A High Performance Interleaved Bridgeless PFC for Nano-grid Systems

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Abstract – A high performance interleaved bridgeless boost power factor correction (PFC) rectifier operating under the critical current conduction mode (CrM) is proposed in this paper to improve the efficiency and system performance of various applications, such as nano-grid systems. By combining the interleaved technique with the bridgeless topology, the circuit contains two independent branches without rectifier diodes. The branches operate in interleaved mode for each respective half-line period. Moreover, when operating in CrM, all the power switches take on soft-switching, thereby reducing switching losses and raising system efficiency. In addition, the input current flows through a minimum amount of power devices. By employing a commercial PFC controller, an effective control scheme is used for the proposed circuit. The operating principle of the proposed circuit is presented, and the design considerations are also demonstrated. Simulations and experiments have been carried out to evaluate theoretical analysis and feasibility of the proposed circuit.

Keywords: PFC, Bridgeless technique, Interleaved topology, Critical conduction mode, Peak current mode control

1. Introduction

With the rapid development of power electronics technology, direct current nano-grid (DCNG) - a DC network that is interconnected with the AC main grid - becomes a good solution for energy management and utilization. It has been reported that using the 400V DC grid improves total energy efficiency by approximately 18% [1]. Power factor correction (PFC) interface can be used to improve the power quality of the AC main grid and integrate DCNG with the existing electrical power system [2, 3].

However, conventional active PFC regulators are generally comprised of a front- end bridge diode and a high frequency boost converter. In the conventional boost PFC circuit, current passes through two slow diodes and one switch during the on-state, while during the off-state through two slow diodes and one fast diode [5-7]. Since typical DCNG system has relatively high power capacity, the high conduction losses caused by the forward voltage drop of the bridge diode and the high switching losses of switches (MOSFET and fast- recovery diode) begin to dramatically degrade overall system efficiency [8-10]. In addition, the heat generated within the semiconductors and inductors can destroy power devices and decreases system reliability [11].

To maximize the conversion efficiency, a lot of research

have been investigated on the attractive techniques on PFC topologies. Two of them are the bridgeless topology and the interleaved technique.

To reduce the conduction losses, the rectifier diode and the boost converter are combined to develop the bridgeless PFC [12-14]. When compared with conventional circuits, this circuit could decrease the conduction losses without input rectifier bridge by allowing the current to flow through a minimum number of switching devices. Dual-boost bridgeless PFC, shown in Fig. 1, is one of the most well-developed bridgeless topologies. In this circuit, two boost circuits operate for two half-line periods, respectively. The working principle of this converter is symmetrical within the positive and negative half-line cycles of the input voltage [7]. In this way, the limitation on power capacity that is caused by the diode bridge is eliminated. In addition, when compared with a single inductor, two inductors can achieve higher thermal performance.

However, the dual-boost bridgeless PFC is still not suitable for high- power DCNG applications. Although there are two boost phases in dual-boost topology, the operation of the circuit is still under single-phase mode. When one phase is operating, another one is in idle state. Single-phase boost converters under high power conditions have significant disadvantages, such as high current ripple,

Fig. 1 Basic bridgeless boost PFC rectifier

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Fig. 2. Basic interleaved boost PFC circuit

high current stress for switches, large inductor size, hard switching operation, and reverse-recovery problems, leading to low efficiency and low energy density. Therefore, more phases should be added to increase the power capacity of the dual-boost PFC. However, direct paralleling more phases can lead low utilization problem of semiconductors.

Another effective technique is known as interleaved PFC, which usually combines more than two phases in parallel to be used in high power applications [15]. A conventional interleaved PFC topology is shown in Fig. 2. Owe to the interleaved operation, lower current ripple and smaller harmonic contents can be obtained for the input and output currents [16]. As a result, the size of the inductors and EMI filter can be decreased, as well as the switching losses.

However, this topology has the same drawbacks as the conventional boost PFC [3,17], such as the high conduction losses caused by the forward voltage drop of the rectifier diode and poor thermal performance, thereby decreasing system efficiency and reliability. To improve the performance of the interleaved PFC, the high conduction losses of the rectifier diode should be firstly considered.

In an attempt to develop a high-efficiency topology that is suitable for high power DCNG applications, this paper proposes a novel boost PFC rectifier that combines the bridgeless topology with the interleaved technique. The circuit contains two interleaved boost phases, two more inductors and one more power switch. Since the two interleaved branches are operated out of phase, lower conduction losses, lower current ripple, and smaller component size are achieved. When it is operated under critical current conduction mode (CrM), the rectifier is able to turn on the switches at zero current to reduce switching losses and to increase efficiency.

This paper mainly contains four sections. Section II describes the configuration and the operation principle of the proposed circuit. Section III proposed the key design considerations of the circuit. A comparison study with conventional topologies is also presented in this section. Section IV illustrates the simulation and experimental results. Finally, a conclusion is given in Section V.

2. Configuration and Operation Principle

2.1 Circuit configuration

The proposed circuit is shown in Fig. 3. The proposed

Fig. 3. The proposed PFC circuit

circuit is composed mainly of two interleaved boost circuits, named as branch I and branch II. Each branch is composed of two phases converters that are parallel connected. Branch I consists of phase I₁ and phase I₂, where I₁ is composed of L₁, S_A, and D₁, I₂ is composed of L₂, D₆, S_C, and D₂. Branch II consists of phase II₁ and phase II₂, where II₁ is composed of L₄, S_B, and D₄, II₂ is composed of L₃, D₅, S_C, and D₃. D_L and D_N denote rectifier diodes, while D₁ ~ D₄ are fast-recovery diodes. L₁ ~ L₄ are boost inductors, and C_O is output capacitor. R_L denotes the equivalent resistive load.

In the proposed circuit, S_C is the common switch shared by the interleaved branches. The current of each branch can be blocked by D₅ and D₆. Either S_A or S_B works in active state while another switch operates as a freewheeling diode during each half-line period of the line voltage, whereas S_C is active for the whole-line period. Since the frequency of the line voltage is low (60Hz or 50Hz), slow-recovery diodes are feasible for D₅ and D₆.

To improve the common-mode EMI performance, branch I operates through D_N during positive half-line cycle, while branch II operates through D_L during negative half-line cycle. Although two additional inductors are employed when compared with conventional interleaved topologies, the proposed circuit can achieve good thermal performance. On the other hand, the two boost inductors of each branch can be coupled on one magnetic core, with considerable size and cost reduction resulting.

As can be seen from Fig. 3, the proposed circuit contains one more MOSFET than the dual-boost PFC. In addition, it requires two more slow diodes and two more fast diodes.

However, a much higher output power can be achieved by using the proposed circuit, while the dual-boost topology is not suitable for high-power applications.

2.2 Circuit Operation Principle and Analysis

The operation circuits of the proposed converter on the positive and negative half-line cycles are shown in Fig. 4(a) and Fig. 4(b), respectively.

In Fig. 4(a), on the positive half-line cycle, phase I₁ (L₁-S_A-D₁) and phase I₂ (L₂-D₆-S_C-D₂) are active alternatively.

When S_A is turned on, the current flows through L₁-S_A-D_N path and the energy is stored in L₁. When S_A is turned off, the current flows through L₁-D₁-D_N path, delivering the energy to the load. Likewise, phase I₂ (L₂-

$D_6-S_C-D_2$) operates with the same principle in the interleaved mode. When S_C is switched on, the current flows through $L_2-S_C-D_6$, and D_N . When S_C is turned off, current flows through L_2 , D_2 , and D_N . During this half-line cycle the boost phases II_1 ($L_4-S_B-D_4$) and phase II_2 ($L_3-D_5-S_C-D_4$) are in an idle state.

The same results can be obtained during the negative half-line cycle. As shown in Fig. 4(b), when S_B is switched on, energy is stored in L_4 through S_B-D_L . When S_B is switched off, energy is released through $L_4-D_4-D_L$. As the interleaved operation, phase II_2 ($L_3-D_5-S_C-D_4$) operates in the same mode with out of phase.

According to the operation principle illustrated in Fig. 4, for branch I_1 and branch II_1 , there is one slow diode, one fast diode, and one active power switch in the current flowing path for one phase, while one more slow diode, D_5 for branch I_2 , and D_6 for branch II_2 , are employed, resulting reduced amount of conduction devices comparing with traditional interleaved PFC circuit. As a result, the heat stresses as well as the conduction losses on the power semiconductors can be reduced, thus, the power convention efficiency can be improved.

It should be noted that S_A and S_B are driven through the same PWM signal, and S_C is driven through the 180° interleaved PWM signal. Therefore, the proposed circuit can operate the same with the conventional interleaved PFC. This control scheme simplifies the configuration of controller and can be implemented through employing standard commercial interleaved PFC controllers.

Fig. 4. Operating principle of the proposed PFC. (a) In positive half-line cycle and (b) in negative half-line cycle of the line voltage

Every interleaved phase of the proposed circuit can be operated under discontinuous current conduction mode or continuous current conduction mode. However, there are several advantages when operating under CrM. For example, the proposed circuit can be operated under zero current switching (ZCS) on the switch turn-on transition. The reverse-recovery issue of fast-recovery diodes can be significantly reduced. As a result, low current ripple and low switching losses can increase the circuit performance and reduce the EMI noise compared with conventional bridgeless PFC shown in Fig.1. The proposed circuit has another advantage of the reduced current stress for the semiconductor switches.

3. Design Consideration and Comparison Study

3.1 Circuit design and comparison study

To design a practical circuit, it is necessary to calculate the key parameters of the power components. During the analysis and evaluation, it is assumed that the unity power factor is achieved.

The input voltage can be described as

$$v_{in}(t) = \sqrt{2}V_{in} \sin(\omega t) \quad (1)$$

where V_{in} is the input RMS voltage. According to the voltage-second balance, we can get

$$i_L(t) = \frac{v_{in}(t)t_{on}}{L} = \frac{(V_o - v_{in}(t))t_{off}}{L} \quad (2)$$

where t_{on} and t_{off} are the on and off periods of the main switch; and V_o is the output voltage of the PFC circuit.

By operating in CrM and taking a constant switching on-time into consideration, the instantaneous switching frequency can be obtained as

$$f_s(t) = \frac{V_o - v_{in}(t)}{V_o t_{ON}} \quad (3)$$

The inductor current should reach zero at the end of the switching period, which can be determined according to the ripple current and input voltage conditions. As the minimum switching frequency should be selected under the low line input voltage, the inductor value is obtained as

$$L = \frac{\eta [v_{in}(t)]_{min}^2 (V_o - \sqrt{2}[(v_{in}(t))]_{min})}{V_o P_o [f_s(t)]_{min}} \quad (4)$$

where $[v_{in}(t)]_{min}$ is the minimum RMS value of the input voltage, $[f_s(t)]_{min}$ is the minimum switching frequency, P_o is the rating output power, and η is the expected convention efficiency.

As a result, the inductor peak current for one boost phase is expressed by

$$I_{Lpk} = \frac{\sqrt{2}P_{out}}{\eta[v_{in}(t)]_{min}} \quad (5)$$

and the inductor RMS current of the boost inductor is

$$I_{Lrms} = \sqrt{\frac{1}{2\pi} \int_0^{2\pi} (i_L(t))^2 dt} = \frac{P_{out}}{\sqrt{6}\eta[v_{in}(t)]_{min}} \quad (6)$$

Every boost circuit transports half of the output power for the whole circuit. As the result, the root mean square (RMS) current of the boost diode can be derived as

$$I_{Drms} = \frac{I_{Lpk}}{2\sqrt{2}} \sqrt{\frac{4\sqrt{2}[v_{in}(t)]_{min}}{9\pi V_o}} \quad (7)$$

Because of the CrM operation, reverse-recovery issues of the fast-recovery diodes can be reduced. As a result, only conduction losses should be considered.

As described in Section II, S_A and S_B work for the positive and negative line cycle, respectively, while S_C is active for the whole line cycle. Therefore, the RMS currents flowing through S_A and S_B can be given by

$$I_{rms_{AB}} = \frac{I_{Lpk}}{2\sqrt{2}} \sqrt{\frac{1}{6} - \frac{4\sqrt{2}[v_{in}(t)]_{min}}{9\pi V_o}} \quad (8)$$

while the RMS current of S_C is obtained as

$$I_{rms_C} = \frac{I_{Lpk}}{2} \sqrt{\frac{1}{6} - \frac{4\sqrt{2}[v_{in}(t)]_{min}}{9\pi V_o}} \quad (9)$$

One of the most essential advantages of the dual-boost bridgeless PFC converter is the reduced conduction losses caused by the rectifier diode bridge. As shown in Fig. 1, the conduction losses of the slow diodes in dual-boost PFC can be calculated as

$$P_{C_DualBoost} = \frac{2\sqrt{2}P_o}{\eta\pi[v_{in}(t)]_{min}} V_F \quad (10)$$

where V_F is the forward voltage of the slow diodes. As there is one more slow diode during the switch on-state in each second boost phase, the conduction losses for the phases can be obtained as

$$P_{C_Proposed} = P_{C_DualBoost} \sqrt{1 + \epsilon_D}, \quad (11)$$

where

$$\epsilon_D = \frac{\pi^2}{32} \left(\frac{1}{6} - \frac{1}{6} - \frac{4\sqrt{2}[v_{in}(t)]_{min}}{9\pi V_o} \right). \quad (12)$$

Assuming the output voltage is 400 V, in the universal input applications where the minimum input voltage is 85 V, the maximum value of ϵ_D is calculated as 0.038 from the above equation. As a result,

$$P_{C_Proposed} = \sqrt{1 + 0.038} P_{C_DualBoost} \approx 1.019 P_{C_DualBoost} \quad (13)$$

Therefore, although there is one more slow diode in the current path than the dual-boost bridgeless PFC circuit shown in Fig.1, the conduction losses of the proposed circuit is almost the same as those of the dual boost PFC circuit.

Based on the analysis above, a comparison among of the proposed circuit, conventional interleaved boost PFC, and dual-boost PFC under CrM operation is summarized in Table 1. It can be seen that the RMS inductor current of the proposed PFC is smaller than the others, which indicates that smaller inductor volume can be used in the circuit. In addition, current stresses of the boost diodes and the main switches of the proposed circuit are lower than those in the other circuits. Therefore, less-ideal and lower cost semiconductors can be used in the proposed circuit.

As the proposed circuit contains fewer power devices in the input current path than the conventional circuits, the conduction losses caused by slow diodes in the proposed

Table 1. Comparison of the conventional and the proposed circuits under CrM operation

Item	Interleaved CrM PFC	CrM Dual-boost PFC	Proposed PFC
Inductor RMS current	$\sqrt{6}I_{Lrms}$	$\sqrt{2}I_{Lrms}$	I_{Lrms}
Inductor value	$L/2$	L	L
Boost diode RMS current	$\sqrt{2}I_{Drms}$	$2I_{Drms}$	I_{Drms}
Slow diode conduction loss	$2P_{C_DualBoost}$	$P_{C_DualBoost}$	$P_{C_DualBoost} \sqrt{1 + \epsilon_D}$
MOSFET RMS current	$\sqrt{2}I_{rms_AB}$	$4I_{rms_AB}$	$S_A, S_B : I_{rms_AB},$ $S_C : \sqrt{2}I_{rms_AB},$
Power capacity	High	Medium	High
Current path (switch on-state)	2 slow diodes, 1 MOSFET	1 slow diodes, 1 MOSFET	$S_A, S_B : 1$ slow diodes, 1 MOSFET $S_C : 2$ slow diodes, 1 MOSFET
Current path (switch off-state)	2 slow diodes, 1 fast diode	1 slow diodes, 1 fast diode	1 slow diodes, 1 fast diode

Fig. 5. Block diagram of the control scheme of the proposed circuit

circuit are much lower than those of the conventional circuits. In addition, the conduction losses in the proposed circuit are similar to those of the dual-boost bridgeless PFC.

As the proposed circuit is built through connecting two boost converters, with each working under interleaved technique, the proposed PFC remains the outstanding features of the bridgeless PFC circuits and the interleaved topologies. This performance can result in lower conduction losses and lower current ripple.

Although there seems more semiconductor devices in the proposed topology, higher efficiency, output power, and thermal management performance can be obtained when compared with the other two topologies under the same power capacity. In addition, because of the lower power stresses of the components, lower component cost can also be achieved with a proper design.

3.2 Control scheme consideration

Peak current mode control (PCMC) method is extensively employed in power converters and has many advantages, such as easy design and fast response. The simplified proposed circuit based on PCMC is shown in Fig. 5.

Fig. 5. includes the signal feedback module, the PWM module, and the phase management module. As shown in Fig. 5, the current loop of the controller is inside of the voltage loop. The input current is controlled by the current loop to follow the input voltage waveform. The voltage loop controller regulates output voltage under CrM, while the switching frequency varies with constant on-time. The current loop and the voltage loop operate together by sampling the current is and output voltage V_o , respectively.

The zero current detection (ZCD) modules, shown as ZCDA and ZCDB, sample the zero current point of the

Fig. 6. Simplified zero current detection circuit of the proposed circuit

inductor currents $i_{L1} \sim i_{L4}$, which are employed as the off command to PWM signals. As a result, S_A , S_B , and S_C are turned on once the valley of inductor currents are detected.

For branch I and branch II, the two phases operate independently. Since all the phases are operated under CrM, the ZCD of the inductor currents becomes very important. An sample and effective ZCD circuit is designed to detect the valley of the inductor for the two phases of the branches. One of the ZCD circuits is shown in Fig. 6.

As shown in the circuit, two inductor current sensing circuits are connected in parallel. L_{sen1} senses the current of L_1 during the positive half-line period as i_{Ls1} . Meanwhile, i_{Ls4} is zero and taken to ZCDA through D_{S3} . In the same way, during the negative half-line period, i_{Ls4} is segregated from ZCDA by D_{S3} while i_{Ls1} is taken through D_{S1} . Therefore, we can get the voltage of ZCDA, v_{ZCDA} , as

$$v_{ZCDA} = \begin{cases} i_{Ls1} R_{S2}, & \text{During positive cycle} \\ i_{Ls4} R_{S4}, & \text{During negative cycle} \end{cases} \quad (13)$$

For the ZCDB circuit, the same results can be obtained. It should be noted that, although there are four channels in the proposed circuit and two ZCD channels in the PFC controller, the ZCD module can capture the valley of the inductor current effectively, making it easy to control the proposed circuit using commercial PFC controllers.

4. Simulation and Experimental Results

4.1 Simulation results

To evaluate the feasibility of the proposed circuit, a simulation circuit based on Pspice was developed. The model was designed according to the specifications shown in Table 2. A commercial PFC controller, UCC28063 from TI, was employed. Fig. 7 illustrates the main power stage of the simulation circuit.

Fig. 8 presents the switching waveforms of the branch I under steady-state. As can be seen, the S_A and S_C are turned on under ZCS conditions. Thus, the proposed circuit can achieve low switching losses and low reverse-recovery

Table 2. Parameters of the simulation circuit

Items	Symbols	Values
Input line frequency	f_L	60 Hz
Input line voltage	V_{ac}	85V - 265V
Rated power	P_{out}	600 W
Output voltage	V_O	385 VDC
Inductor	$L_1 \sim L_4$	80 μ H
Output capacitor	C_{out}	500 μ F

Fig. 7. Simulated power stage of the proposed circuit based on Pspice

Fig. 8. Waveforms of drain-to-source voltage, driver signal, and seitching current of S_A and S_C

losses.

The currents of L_1 and L_2 are shown in Fig. 9, as well as the driver signals of S_A and S_C . It can be seen that the effective frequency of ripple current is increased twice. The peak value of the output current ripple is also reduced half compared to the inductor current.

Fig. 10 illustrates the current waveforms of L_1 and L_2 in branch I under 220 V_{AC} line input and rated-load conditions. The switching waveforms of S_A and S_C are also presents in Fig. 10. As analyzed in the operation principle, S_A operates in half-line period and S_C is active during the whole line periods.

4.2 Experimental results

A 600 W experimental prototype circuit was built and

Fig. 9. Waveforms of the current waveforms of L_1 and L_2 as well as output current.

Fig. 10. Waveforms of inductor current waveforms of branch I, switching waveforms of S_A and S_C , and the input current waveform

Fig. 11. Photograph of the prototype converter

tested for a laboratory DCNG system, as shown in Fig. 11. The specifications of the circuit are the same as the parameters of the simulated circuit shown in Table 2. In the future work, higher power prototype circuit will be designed to test detailed performance.

The switching voltages of the switches S_A and S_B , V_{ds_SA} and V_{ds_SB} , are shown in Fig. 12, respectively. The current of L_1 and L_4 , noted as I_{L1} and I_{L2} , and the input voltage, V_{input} , are presented in Fig. 13, respectively. As can be seen, S_A and S_B are active during each half-line period, which can decrease the heat density. The inductor currents follow

Fig. 12. Experiential switching voltages waveforms of S_A and S_B

Fig. 13. Inductor current and the input voltage waveforms

Fig. 14. Switching waveforms of S_C , L_2 , and the input current I_{in}

the shape of input voltage in CrM operation. When L_1 operates in the positive half-line period, L_4 keeps idle with a zero inductor current. Similarly when L_4 is active in the negative half-line period, L_1 is in idle state.

The current waveform of L_2 , the switching waveforms of S_C , and the input current I_{in} are shown in Fig.14. S_C operates during the whole-line cycle, while L_2 and L_3 are employed for each half-line cycle. Two switches, S_A and S_C or S_B and S_C , are active under interleaved operation during each half-line period, resulting the same operation as the conventional interleaved circuit. However, it should be noted that the number of conduction devices is decreased by employing bridgeless topology.

The input current and voltage, the output voltage, and the sensing signal of output current under full load

Fig. 15. Waveforms of the proposed PFC circuit under rated load and 220V input voltage.

Fig. 16. Waveforms of the inductor currents and the total output current

condition are shown in Fig. 15. The input current and the input voltage are in the same phase and are practically sinusoidal. The output voltage is constant with a very low ripple.

The output current and the inductor current waveforms in branch I are shown in Fig. 16. CrM operation of the circuit is shown through the waveforms. As the same as the simulation results, the ripple frequency of total current is increased by twice and the peak value of current ripple was significantly decreased when compared with the inductor ripple current, which is a benefit of the interleaved operation. Therefore, the size of EMI filter can be reduced. In addition, reserve-recovery noise can also be decreased significantly.

The measured power factors (PF) results of the proposed circuit under low-line and high-line input voltages are shown in Fig. 17. It can be seen that the proposed circuit achieves very high PF under low line input voltage of 85 V, which is above 0.98 from 10% light load to the rated load. Under the high line input voltage of 265 V and the rated load, the PF of the proposed circuit is higher than 99.6%.

The measured total harmonic distortions (THD) results under rated load and input voltage of 265 V conditions are shown in Fig. 18. The results show that very low THD results are achieved for the proposed PFC circuit, which can meet the IEC61000-3-2 Class D requirements.

The measured system efficiencies of the proposed circuit and a conventional interleaved PFC under 85 V and 265 V

Fig. 17. System power factor of the proposed circuit under high line and low line voltages.

Fig. 18. Measured system THD under 265V input and rated load

Fig. 19. Experimental results of system efficiency

input voltages are shown in Fig. 19. As the dual-boost PFC is in one-phase structure, which is not suitable for DCNG applications, it has been left off the comparison list. According to Fig. 19, the proposed circuit has higher efficiency than the conventional one within the whole load range. Under full load conditions of 85 V input, the improved efficiency can reach as high as 0.8%, while the

value is up to 0.35% at 265 V input. It can be seen that high efficiency improvement can be obtained under heavy load for the proposed circuit, because the device number in the current flowing path is decreased.

For the proposed circuit, the efficiency under rated load and input voltage of 85 V is up to 95.6%. The maximum efficiency is obtained as 97.6%, that is achieved at 500 W load at 265 V input. During the whole load range of 265 V input, the system efficiencies are always higher than 90%. As the power capacity becomes higher, the benefit in terms of the efficiency of the proposed circuit becomes larger than the conventional circuits. Therefore, the proposed topology is more suitable for higher power applications, such as nano-grid systems, high power battery chargers, etc.

5. Conclusion

In this paper, a high performance interleaved bridgeless boost PFC circuit has been proposed for nano-grid applications. The circuit combines the conventional interleaved technique with bridgeless topology. The high conduction losses caused by the forward voltage of the rectifier diode bridge have been reduced significantly by the bridgeless technique. With operating as interleaved topology in CrM, this circuit is suitable for higher power applications with lower current ripple requirement. Simulated circuit based on Pspice was developed and a 600 W prototype circuit was designed to evaluate the feasibility of the proposed circuit. Very high PF and low THD were obtained through the experiments. In addition, the maximum system efficiency of 97.6% was obtained under 500 load and input voltage of 265 V.

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