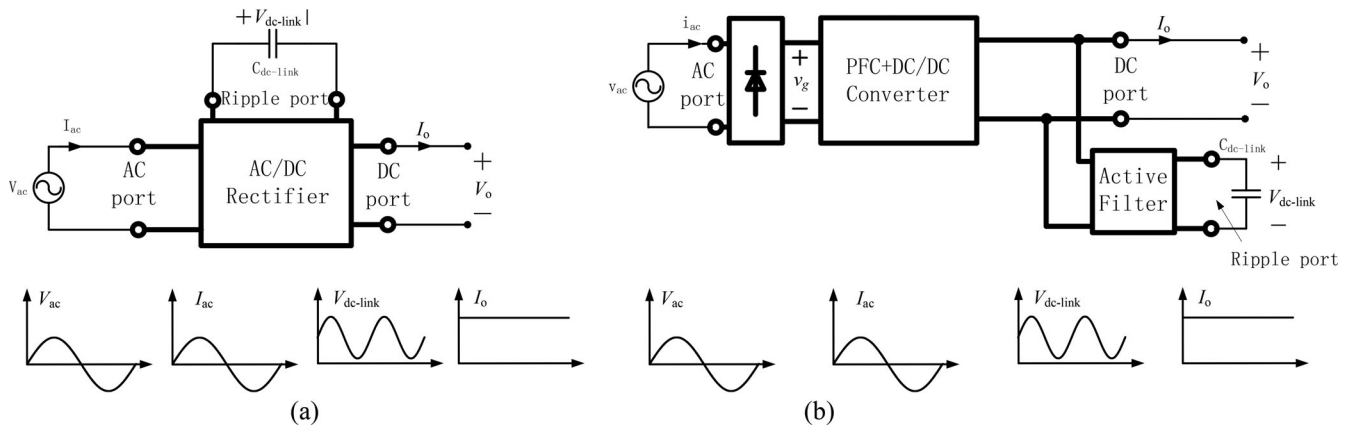


Fig. 1. (a) Conceptual diagram of three-port model of an ac/dc rectifier and (b) a rectification system with parallel active filtering.

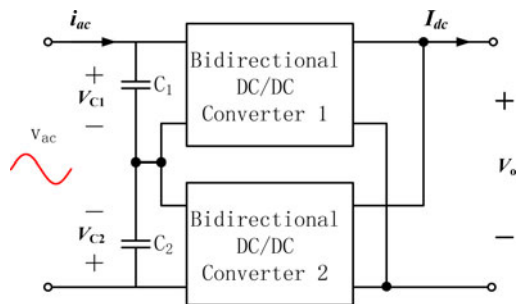


Fig. 2. Schematic of the differential rectifier system used for the inductor-current waveform control.

## II. PROPOSED DIRECT AC/DC RECTIFIER SYSTEM

### A. Concept

Fig. 2 shows the general concept of the direct ac/dc differential rectification system, in which two bidirectional dc/dc converters are connected in a series-input and parallel-output configuration [20]. Here,  $v_{c1}$  and  $v_{c2}$  are, respectively, the input voltage of each of the two converters, whereby  $v_{c1}$  and  $v_{c2}$  are in opposing polarity, and their difference is a pure sinusoidal waveform that follows the shape of the input ac source  $v_{ac}$ . It must be emphasized that

- 1) The proposed differential rectifier is in compliance with existing requirements for typical single-stage ac/dc power converters, e.g., conversion functionality, isolation, PF, input/output filter requirement, etc. The front-stage diode rectifier, large input filter, and the postregulation dc/dc conversion stage are not required.
- 2) The input voltage of the two converters,  $v_{c1}$  and  $v_{c2}$ , can be of many forms, as long as their differential value is equal to the ac voltage source  $v_{ac}$ . This facilitates the use of the waveform control technique, which allows the system to concurrently achieve good-quality single-stage ac/dc rectification and double-line frequency ripple mitigation without the use of E-Cap.

- 3) In the case of adopting bidirectional dc/dc converters in the proposed configuration, the bidirectional power flow ability of the two converters implies both power rectification and inversion capability of the proposed ac/dc differential system.

### B. Topology

The bidirectional dc/dc converters in Fig. 2 can be of any type, depending on the required application. They can be converters without (e.g., buck, boost, buck–boost, Cuk, Zeta) or with galvanic isolation (flyback, push–pull, forward). In fact, many existing rectifier topologies can be treated as two converters with a differential configuration. For instance, the widely applied full-bridge PWM rectifier is functionally equivalent to two bidirectional boost converters connected differentially, as shown in Fig. 3. Fig. 3(a) shows a full-bridge rectifier, whose input filter capacitors are formed by  $C_1$  and  $C_2$ . If the joint point between  $C_1$  and  $C_2$  is connected to the dc ground, as indicated by the dashed line, the full-bridge rectifier can be redrawn in the form of the proposed differential rectifier configuration, as given in Fig. 3(b). In other words, a full-bridge rectifier is a boost type differential rectifier. The step-up property of the boost converter ensures that the full-bridge rectifier can only output a dc voltage with an amplitude higher than that of the ac utility.

In this paper, a buck-type differential rectifier is used as a case-study example to illustrate the concept (see Fig. 4). The rectifier comprises two bidirectional buck converters connected under the proposed differential configuration [20]. Here,  $C_1$ ,  $C_2$ , and  $L_1$ ,  $L_2$  are the input capacitors and output inductors for the respective converter, and  $T_1 - T_4$  are their power switches. In this system, the upper converter is called the high-side converter (formed by  $C_1$ ,  $T_1$ ,  $T_2$ ,  $L_1$ ), and the lower converter is called the low-side converter ( $C_2$ ,  $T_3$ ,  $T_4$ ,  $L_2$ ). The rectifier is capable of directly generating a dc voltage with amplitude either higher or lower than the amplitude of the ac mains. The direct step-down ability makes the buck differential rectifier an interesting solution for applications where the dc output voltage is low since the rectifier requires no further processing stage.

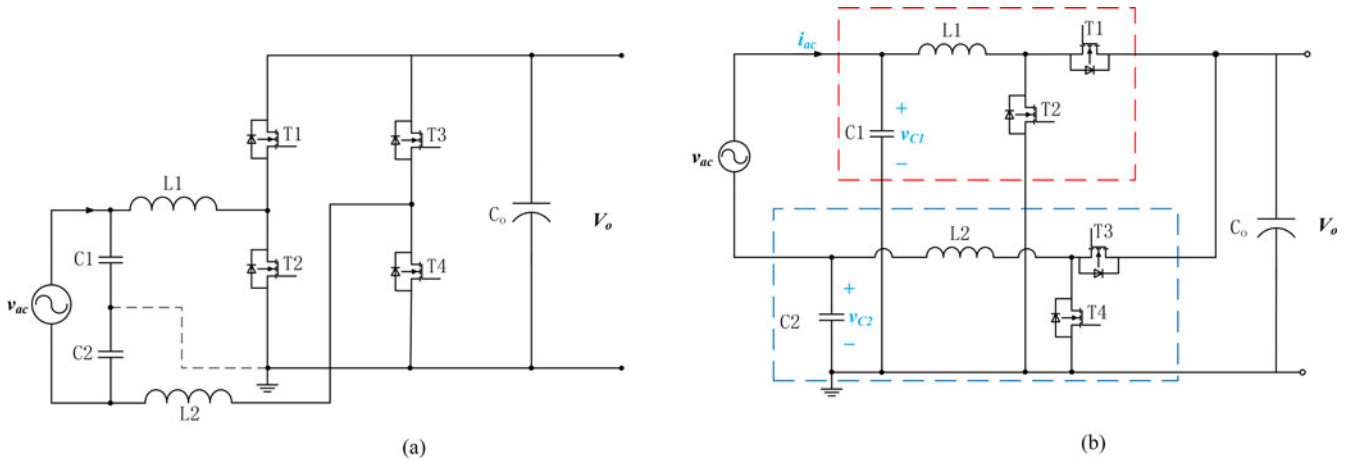


Fig. 3. (a) Conventional full-bridge PWM rectifier and (b) its equivalent differential rectifier model configuration.

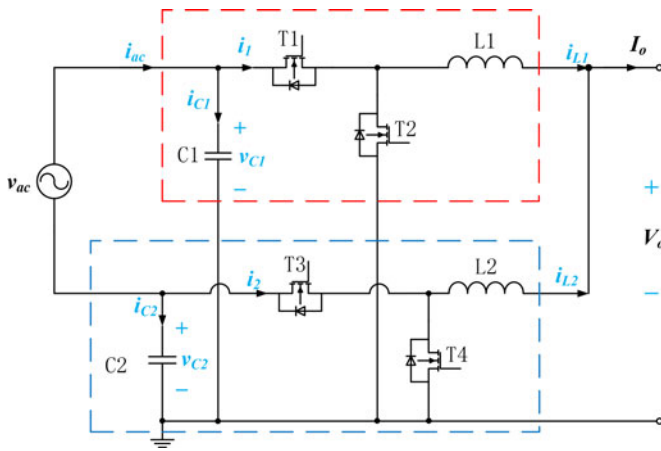


Fig. 4. Buck differential rectifier.

TABLE I  
OPERATING STATES OF THE BUCK DIFFERENTIAL RECTIFIER

Switch State	$T_1$ (high side)	$T_2$ (high side)	$T_3$ (low side)	$T_4$ (low side)
$S_1$	OFF	ON	OFF	ON
$S_2$	OFF	ON	ON	OFF
$S_3$	ON	OFF	ON	OFF
$S_4$	ON	OFF	OFF	ON

### C. Operation of the Buck Differential Rectifier

The capacitor-voltage waveform control method is originally proposed for differential inverters [19], the control of which involves only the regulator for the ac side capacitor voltages  $v_{c1}$  and  $v_{c2}$ . In the case of the differential rectifier, a different set of control variables and a different controller are needed.

Table I shows the four possible operating states  $S_1$ ,  $S_2$ ,  $S_3$ , and  $S_4$  of the buck differential rectifier in a complete line cycle with respect to the ON/OFF states of the switches  $T_1$ - $T_4$ . The corresponding equivalent circuits are given in Fig. 5. In order to operate the rectifier properly, it must be highlighted that it is

not possible for the four states to exist at the same time in one switching cycle, neither during positive line cycle nor during the negative cycle. In the positive half of the line cycle, the possible states are  $S_1$ ,  $S_2$ , and  $S_3$ , and in the negative half of the line cycle, the possible states are  $S_1$ ,  $S_3$ , and  $S_4$ . This phenomenon is caused by the unbalance of the instantaneous input power (dc power plus a double-line frequency component) and the constant output power. Take the positive line cycle as an example. In this line cycle, the high-side converter is providing more power than the dc side requires. In order to keep the output power constant, the excessive energy has to be released back to the ac side into  $C_2$ . Therefore, during the interval when ac power is delivered through high-side converter (when  $T_1$  is ON,  $T_3$  is ON,  $S_3$  mode) without sourcing power to the output ( $T_1$  ON,  $T_4$  ON,  $S_4$  mode). The process is physically inevitable, no matter what control techniques are used.

All possible operating conditions for switches  $T_1$  and  $T_3$  during the positive and negative half-line cycles are depicted in Figs. 6 and 7.

The duty cycles of the rectifier must satisfy (1) such that  $d_{low}$  (duty ratio of the low-side converter with respect to  $T_3$ ) is always larger than or equal to  $d_{high}$  (duty ratio of the high-side converter with respect to  $T_1$ ) during the positive half-line cycle, and vice versa. Here, only the positive half line cycle is examined since the operation is similar between the positive and negative half of the line cycle

$$\begin{cases} v_{ac} > 0 \Rightarrow d_{low} > d_{high} \\ v_{ac} < 0 \Rightarrow d_{low} < d_{high} \\ v_{ac} = 0 \Rightarrow d_{low} = d_{high} \end{cases} \quad (1)$$

By setting  $d_1$ ,  $d_2$ ,  $d_3$ , and  $d_4$  as the ratios of the respective time interval of each state  $S_1$ - $S_4$  shown in Fig. 7 over one switching period,  $d_{low}$  and  $d_{high}$  can be expressed as

$$d_{low} = d_2 + d_3 \quad (2)$$

$$d_{high} = d_3. \quad (3)$$

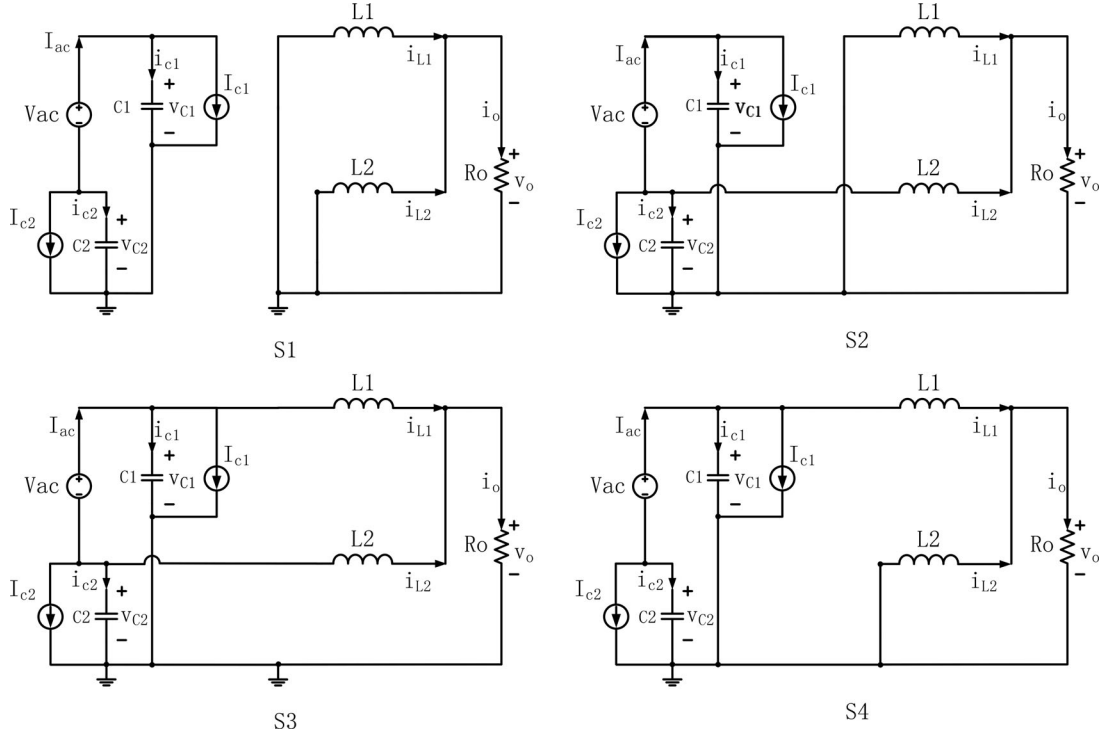


Fig. 5. Equivalent circuits of the buck differential rectifier in various operating states.

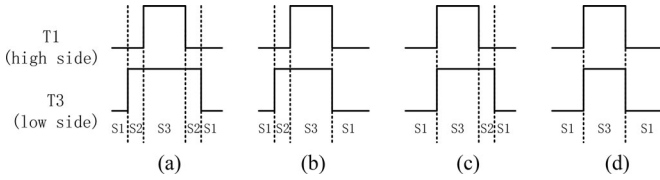


Fig. 6. Possible operation conditions during positive half-line cycle.

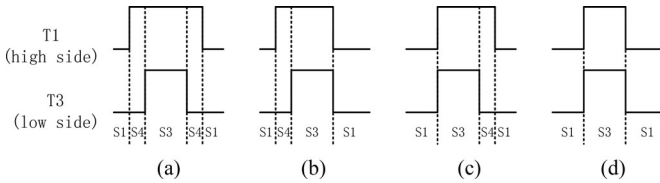


Fig. 7. Possible operation conditions during negative half-line cycle.

Assuming that the switching frequency is much higher than the line frequency, the input ac source during a switching cycle can be treated as a constant dc source. The symbol  $v_{ac}$  used in Fig. 4 is hence modified as  $V_{ac}$  in Fig. 5 to represent the dc source property over the switching cycle in which it is examined. Similarly, the ac line current over the switching cycle is assumed to be a dc current source  $I_{ac}$ . With this consideration, the capacitor currents of  $C_1$  and  $C_2$  will each comprise two components, a high-frequency component ( $i_{c1}$  and  $i_{c2}$ , respectively), introduced by the switching dynamics, and a dc component ( $I_{C1}$  and  $I_{C2}$ ) introduced by the ac source. In performing the control,

the values for  $i_{c1}$  and  $i_{c2}$  will be determined by the waveform control method.

The state-space-averaged equation of the buck differential rectifier can be derived from Fig. 5 as

$$\begin{aligned}
 & \begin{bmatrix} 0 \\ 0 \\ I_{C1} \\ I_{C2} \end{bmatrix} + \begin{bmatrix} L_1 & 0 & 0 & 0 \\ 0 & L_2 & 0 & 0 \\ 0 & 0 & C_1 & 0 \\ 0 & 0 & 0 & C_2 \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{L1}(t) \\ i_{L2}(t) \\ v_{c1}(t) \\ v_{c2}(t) \end{bmatrix} \\
 & = \begin{bmatrix} -R_o & -R_o & d_3 & 0 \\ -R_o & -R_o & 0 & d_2 + d_3 \\ -d_3 & 0 & 0 & 0 \\ 0 & -(d_2 + d_3) & 0 & 0 \end{bmatrix} \begin{bmatrix} i_{L1}(t) \\ i_{L2}(t) \\ v_{c1}(t) \\ v_{c2}(t) \end{bmatrix} \\
 & + \begin{bmatrix} 0 & 0 \\ 0 & 0 \\ 0 & 1 \\ 0 & -1 \end{bmatrix} \begin{bmatrix} v_{ac}(t) \\ i_{ac}(t) \end{bmatrix} \quad (4)
 \end{aligned}$$

where  $R_o$  is the load resistance, and  $i_{L1}$  and  $i_{L2}$  are the inductor currents.

The steady-state-averaged equation of the rectifier can be derived from (4), by setting  $\frac{di_{L1}}{dt} = \frac{di_{L2}}{dt} = \frac{dv_{c1}}{dt} = \frac{dv_{c2}}{dt} = 0$ ,

as

$$\begin{bmatrix} 0 \\ 0 \\ I_{C1} \\ I_{C2} \end{bmatrix} = \begin{bmatrix} -R_o & -R_o & D_3 & 0 \\ -R_o & -R_o & 0 & D_2 + D_3 \\ -D_3 & 0 & 0 & 0 \\ 0 & -(D_2 + D_3) & 0 & 0 \end{bmatrix} \times \begin{bmatrix} I_{L1}(t) \\ I_{L2}(t) \\ V_{c1}(t) \\ V_{c2}(t) \end{bmatrix} + \begin{bmatrix} 0 & 0 \\ 0 & 0 \\ 0 & 1 \\ 0 & -1 \end{bmatrix} \begin{bmatrix} V_{ac}(t) \\ I_{ac}(t) \end{bmatrix} \quad (5)$$

where  $I_{L1}$ ,  $I_{L2}$ ,  $V_{C1}$ ,  $V_{C2}$ ,  $D_2$ , and  $D_3$  are the steady-state values of  $i_{L1}$ ,  $i_{L2}$ ,  $v_{C1}$ ,  $v_{C2}$ ,  $d_2$ , and  $d_3$ .

By solving (5),  $V_{C1}$ ,  $V_{C2}$ ,  $I_{L1}$ , and  $I_{L2}$ , can be derived as

$$\begin{cases} V_{C1} = \frac{D_2 + D_3}{D_2} V_{ac} = \frac{D_{low}}{D_{low} - D_{high}} V_{ac} \\ V_{C2} = \frac{D_3}{D_2} V_{ac} = \frac{D_{high}}{D_{low} - D_{high}} V_{ac} \\ I_{L1} = \frac{1}{D_3} (I_{ac} - I_{C1}) = \frac{1}{D_{high}} I_1 \\ I_{L2} = -\frac{1}{D_2 + D_3} (I_{ac} + I_{C2}) = \frac{1}{D_{low}} I_2 \end{cases} \quad (6)$$

where  $I_1$  and  $I_2$  are the averaged input currents of the high side (i.e.,  $T_1$ ) and the low side (i.e.,  $T_3$ ) converters, respectively.

From (6), the averaged output current and voltage of the rectifier can be resolved as

$$\begin{cases} I_o = I_{L1} + I_{L2} = \frac{1}{D_{high}} I_1 + \frac{1}{D_{low}} I_2 \\ V_o = (I_{L1} + I_{L2}) R_o = \left( \frac{1}{D_{high}} I_1 + \frac{1}{D_{low}} I_2 \right) R_o. \end{cases} \quad (7)$$

It is found from (6) that  $V_{C1}$  and  $V_{C2}$  are coupled functions of  $D_{low}$  and  $D_{high}$ . Therefore, if  $V_{C1}$  and  $V_{C2}$  are chosen as the control variables, the control is not straightforward since it involves the operation of both converters simultaneously. Here,  $D_{low}$  and  $D_{high}$  cannot be resolved explicitly with respect to  $V_{C1}$ ,  $V_{C2}$ , and  $V_{ac}$  since there is no unique solution. Therefore, it is difficult to apply capacitor voltage waveform control.

On the other hand, inductor currents  $I_{L1}$  and  $I_{L2}$  are decoupled between the two converters and are independent functions of  $D_{low}$  and  $D_{high}$ , as can be seen in (6). The duty cycles of the two converters can be easily resolved as

$$\begin{cases} D_{low} = \frac{I_2}{I_{L2}} \\ D_{high} = \frac{I_1}{I_{L1}}. \end{cases} \quad (8)$$

Therefore, by choosing  $I_{L1}$  and  $I_{L2}$  as control variables for the buck differential rectifier, the control becomes simple and straightforward as compared with the original capacitor-voltage waveform control [19] since the two current variables are fully decoupled from one another.

### III. APPLICATION OF THE GENERAL WAVEFORM CONTROL METHOD TO THE PROPOSED DIFFERENTIAL RECTIFIER

#### A. General Waveform Control Method

For the proposed differential rectifier system, a general waveform control technique for mitigating the double-line frequency power at the dc side of the circuit without using an E-cap, is proposed. This method is evolved from the work described in [20], which is originally developed for the full-bridge (boost-type differential) inverters. The waveform control method described here is general and applicable to any differential rectifier systems.

Traditionally, without waveform control, the two capacitor voltages are in the form of

$$v_{c1} = V_d + 0.5V_{max} \sin(\omega t) \quad (9)$$

$$v_{c2} = V_d - 0.5V_{max} \sin(\omega t). \quad (10)$$

Due to the differential connection of  $C_1$  and  $C_2$ , their differential voltage is equal to the ac line voltage, i.e.,

$$v_{ac} = v_{c1} - v_{c2} = V_{max} \sin(\omega t). \quad (11)$$

The dc output  $i_o$  will inevitably contain a significant level of double-line frequency current ripple  $i_o(2\omega)$  as given in

$$i_o = I_o + i_o(2\omega). \quad (12)$$

It is demonstrated in [20] that, given the operation waveforms described in (9) and (10), the two capacitors  $C_1$  and  $C_2$  do not contribute to the absorption of the double-line frequency power from the ac input, but serve only as high-frequency filters. By introducing a double-line frequency component to control  $v_{C1}$  and  $v_{C2}$ ,  $C_1$  and  $C_2$  are able to additionally store the double-line pulsation power, thereby mitigating the double-line frequency ripple at the dc output.

Assuming that the input capacitor voltages are in the general form of

$$v_{c1} = V_d + kV_{max} \sin(\omega t) + B \sin(2\omega t + \varphi) \quad (13)$$

$$v_{c2} = V_d + (k - 1)V_{max} \sin(\omega t) + B \sin(2\omega t + \varphi) \quad (14)$$

where  $k$  is the ratio of the amplitude of the line-frequency component for the high-side converter against the amplitude of the line voltage  $V_{max}$ , and  $B$  is the amplitude of the double-line frequency components that are newly introduced into  $v_{C1}$  and  $v_{C2}$ . The double-line frequency components are injected differentially such that (11) is still satisfied.

The current flow into  $C_1$  and  $C_2$  can be derived, respectively, as

$$i_{c1} = C_1 \frac{dv_{c1}}{dt} = kC_1\omega V_{max} \cos(\omega t) + 2\omega C_1 B \cos(2\omega t + \varphi) \quad (15)$$

$$i_{c2} = C_2 \frac{dv_{c2}}{dt} = (k - 1)C_2\omega V_{max} \cos(\omega t) + 2\omega C_2 B \cos(2\omega t + \varphi). \quad (16)$$

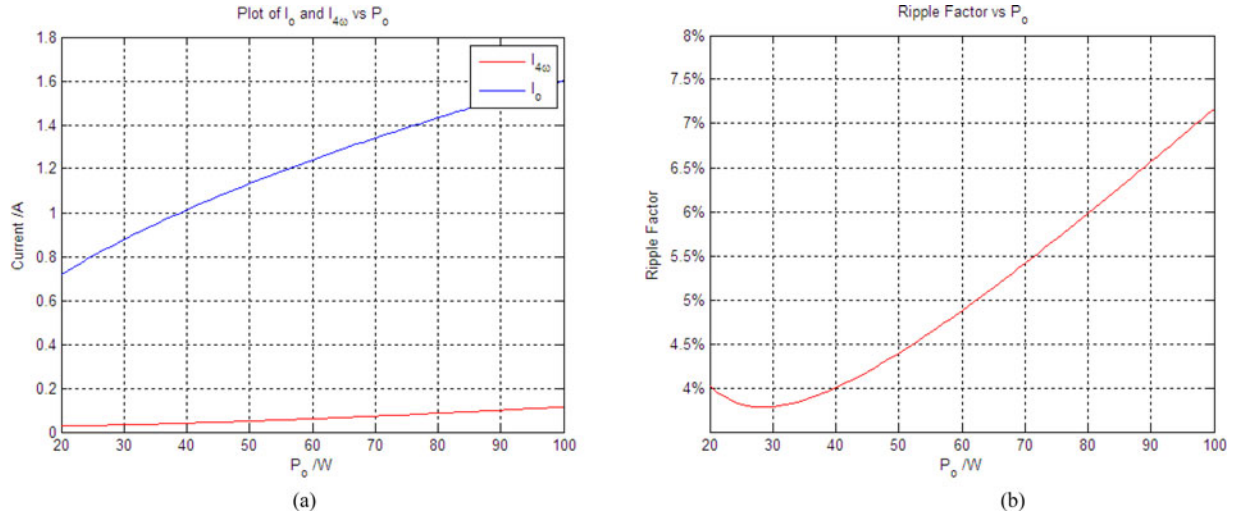


Fig. 8. Fourth-harmonic output current ripple analysis ( $V_{m\max} = 155.5$  V,  $V_d = 200$  V,  $C_1 = C_2 = 15\mu\text{F}$ , and  $R_o = 39\ \Omega$ ). (a) Averaged dc current  $I_o$  and ripple amplitude  $I_{(4\omega)}$  versus output power  $P_o$ . (b) Ripple factor versus output power  $P_o$ .

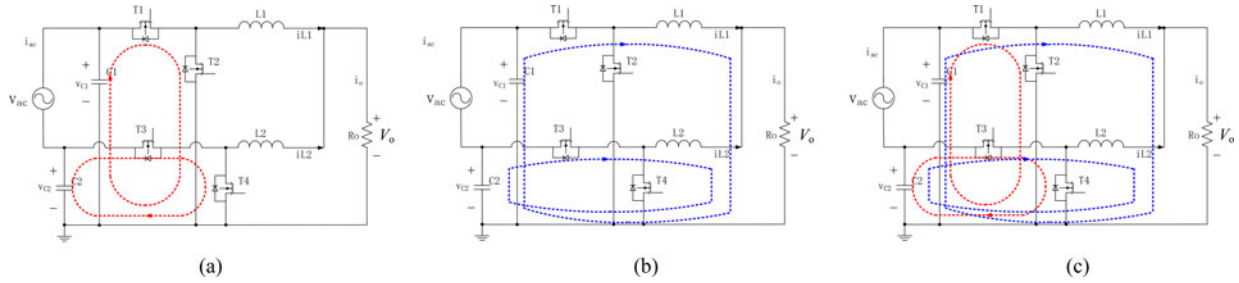


Fig. 9. Flow path of the current of double-line frequency power component in the proposed differential rectifier. (a) Through capacitors. (b) Through inductors. (c) Through capacitors and inductors.

The average input current of the converters in the rectifier can be derived as

$$i_1 = i_{ac} - i_{c1} = I_{\max} \sin(\omega t) - kC_1\omega V_{\max} \cos(\omega t) - 2\omega C_1 B \cos(2\omega t + \varphi) \quad (17)$$

$$i_2 = -i_{ac} - i_{c2} = -I_{\max} \sin(\omega t) - (k-1)C_2\omega V_{\max} \cos(\omega t) - 2\omega C_2 B \cos(2\omega t + \varphi) \quad (18)$$

where  $I_{\max}$  is the peak amplitude of line current.

Assuming the converters to be lossless, the input power equals the output power, of which the inductor currents can be derived as

$$i_{L1} = \frac{i_1 \times v_{c1}}{V_o} = \frac{i_1}{d_{\text{high}}} \text{ and } i_{L2} = \frac{i_2 \times v_{c2}}{V_o} = \frac{i_2}{d_{\text{low}}} \quad (19)$$

From (17)–(19), the total output current is given as

$$i_o = i_{L1} + i_{L2} = I_o + i_{o(\omega)} + i_{o(2\omega)} + i_{o(3\omega)} + i_{o(4\omega)} \quad (20)$$

where  $I_o$  is the dc component of total output current, and  $i_{o(\omega)}$ ,  $i_{o(2\omega)}$ ,  $i_{o(3\omega)}$ ,  $i_{o(4\omega)}$  are, respectively, the first-, second-, third-, and fourth-harmonic current components present in the total output current  $i_o$ . Detailed expressions of these components are

given as follows:

$$I_o = \frac{I_{\max} V_{\max}}{2V_o} = \frac{P_{ac\_avg}}{V_o} \quad (21)$$

$$i_{o(\omega)} = [kC_1 - (1-k)C_2] \left( -\frac{\omega V_{\max} V_d}{V_o} \cos(\omega t) + \frac{\omega B V_{\max}}{2V_o} \sin(\omega t + \varphi) \right) \quad (22)$$

$$i_{o(2\omega)} = -\frac{I_{\max} V_{\max}}{2V_o} \cos(2\omega t) - \frac{2\omega B(C_1 + C_2)V_d}{V_o} \times \cos(2\omega t + \varphi) - (k^2 C_1 + (1-k)^2 C_2) \frac{\omega V_{\max}^2}{2V_o} \times 2V_o \sin(2\omega t) \quad (23)$$

$$i_{o(3\omega)} = [kC_1 - (1-k)C_2] \left( -\frac{3\omega V_{\max} B}{2V_o} \sin(3\omega t + \varphi) \right) \quad (24)$$

$$i_{o(4\omega)} = \frac{\omega B^2(C_1 + C_2)}{V_o} \sin(4\omega t + 2\varphi) \quad (25)$$

where  $P_{ac\_avg}$  is the average value of the source power.

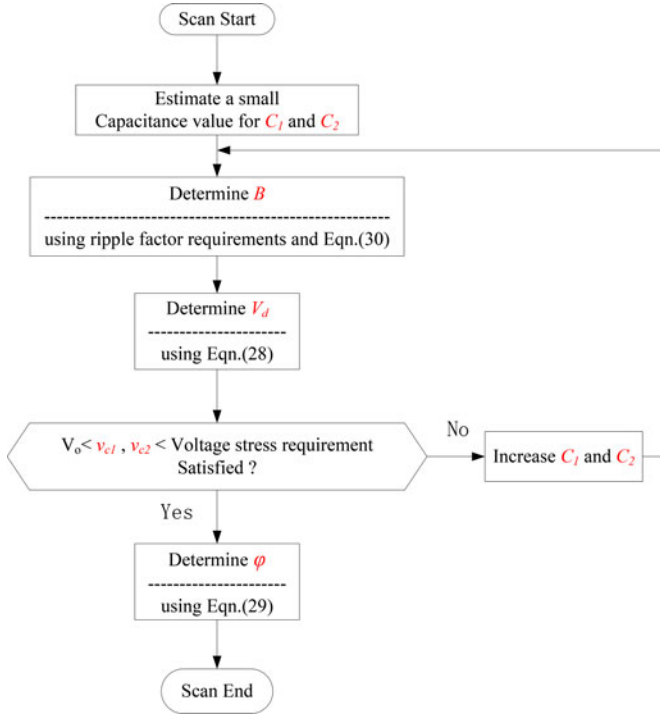


Fig. 10. Waveform control parameters determination flowchart using capacitor scanning.

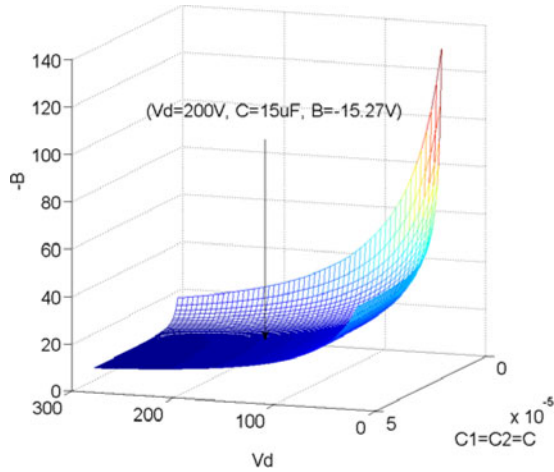


Fig. 11. Plot of  $C$ ,  $B$ , and  $V_d$  (where  $C = C_1 = C_2$ ) for a  $P_o = 50$  W design that is based on (33).

Careful examination of (22) and (24) shows that  $i_{o(\omega)}$  and  $i_{o(3\omega)}$  will always be zero as long as

$$kC_1 = (1 - k)C_2 \quad (26)$$

is satisfied.

Moreover, by equating  $i_{o(2\omega)}$  in (23) to zero and combining it with (26), it can be derived that  $i_{o(\omega)}$ ,  $i_{o(2\omega)}$  and  $i_{o(3\omega)}$  will be eliminated when (27) to (29) are complied

$$k = \frac{C_2}{C_1 + C_2} \quad (27)$$

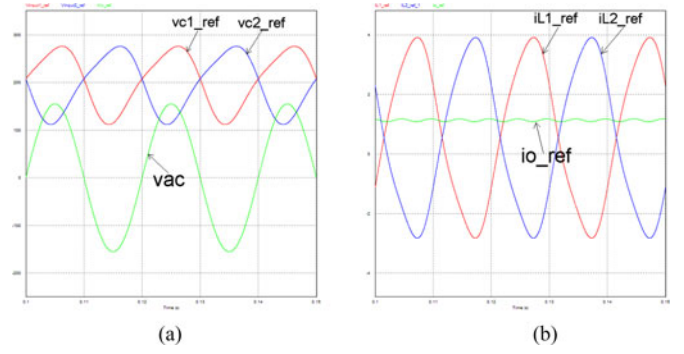


Fig. 12. Simulated reference voltage and current waveforms of the buck differential rectifier for  $C_1 = C_2 = 15 \mu\text{F}$ ,  $P_o = 50$  W,  $V_{in} = 110 \sin(\omega t)$ ,  $R_o = 39 \Omega$ .

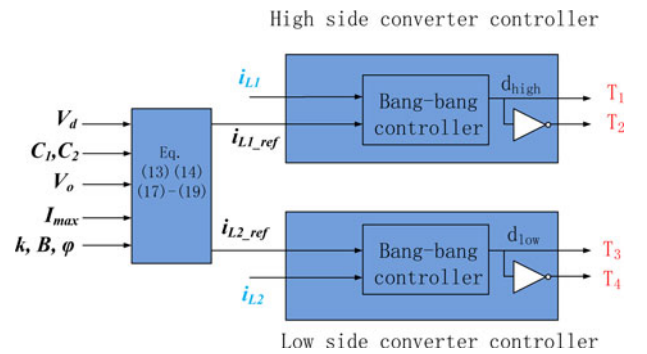


Fig. 13. Control blocks of the differential buck rectifier with a current bang-bang controller.

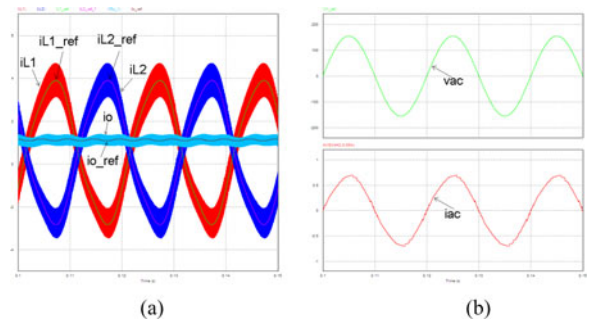


Fig. 14. Simulated operation waveforms of differential buck converter for  $V_d = 200$  V,  $P_o = 50$  W,  $R_o = 39 \Omega$ .

$$B = -\frac{V_{\max}}{4\omega(C_1 + C_2)V_d} \sqrt{I_{\max}^2 + (kC_1\omega V_{\max})^2} \quad (28)$$

$$\varphi = \frac{\pi}{2} - \sin^{-1} \frac{I_{\max}}{\sqrt{I_{\max}^2 + (kC_1\omega V_{\max})^2}} \quad (29)$$

The component  $i_{o(4\omega)}$  cannot be eliminated when only the double-line frequency component is introduced in the control of  $v_{C1}$  and  $v_{C2}$ . This component will eventually present as output current ripples. The ripple factor (which is the ratio of the current ripple amplitude over the dc averaged current) of output current can therefore be derived as shown in (30). Fig. 8 shows the calculated results of the ripple current amplitude  $I_{(4\omega)}$ , the

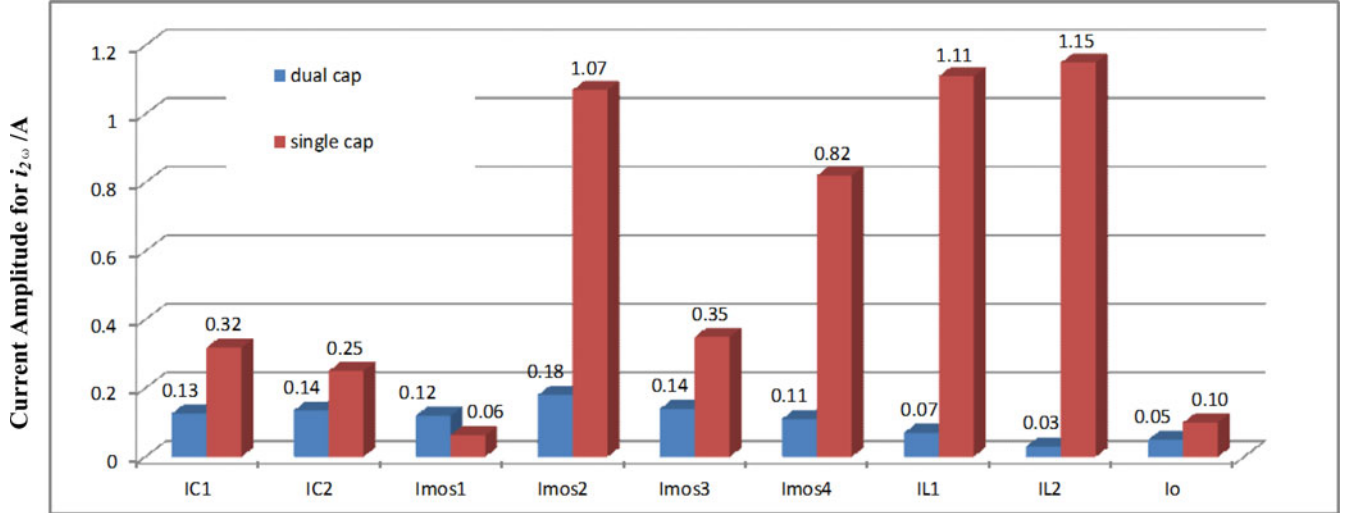


Fig. 15. Amplitude of double-line-frequency current component on the respective power devices for different combinations of  $C_1$  and  $C_2$  using waveform control (dual-cap case:  $C_1 = C_2 = 15 \mu\text{F}$ , single-cap case:  $C_1 = 0$ ,  $C_2 = 30 \mu\text{F}$ ).

averaged dc current  $I_o$ , and the ripple factor under the condition of  $V_{\text{max}} = 155.5 \text{ V}$ ,  $V_d = 200 \text{ V}$ ,  $C_1 = C_2 = 15 \mu\text{F}$ , and  $R_o = 39 \Omega$ . It can be seen from Fig. 8(b) that the ripple factor is low over a wide power range. A more complicated form of  $v_{C1}$  and  $v_{C2}$  would be required for the full elimination of all the low-frequency ripples. However, this is often unnecessary since such ripples are normally very low. Therefore, with the proposed inductor-current waveform control method, only a small non-E-Cap is required for the dc output of the circuit

$$\text{factor}_{\text{ripple}}(i_o) = \frac{I_{4\omega}}{I_o} = \frac{2\omega B^2(C_1 + C_2)}{I_{\text{max}} V_{\text{max}}}. \quad (30)$$

It must be emphasized that the inductor-current waveform control method is applicable to the general family of differentially connected rectification systems (e.g., full-bridge or half-bridge topologies) for eliminating the dc-link E-Cap. Considering that many existing rectification systems have already been implemented with the full-bridge topologies, one immediate advantage of the proposed method is that, without any major hardware modification to the existing systems, the prolonging of the system's lifetime is possible with only a slight change of the control methodology.

### B. Input Capacitors Optimization and Waveform Control Parameter Design

For the same system where the line voltage and current are the same, (27)–(29) indicate that a different combination of the input capacitors  $C_1$  and  $C_2$  will require different waveforms of  $v_{C1}$  and  $v_{C2}$ . From an efficiency viewpoint, the optimum selection of the input capacitors can be found by studying the flow path of the double-line frequency power component. Fig. 9 shows the three possible current flow paths of the double-line frequency power, i.e., either 1) through the capacitors, 2) the inductors, or, 3) both the inductors and capacitors. Since the inductors are usually more lossy than capacitors [19], it is preferable to have

TABLE II  
SPECIFICATIONS OF BUCK DIFFERENTIAL RECTIFIER

Input line parasitic inductance	3.67 $\mu\text{H}$
Input voltage $V_{\text{ac}}$ (RMS)	110 V
Output voltage $V_o$	43.6 V
Fundamental frequency $f$	50 Hz
Switch frequency $f_s$	around 50 kHz
Inductors ( $L_1, L_2$ )	600 $\mu\text{H}$ , 6 A
Capacitors ( $C_1, C_2$ )	Dual cap case: $C_1 = C_2 = 15 \mu\text{F}$ , 600 V, film cap Single cap case: $C_2 = 30 \mu\text{F}$ , 600 V, film cap

the double-line frequency power flowing through the capacitors instead of the inductors. When the flow path in Fig. 9(a) is realized, no double-line frequency power flows through the inductors.

As  $L_1$  and  $L_2$  are placed on the dc side, the study of the double-line frequency power in the inductors is equivalent to the study of their respective double-line frequency current. From (19) and by considering (27)–(29), the double-line frequency content in the inductor currents can be derived as

$$i_{L1(2\omega)} = \left( \frac{C_1 - C_2}{C_1 + C_2} \right) \left( \frac{V_{\text{max}} I_{\text{max}}}{2V_o} \cos(2\omega t) + \frac{\omega V_{\text{max}}^2 C_1 C_2 / (C_1 + C_2)}{2V_o} \sin(2\omega t) \right) \quad (31)$$

$$i_{L2(2\omega)} = - \left( \frac{C_1 - C_2}{C_1 + C_2} \right) \left( \frac{V_{\text{max}} I_{\text{max}}}{2V_o} \cos(2\omega t) + \frac{\omega V_{\text{max}}^2 C_1 C_2 / (C_1 + C_2)}{2V_o} \sin(2\omega t) \right). \quad (32)$$

It is evident from (31) and (32) that the double-line frequency current component in the inductors will be simultaneously zero if  $C_1$  and  $C_2$  are equal. Otherwise, the double-line frequency

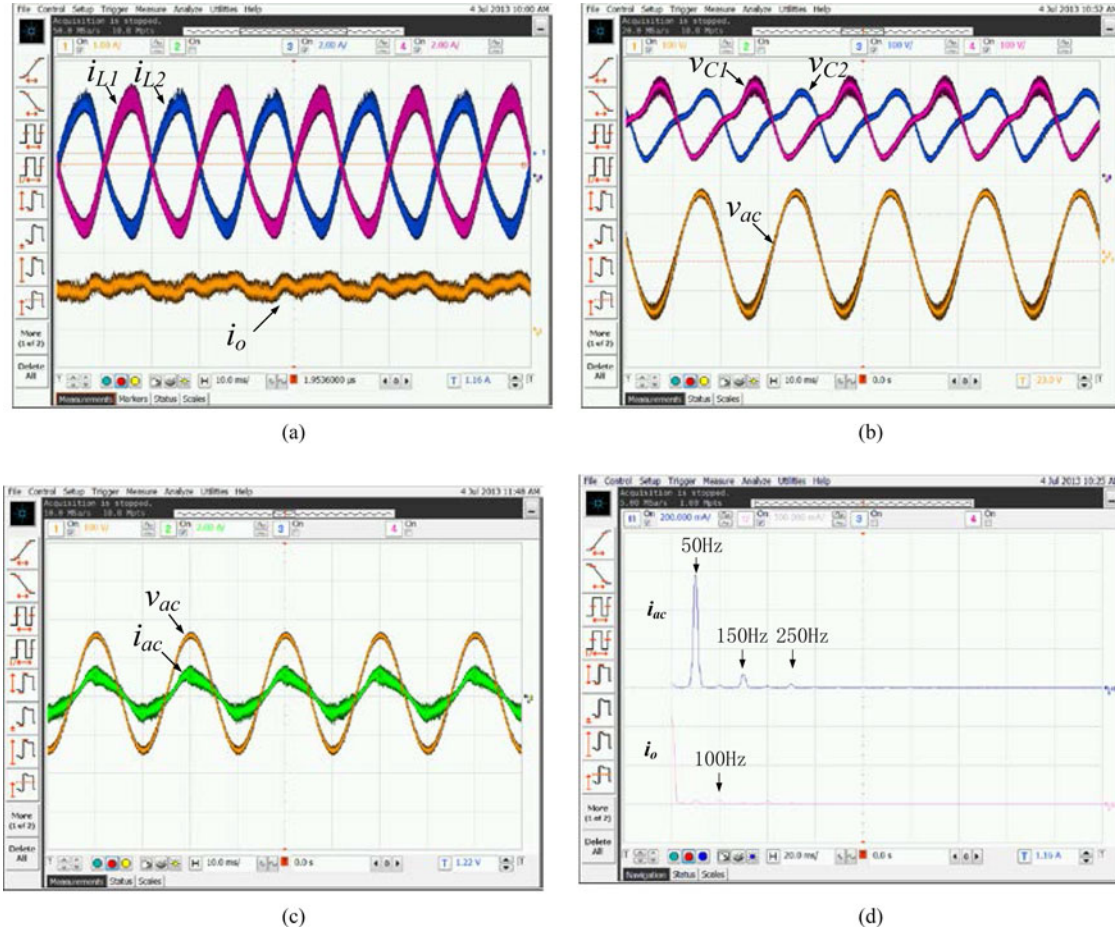


Fig. 16. Experimental results of differential buck converter with operating conditions  $P_o = 50$  W,  $R_o = 39$   $\Omega$ ,  $C_o = 0.47$   $\mu$ F for the dual-cap case of  $C_1 = C_2 = 15$   $\mu$ F. [(a) Inductor currents  $i_{L1}$  (pink),  $i_{L2}$  (blue) and output current  $i_o$  (orange), (b)  $v_{C1}$  (pink),  $v_{C2}$  (blue),  $v_{ac}$  (orange), (c)  $v_{ac}$  (orange),  $i_{ac}$  (green), (d) FFT analysis results for  $i_{ac}$  (orange) and for  $i_o$  (green). Scales:  $i_{L1}$ ,  $i_{L2}$ : 2 A/div,  $i_o$ : 1 A/div,  $v_{C1}$ ,  $v_{C2}$ ,  $v_{ac}$ : 100 V/div,  $i_{ac}$ : 2 A/div, FFT for  $i_{ac}$ : 200 mA/div, FFT for  $i_o$ : 500 mA/div.

current component will flow through the inductors. If  $C_1 = C_2 = C$ , then  $k$ ,  $B$ , and  $\varphi$  of (27)–(29) will be simplified as

$$k = \frac{1}{2} \quad (33)$$

$$B = -\frac{V_{\max}}{8V_d\omega C} \sqrt{I_{\max}^2 + \omega^2 C^2 V_{\max}^2 / 4} \quad (34)$$

$$\varphi = \sin^{-1} \frac{I_{\max}}{\sqrt{I_{\max}^2 + \omega^2 C^2 V_{\max}^2 / 4}} - \frac{\pi}{2}. \quad (35)$$

In this way, the double-line frequency power flow path described in Fig. 9(a) for reducing the conduction loss in the rectifier circuit will be realized, and a higher energy efficiency of the circuit can be obtained.

Once the optimal  $k$  parameter is determined, the optimal capacitance ratio is also derived, that is,  $C_1 = C_2$ . The capacitances of  $C_1$  and  $C_2$  should be selected in such a way that 1) the ripple factor defined in (30) satisfies the design criteria, e.g., less than 5%; 2) the capacitor voltages  $v_{C1}$  and  $v_{C2}$  in (13) and (14) are always higher than the output dc voltage  $V_o$  at all desired loading conditions, of which otherwise the two bidirectional buck converters cannot operate properly; 3) the peak value of

$v_{C1}$  and  $v_{C2}$  must be within acceptable range, e.g., within specified voltage stress range; and 4) the capacitance of  $C_1$  and  $C_2$  should be minimal such that the use of E-Cap in the circuit can be avoided.

Bearing in mind the above capacitor selection rules, the waveform control parameters can be designed through capacitor scanning using the flowchart shown in Fig. 10. First, estimate a small capacitor value for  $C_1$  and  $C_2$ . Then, the maximum waveform parameter  $B$  is derived based on the ripple factor requirements according to (30). With the derived  $B$  and by using (28), parameter  $V_d$  can be determined. The capacitor voltage must fall within the limits of the output voltage and satisfy the voltage stress requirement. Otherwise, the capacitance is increased, and then, the calculation is repeated until a suitable value of the capacitance is found.

### C. Simulation Verification

Based on (34), the parametric relationships of  $C$ ,  $B$ , and  $V_d$  for a buck differential rectifier is derived and plotted as shown in Fig. 11 (assuming that  $P_o = 50$  W,  $v_{ac} = 110\sqrt{2}\sin(2 * \pi * 50t)$  and  $C_1 = C_2 = C$ ). Using the above capacitance

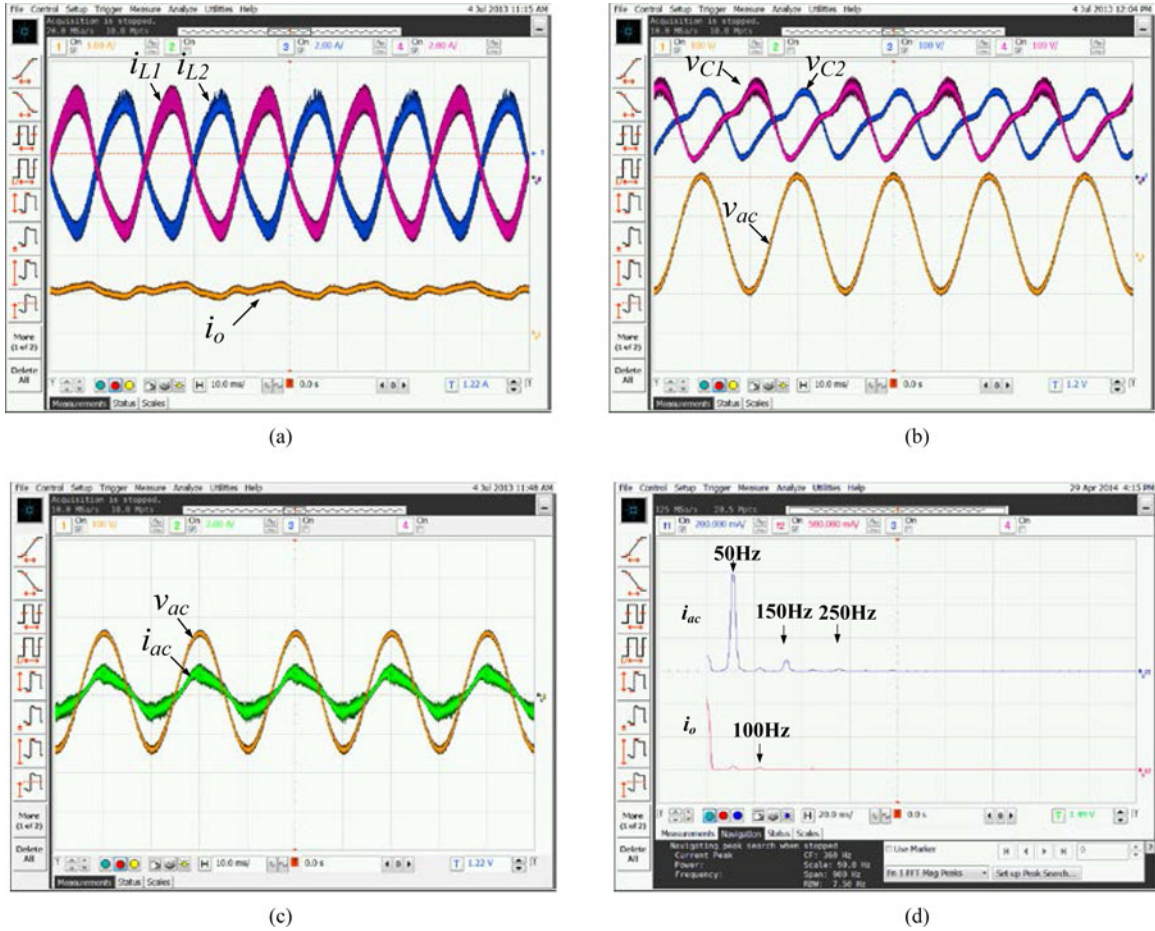


Fig. 17. Experimental waveforms of the differential buck converter with a resistive-capacitive load ( $P_o = 50$  W,  $R_o = 39 \Omega$ ,  $C_o = 47 \mu\text{F}$ , dual-cap case). [(a) inductor currents  $i_{L1}$  (pink),  $i_{L2}$  (blue) and output current  $i_o$  (orange). (b)  $v_{C1}$  (pink),  $v_{C2}$  (blue),  $v_{ac}$  (orange). (c)  $v_{ac}$  (orange),  $i_{ac}$  (green). (d) FFT analysis results for  $v_{ac}$  (orange) and for  $i_o$  (green). Scales:  $i_{L1}$ ,  $i_{L2}$ : 2 A/div,  $i_o$ : 1 A/div,  $v_{C1}$ ,  $v_{C2}$ ,  $v_{ac}$ : 100 V/div,  $i_{ac}$ : 2 A/div, FFT for  $v_{ac}$ : 200 mA/div, FFT for  $i_o$ : 500 mA/div]

selection rule and waveform control parameter design flowchart, the following parameters are chosen and applied in this study:  $C_1 = C_2 = C = 15 \mu\text{F}$ ,  $k = 0.5$ ,  $V_d = 200$  V,  $B = -15.27$  V, and  $\varphi = 0.52$  rad. The voltages across  $C_1$  and  $C_2$  are, respectively

$$v_{c1} = 200 + 77.78 \sin(\omega t) - 15.27 \sin(2\omega t - 0.52) \quad (36)$$

$$v_{c2} = 200 - 77.78 \sin(\omega t) - 15.27 \sin(2\omega t - 0.52). \quad (37)$$

Similarly, the inductor currents and output current can be derived from (19) and (20). The simulated reference and expected waveforms of  $v_{C1}$ ,  $v_{C2}$ ,  $i_{L1}$ ,  $i_{L2}$ , and  $I_o$  for operating the rectifier with the waveform control are recorded in Fig. 12.

From Fig. 12(a), it is shown that the ac content of  $v_{C1}$  and  $v_{C2}$  are clearly of a distorted form, but their differential value is still the ac line voltage. From Fig. 12(b), it is shown that the low-frequency ac components of  $i_{L1}$  and  $i_{L2}$  cancel out with one another, leading to the output current  $i_o$  containing only a small portion of the low-frequency content. The amplitude of the ripple current  $i_o$  is 4.43% (0.05 A) of the average dc current (1.13 A) and is in good agreement with the calculated

results shown in Fig. 8. This is in contrast to  $i_o$  that is achievable by traditional means without waveform control [refer to (12)], where the ripple amplitude is 100% (1.13 A) of the average dc current. The low-frequency ripple has been suppressed almost 23 times with the inclusion of waveform control.

Given the inductor current references in Fig. 12, two simple bang-bang controllers are then employed for regulating the inductor currents. The control blocks are shown in Fig. 13, and the simulated waveforms of the differential buck rectifier are shown in Fig. 14. As illustrated in Fig. 14(a),  $i_{L1}$  and  $i_{L2}$  are capable of tracking their respective references. The simulated output current ripple amplitude of  $i_o$  is 0.3 A, which is 27% of the average dc current of 1.13 A. It should be noted that no output capacitor is used in the system, and most of the presented ripples are actually of high-frequency content, which can be easily filtered out using an output high-frequency filter if desired. Fig. 14(b) shows the input voltage and current waveforms. A sinusoidal ac current in phase with the input line voltage is observed.

The flow of the double-line frequency current components through all devices is analyzed. The results are presented in Fig. 15. It is clear that when  $C_1 = C_2$ , the double-line frequency current flows mainly through  $C_1$  and  $C_2$ , and  $T_1$ - $T_4$ .

A small amount of  $i_{L1(2\omega)}$  and  $i_{L2(2\omega)}$  still exists due to control imperfection.

A different set of simulation with  $C_1 = 0$ ,  $C_2 = 30 \mu\text{F}$  (known as single-cap case) is conducted to demonstrate the effectiveness of optimizing the capacitor combination. From (27)–(29), the waveform parameters are determined as  $k = 1$ ,  $B = -13.26 \text{ V}$ ,  $V_d = 200 \text{ V}$ , and  $\varphi = 0$ . The associated double-line frequency currents are presented in Fig. 15. The double-line frequency current flows through every device but with a higher amplitude in  $C_1$ ,  $C_2$ ,  $T_2$ ,  $T_3$ ,  $T_4$ ,  $L_1$ , and  $L_2$  than those of the dual-cap case. Higher values of  $i_{L1(2\omega)}$  and  $i_{L2(2\omega)}$  imply more double-line frequency power flowing through the inductors and therefore a higher power loss and lower energy efficiency of the circuit.

#### IV. EXPERIMENTAL VALIDATION

##### A. Experimental Setup

A prototype of the buck differential rectifier with the waveform control and conduction loss reduction is constructed and tested. The design of the rectifier which comprises two buck converter follows that of conventional PFC buck converter design principles, but with a smaller input voltage range. To avoid obscuring the essential of this paper which is on the waveform control method, the design details of the differential rectifier will be omitted. Table II gives the rectifier's specifications. LABVIEW is used to calculate and generate the desired inductor current reference signals according to (19) based on the optimal values of  $k$ ,  $B$ , and  $\varphi$ . Two hysteresis controllers are used to independently control  $i_{L1}$  and  $i_{L2}$  such that the respective references are tracked. It will be shown that with the proposed waveform control method, a direct ac/dc differential rectifier with mitigated double-line frequency ripple on the dc side can be realized. It must be emphasized that no E-cap is used in the experimental setup. For the dual-cap experiment ( $C_1 = C_2 = 15 \mu\text{F}$ ), the adopted parameters are  $k = 0.5$ ,  $V_d = 200 \text{ V}$ , and  $B = -15.27 \text{ V}$ , and  $\varphi = 0.52$ . For single-cap experiment ( $C_1 = 0$ ,  $C_2 = 30 \mu\text{F}$ ), the parameters are  $k = 1$ ,  $V_d = 200 \text{ V}$ ,  $B = -13.26 \text{ V}$ , and  $\varphi = 0$ .

##### B. With Resistive and Resistive-Capacitive Load (Dual-Cap Case)

Fig. 16(a)–(c) shows the measured voltage and current waveforms of the proposed buck differential rectifier in a dual-cap configuration, with the waveform control and operating at 50-W output, with a pure resistive load of  $R_o = 39 \Omega$ . In this experiment,  $R_o$  is connected in parallel with a very small capacitor ( $C_o = 0.47 \mu\text{F}$ ) to filter the high-frequency switching ripple. The capacitor does not affect the dominant resistive property of the load.

Fig. 16(a) shows the waveforms of  $i_{L1}$  and  $i_{L2}$ , and the output current  $i_o$ , which is the sum of  $i_{L1}$  and  $i_{L2}$ . With the proposed waveform control, the low-frequency current ripple at the dc output is substantially mitigated. The amplitude of output current ripple is 0.27 A (23% of average (dc) output current 1.15 A). The result is similar with that obtained in simulation (27%).

TABLE III  
RMS CURRENT IN COMPONENTS FOR DIFFERENT  $C_1$  AND  $C_2$  COMBINATIONS ( $P_o = 50 \text{ W}$ )

	$L_1$	$L_2$	$T_1$	$T_2$	$T_3$	$T_4$	$C_1$	$C_2$
Dual-cap case (A)	2.43	2.43	1.15	2.14	1.15	2.14	0.52	0.52
Single-cap case (A)	2.57	2.23	1.07	2.34	1.1	1.9	0	0.91

TABLE IV  
CURRENT STRESS ON POWER DEVICES FOR DIFFERENT  $C_1$  AND  $C_2$  COMBINATIONS ( $P_o = 50 \text{ W}$ )

	$T_1$	$T_2$	$T_3$	$T_4$
Dual-cap case (A)	4.68	4.71	4.68	4.71
Single-cap case (A)	6.06	6.1	4.7	4.7

Fig. 16(b) shows the waveforms of  $v_{C1}$ ,  $v_{C2}$ , and  $v_{ac}$ . Since no direct capacitor voltage regulation is employed, the dc component of  $v_{C1}$  and  $v_{C2}$  presents some deviation from the expected value of  $V_d$ , which is attributed by the power losses of the converters. The results will be improved if a voltage regulation controller is applied, by which means the averaging voltage of  $v_{C1}$  and  $v_{C2}$ , that is  $V_d$ , can be regulated. Fig. 16(c) and (d) shows the waveforms of  $v_{ac}$ ,  $i_{ac}$ , and the FFT spectrum of the line ( $i_{ac}$ ) and the output current ( $i_o$ ), respectively. It is shown that a good PF with low-harmonic contents is achieved. Line current contains negligible (2.69%) 100-Hz (second harmonic) current component and 11.9% of the third-harmonic current component. The measured total harmonic distortion of the input current is 10.45% with a PF of 0.97. These measurements are within the limit of Class C regulation. It should be noted that only a small input filter is included in the circuit since  $C_1$  and  $C_2$  are adequate to filter the high-frequency content in  $i_{ac}$ . For output current  $i_o$ , the 100 Hz content is only 4.2% of the average current.

Fig. 17 shows the operation waveforms of the rectifier with a resistive-capacitive load ( $R_o = 39 \Omega$ ,  $C_o = 47 \mu\text{F}$ ). Compared with the results given in Fig. 16, similar operating waveforms have been captured. With increased output capacitance, the output current ripple, especially the high-frequency components, has been further suppressed.

##### C. Comparative Study of Different Input Capacitors Combination

Another set of experiment with  $C_1 = 0$  and  $C_2 = 30 \mu\text{F}$  (single-cap case) is carried out to validate the result of input capacitor optimization. Table III illustrates the measured RMS current of the inductors  $L_1$  and  $L_2$ , capacitors  $C_1$  and  $C_2$ , and the four active switches with an output power of 50 W. It can be seen that  $i_{L1}$  and  $i_{L2}$  in the single-cap case are 2.57 and 2.23 A, respectively; while they are both 2.43 A in the dual-cap case.

Table IV shows the current stress of various switches for both cases. Theoretically, the current stress of  $T_1$  and  $T_2$ , as well as  $T_3$  and  $T_4$  should be the same. The tiny current stress differences in Table IV are mainly caused by measurement inaccuracy. It is

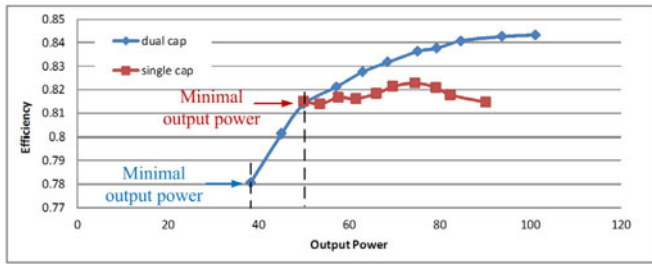


Fig. 18. Efficiency comparison between dual-cap case and single-cap case.

found that the single-cap case has a higher current stress on  $T_1$  and  $T_2$  than the dual-cap case. In the dual-cap case, the current stress has been suppressed by more than 22.7% (1.39 A). Also, the stress is more evenly distributed among all switches than in the single-cap case.

Finally, the system efficiency curves of both cases are compared and are shown in Fig. 18. It is observed that when the output power level is higher than 50 W, the dual-cap case has a higher energy efficiency than the single-cap case. The improvement in efficiency increases as the power level goes higher. When the output power is below 50 W, it is found that the single-cap case cannot function properly since there are instances when the input capacitor voltages  $v_{c1}$  and  $v_{c2}$  are lower than the expected output voltage, as have been discussed in Section III-B. Different capacitance and waveform parameters must be selected for the differential rectifier to operate below 50 W. For a detailed design flowchart, refer to Section III-B. As for dual-cap case, however, a wider operation range has been observed.

## V. CONCLUSION

A type of E-cap-less ac/dc differential rectifier system with current-waveform control technique is proposed in this paper for direct rectification applications. The configuration is based on a structure of having two bidirectional dc/dc converters in series-input and parallel-output connections. To complement the feature of this rectifier configuration, waveform control is applied to mitigate the double-line frequency current ripple of the dc output, thereby allowing the removal of the electrolytic storage capacitor that is typically required in such systems. Mathematical derivations and analysis are provided in the paper to validate the concept, using the buck differential rectifier as a case study example. Moreover, the conduction loss of double-line frequency power flow can be minimized through parameter optimization. Experimental results show that the idea is feasible and workable for this configuration. The idea of the direct ac/dc differential rectifier is extendable to other combination of converters, connected differentially, with or without galvanic isolation, for single phase or multiphase applications.

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