A Cascaded High Step-up DC-DC Converter with Single Switch for Microsource Applications

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Abstract—This paper proposes a new high step-up DC-DC converter designed especially for regulating the DC interface between various microsources and a DC-AC inverter to electricity grid. The configuration of proposed converter is a quadratic boost converter with the coupled inductor in the second boost converter. The converter achieves high step-up voltage gain with appropriate duty ratio and low voltage stress on the power switch. Additionally, the energy stored in the leakage inductor of the coupled-inductor can be recycled to the output capacitor. The operating principles and steady-state analyses of continuous-conduction mode and boundary-conduction mode are discussed in detail. To verify the performance of the proposed converter, a 280W prototype sample is implemented with an input-voltage range of 20 ~ 40 V and output voltage of up to 400 V. The upmost efficiency of 93.3% is reached with high line input; on the other hand, the full-load efficiency remains at 89.3% during low line input.

Index Terms – microgrid, microsource, high step-up voltage gain, single switch, boost-flyback converter, coupled-inductor.

I. INTRODUCTION

Renewable energy is becoming increasingly important and prevalent in distribution systems, which provide different choices to electricity consumers whether they receive power from the main electricity source or in forming a microsource not only to fulfill their own demand but alternatively to be a power producer supplying a microgrid [1, 2, 31-33]. A microgrid usually includes various microsources and loads, which operate as an independent and controllable system when they are either grid-connected or islanded, as well as when they can reliably connect or disconnect [2, 33]. The microsource is classified as either a DC source or as a high-frequency AC source [3]. These two microsource categories are comprised of diverse renewable energy applications, such as solar-cell modules, fuel-cell stacks, wind turbines, and reciprocating engines [4, 31,
Fig. 1 shows a regular schematic of a microgrid unit supplied by various microsources; the high step-up converter is used to increase the output voltage of the microsource to 380 ~ 400 V for the DC interface to the main electricity source through the DC-AC inverter [2, 4, 31]. The single solar-cell module or fuel-cell stack both are essentially low-voltage sources, and thus a high step-up voltage gain DC-DC converter is required to regulate the voltage of the DC interface.

Previous research on various converters for high step-up applications has included analyses of the switched-inductor and switched-capacitor types [5, 6, 24, 25], the boost type integrating with the switched-capacitor technique [7, 25], the voltage-lift type [8, 9] the capacitor-diode voltage multiplier type [10], and the transformerless DC-DC converters [11, 23]. The voltage gain of these converters is insufficiently convert to a suitable AC source as a model microsource [1], in case of extremely high voltage gain is required, to using series connection of converter is able to reach much higher voltage gain. As known the efficiency and voltage gain of DC-DC boost converter are restrained by either the parasitic effect of power switches or the reverse-recovery issue of diodes. In addition, the equivalent series resistance (ESR) of capacitor and the parasitic resistances of inductor are also affecting overall efficiency. [5-26]. Although an alternative solution is the DC-DC flyback converter along with some advantages such as simple structure, easy control, and cost effective; the energy of leakage inductor of the transformer leaded to low efficiency and high voltage stress across the active switch. To employs active clamp technique not only to recycle the leakage inductor energy of the transformer but to constrained the voltage stress is crossed the active switch [12-14], however the tradeoff is higher cost and complex control circuit. Some converters effectively combined boost and flyback both converters as one or other different converter combinations are developed to carry out high step-up voltage gain by using the coupled-inductor technique [15-17]. In terms of higher voltage gain is constricted by the voltage stress on the active switch, once the leakage inductor energy of the coupled-inductor can be recycled that consequent the voltage stress on the active switch is reduced, this leads to the coupled-inductor and the voltage-multiplier or voltage-lift techniques are successfully accomplished the goal of higher voltage gain [18-26].

This paper presents a cascaded high step-up DC-DC converter to increase the output voltage of the microsource to a properly voltage level for the DC interface through DC-AC inverter to the main electricity grid. Proposed converter is a quadratic boost converter with the coupled inductor in the second boost converter.
The circuit diagram of proposed converter is shown in Fig. 2, the proposed circuit can be divided as a conventional boost converter and a boost-flyback converter [15]. These two segments are named first-boost stage and a second-boost stage. The first-boost stage is like a boost converter which including an input-inductor $L_{in}$, two diodes $D_1$ and $D_2$, and a pumping capacitor $C_1$. The second-boost stage is a boost-flyback converter which also including a dual winding coupled-inductor $T_1$, two diodes $D_3$ and $D_4$, and two output capacitors $C_{O1}$, and $C_{O2}$. Especially, these two stages are driven by a single switch $S_1$. The features of this converter as follows: 1) the quadratic boost converter is effectively extended voltage conversion ratio and the first-boost stage also benefited the input current ripple reduction. 2) the leakage inductor energy of the coupled-inductor can be recycled, which not just reducing the voltage stress on active switch, and also the conversion efficiency is significantly improved.

II. OPERATING PRINCIPLE OF THE PROPOSED CONVERTER

The simplified circuit model of the proposed converter is shown in Fig. 3. The dual-winding coupled-inductor consisted of a magnetizing-inductor $L_m$, primary leakage inductor $L_{k1}$, secondary leakage-inductor $L_{k2}$, and an ideal transformer, which is constituted the primary and secondary windings, $N_1$ and $N_2$, respectively. In order to simplify the circuit analysis of the proposed converter, some assumptions are declared as below:

1) All components are ideally except the leakage inductor of coupled-inductor is considered. The ON-state resistance $R_{DS(ON)}$ and all parasitic capacitors of the main switch $S_1$ are neglected, in addition the forward voltage drop of the diodes $D_1 – D_4$ are ignored.

2) The capacitor $C_1$ and the output capacitors $C_{O1}$ and $C_{O2}$ are sufficiently large, and the voltages cross $C_1$, $C_{O1}$ and $C_{O2}$ are considered as constant during one switching period.

3) The equivalent series resistance of capacitors $C_1$, $C_{O1}$ and $C_{O2}$ all are neglected.

4) The turn ratio $n$ of dual-winding coupled-inductor $T_1$ is equal to $N_2/N_1$.

The operating principle of continuous conduction mode (CCM) is presented in detail as below.

(A) CCM Operation

Fig. 4 is rendering several typical waveforms during five operating modes at one switching period $T_s$ while input-inductor $L_{in}$ and magnetizing-inductor $L_m$ both are operated in CCM. The operating modes are described as follows:
1) Mode I \([t_0, t_1]\): In this transition interval, switch \(S_1\) is turned on. Diodes \(D_1\) and \(D_3\) are conducted but diodes \(D_2\) and \(D_4\) are turned off. The current flow path is shown in Fig. 5(a). The energy of the DC source \(V_{in}\) is transferred to input-inductor \(L_{in}\) through diode \(D_1\), the voltage across input-inductor \(L_{in}\) is \(V_{in}\); the input current \(i_{in}\) is equal to \(i_{D1}\) and increased. The capacitor \(C_1\) delivering its energy to magnetizing-inductor \(L_m\) and primary leakage-inductor \(L_{k1}\). The voltage across magnetizing-inductor \(L_m\) and primary leakage-inductor \(L_{k1}\) is \(V_{c1}\), but the magnetizing-ductor \(L_m\) keep transferring its energy through secondary leakage-inductor \(L_{k2}\) to charge capacitor \(C_{O2}\), that both currents \(i_{Lk2}\) and \(i_{Lm}\) are decreased, until the increasing \(i_{Lk1}\) is reached and equaled to decreasing \(i_{Lm}\), in the meantime, current \(i_{Lk2}\) is down to zero at \(t = t_1\) that this mode is ended. The energies stored in capacitors \(C_{O1}\) and \(C_{O2}\) are constantly discharged to the load \(R\).

2) Mode II \([t_1, t_2]\): During this interval, switch \(S_1\) is remained on. Only diode \(D_1\) is conducted and rest of other diodes \(D_2, D_3\) and \(D_4\) are turned off. The current flow path is shown in Fig. 5(b). The energy of the DC source \(V_{in}\) is still stored into input-inductor \(L_{in}\) through diode \(D_1\). The energy has charged in capacitor \(C_1\) is still delivered to magnetizing-inductor \(L_m\) and primary leakage-inductor \(L_{k1}\). The voltage across magnetizing-inductor \(L_m\) and primary leakage-inductor \(L_{k1}\) is \(V_{c1}\). Thus, currents \(i_{in}\), \(i_{D1}\), \(i_{Lm}\) and \(i_{Lk1}\) are increased. The energies stored in capacitors \(C_{O1}\) and \(C_{O2}\) are still discharged to the load \(R\). This mode is ended when switch \(S_1\) is turned off at \(t = t_2\).

3) Mode III \([t_2, t_3]\): During this interval, switch \(S_1\) and diode \(D_1\) are turned off; the diodes \(D_2, D_3\) and \(D_4\) are conducted. The current flow path is shown in Fig. 5(c). The DC source \(V_{in}\) and input-inductor \(L_{in}\) are serially to charge capacitor \(C_1\) with their energies. Meanwhile, primary leakage-inductor \(L_{k1}\) is in series with capacitor \(C_1\) as a voltage source \(V_{c1}\) through magnetizing-inductor \(L_m\) then delivered their energies to charge capacitor \(C_{O1}\). The magnetizing-inductor \(L_m\) also transferred the magnetizing energy through coupled-inductor \(T_1\) to secondary leakage-inductor \(L_{k2}\) and to charge capacitor \(C_{O2}\). Thus, currents \(i_{in}\), \(i_{D2}\), \(i_{D4}\), \(i_{Lm}\) and \(i_{Lk1}\) are decreased. But, currents \(i_{C1}\), \(i_{Lk2}\) and \(i_{D3}\) are increased. The energy stored in capacitors \(C_{O1}\) and \(C_{O2}\) are discharged to the load \(R\). This mode is ended when current \(i_{CO1}\) is dropped till zero at \(t = t_3\).

4) Mode IV \([t_3, t_4]\): During this transition interval, switch \(S_1\) and diode \(D_1\) are remained off; and diodes \(D_2, D_3\) and \(D_4\) are still conducted. The current flow path is shown in Fig. 5(d). Almost statuses are remained as Mode III except the condition of primary leakage-inductor \(L_{k1}\) is in series with capacitor \(C_1\) as
a voltage source $V_{C1}$ through magnetizing-inductor $L_m$ then discharged or released their energies to load. Thus, currents $i_{in}, i_{D2}, i_{D4}, i_{Lm}$ and $i_{Lk1}$ are persistently decreased, but currents $i_{CO2}, i_{Lk2}$ and $i_{D3}$ are still increased. The energy stored in capacitors $C_{O1}$ and $C_{O2}$ is discharged to the load $R$. This mode is ended when current $i_{Lk1}$ is decreased until zero at $t = t_4$.

5) Mode V [$t_4$, $t_6$]: During this interval, switch $S_1$ and diode $D_1$ are remained off; diode $D_4$ is turned off and diodes $D_2$ and $D_3$ are keep in conducted. The current flow path is shown in Fig. 5(e). The DC source $V_{in}$ and input-inductor $L_{in}$ are serially and still charged to capacitor $C_1$ with their energies. The magnetizing-inductor $L_m$ continuously transferred its own magnetizing energy through coupled-inductor $T_1$ and diode $D_3$ to secondary leakage-inductor $L_{k2}$ and to charge capacitor $C_{O2}$. Thus, currents $i_{in}, i_{D2}, i_{D3}, i_{Lk2}$ and $i_m$ are decreased. The energies stored in capacitors $C_{O1}$ and $C_{O2}$ are discharged to the load. This mode is end when switch $S_1$ is turned on at the beginning of next switching period.

III. STEADY-STATE ANALYSIS OF THE PROPOSED CONVERTER

(A) CCM Operation

Since the time durations of modes I and IV are transition periods, only modes II, III and V are considered at CCM operation for the steady state analysis. During the time duration of modes II that main switch $S_1$ is conducted, and the coupling coefficient of the coupled-inductor $k$ is considered as $L_m / (L_m + L_{k1})$. The following equations can be written from Fig 5(b).

$$v_{Lm} = V_{in}$$ (1)

$$v_{Lm} = \frac{L_m}{L_m + L_{k2}} V_{C1} = k \cdot V_{C1}$$ (2)

$$v_{Lk1} = V_{C1} - V_{Lm} = (1 - k) \cdot V_{C1}$$ (3)

$$v_{Lk2} = n \cdot v_{Lm}$$ (4)

During the period of modes III and V that main switch $S_1$ is turned off, the following equations can be found as below,

$$v_{Lm} = V_{in} - V_{C1}$$ (5)

$$v_{Lm} = V_{C1} - V_{CO1} - V_{Lk1}$$ (6)

$$v_{Lk2} = -n v_{Lm} - V_{CO2}$$ (7)
where the turn ratio of the coupled-inductor $n$ is equal to $N_2/N_1$. The voltage across inductor $L_{in}$ by the volt-second balance principle as below,

$$\int_0^{T_{DS}} V_{in} \, dt + \int_0^{T_{DS}} (V_{in} - V_{c1}) \, dt = 0$$

(8)

$$V_{c1} = \frac{I}{1 - D} V_{in}$$

(9)

The voltage across magnetizing-inductor $L_m$ by the volt-second balance principle as below,

$$\int_0^{T_{DS}} kV_{c1} \, dt + \int_0^{T_{DS}} (V_{c1} - V_{c1} + V_{L1}) \, dt = 0$$

(10)

$$\int_0^{T_{DS}} kV_{c1} \, dt + \int_0^{T_{DS}} \frac{(V_{c2} - V_{L2})}{n} \, dt = 0$$

(11)

Substitute (9) into (10) and (11), and $L_{k2}$ is assumed as equal to $nL_{k1}$, thus $V_{c1}$ and $V_{c2}$ can be obtained the following equations,

$$V_{c1} = \frac{I - D + kD}{I - D} V_{c1} - V_{L1} = \frac{I - D + kD}{(1 - D)^2} V_{in} - V_{L1}$$

(12)

$$V_{c2} = \frac{n \cdot kD}{I - D} V_{c1} - V_{L2} = \frac{n \cdot kD}{(1 - D)^2} V_{in} - n \cdot V_{L1}$$

(13)

The output voltage $V_o$ can be express as

$$V_o = V_{c1} + V_{c2}$$

(14)

Substituting (3), (12) and (13) into (14) can be obtain the voltage gain $M_{CCM}$.

$$M_{CCM} = \frac{V_o}{V_{in}} = \frac{k(n + 1) + n(D - 1)}{(1 - D)^2}$$

(15)

Fig. 6 is shown a line chart of the voltage gain versus the duty-ratio $D$ under three different coupling coefficients of the coupled-inductor while the $n = 4.4$ is given. It revealed the coupling coefficient $k$ is almost unaffected. Substituting $k = 1$ into (15) and (13) the input-output voltage gain can be simplified as below

$$M_{CCM} = \frac{V_o}{V_{in}} = \frac{1 + nD}{(1 - D)^2}$$

(16)

$$M_{CCM, TT} = \frac{V_{c2}}{V_{c1}} = \frac{nD}{1 - D}$$

(17)

Fig. 7 is demonstrated the voltage gain versus the duty ratio of the proposed converter and other converters in [16], [18] and [19] at CCM operation under $k = 1$ and $n = 4.4$. As long as the duty ratio of the
proposed converter is larger than 0.55 the voltage gain is higher than the converters in [16], [18] and [19]. Referring to the description of CCM operating modes, the voltage stresses on \( S_1 \) and \( D_1 - D_4 \) are given as

\[
V_{DS} = V_{D4} = \frac{V_o}{I + nD}
\]  
(18)

\[
V_{D1} = \frac{DV_o}{I + nD}
\]  
(19)

\[
V_{D2} = \frac{(1-D)V_o}{I + nD}
\]  
(20)

\[
V_{D3} = \frac{nV_o}{I + nD}
\]  
(21)

\( (B) \) BCM Operation

In terms of power utility ratio of some microsources as PV module is about 60% to 90% usually. The boundary condition is designed at 25% to 40% full load regularly for practical application and commercial products. That’s a major design feature of the proposed converter is operated at CCM. For steady-state analysis of the boundary condition mode (BCM) is presented in detail as below.

The two peak value of the magnetizing-inductor current \( I_{Lmp} \) and the input-inductor current \( I_{inp} \) are given

\[
I_{Lmp} = \frac{D T TV_{cl}}{L_n}
\]  
(22)

\[
I_{inp} = \frac{D T V_n}{L_{in}}
\]  
(23)

the \( D_L \) is defined as the duration of magnetizing-inductor current from peak ramped down to zero, and the \( D_X \) is defined as the duration of diode current \( i_{D4} \) from peak ramped down to zero, the average value of \( i_{D3} \) and \( i_{D4} \) during each switching period are written as

\[
I_{D3} = \frac{(D_L - D_X) I_{Lmp}}{2n}
\]  
(24)

\[
I_{D4} = \frac{D_X I_{Lmp}}{2}
\]  
(25)

from (24) and (25), \( D_X \) is obtained

\[
D_X = \frac{D_L}{I + n}
\]  
(26)

the output current \( I_O \) is derived as below
\[ I_o = \frac{V_o}{R} = \frac{V_{CO1} + V_{CO2}}{R} = \frac{nV_{CI} + (n+1)V_{CO2}}{R} \]  
(27)

Since the average currents of output capacitor \( I_{CO1} \) and \( I_{CO2} \) are zero in steady state, the average value of \( i_{D3} \) and \( i_{D4} \) are respectively equal to the average value of \( i_o \). Substituting (24), (25), and (26) into (27), \( I_o \) can be rewritten as

\[ I_o = \frac{V_o}{R} = \frac{(D_L - \frac{D_L}{1+n})V_{CI} \cdot T_s \cdot D}{2L_m} \]  
(28)

the normalized magnetizing-inductor time constant \( \tau_{Lm} \) is defined as

\[ \tau_{Lm} = \frac{L_m}{R \cdot T_s} = \frac{L_m \cdot f_S}{R} \]  
(29)

where \( f_S \) is the switching frequency. Substituting (28) into (29), the voltage conversion of \( V_{CO2} \) and \( V_{CI} \) can be obtained

\[ \frac{V_{CO2}}{V_{CI}} = -n + \sqrt{n^2 + 2(n+1) \frac{D^2n^2}{(1+n) \cdot \tau_{Lm}}} \]  
(30)

When the proposed converter is operated in BCM, that equation (17) is equal to (30) yields the boundary normalized magnetizing-inductor time constant \( \tau_{LmB} \)

\[ \tau_{LmB} = \frac{L_{mB}}{R \cdot T_s} = \frac{D \cdot (1-D)^2}{2(n^2D + nD + n + 1)} \]  
(31)

the curve of \( \tau_{LmB} \) is plotted in Fig. 8. Once the \( \tau_{Lm} \) is higher than \( \tau_{LmB} \) the couple-inductor is operated in CCM.

The voltage gain of the second-boost stage can be derived

\[ \frac{V_{CI}}{V_{IN}} = \sqrt{1 + \frac{1 - 4\left(\frac{L_m}{L_{in}} \cdot \frac{D \cdot (1+n)}{L_m \cdot D \cdot (1+n) + D_L}\right)}{2}} \]  
(32)

while the second-boost stage of the proposed converter is operated in BCM, the \( D_L \) is equal to \((1 - D)\) and also the (9) is equal to (32), the ratio of magnetizing-inductor \( L_m \) and input-inductor \( L_{in} \) can be obtained as

\[ \frac{L_m}{L_{in}} = \frac{nD + 1}{(1+n) \cdot (1-D)^2} \]  
(33)

The curves in Fig. 9 are illustrated the ratio of magnetizing-inductor and input-inductor, \( L_m / L_{in} \), versus duty ratio for different turn-ratios, which is shown the correlation of \( L_m \) and \( L_{in} \). These two inductors are
applied to the same of duty ratio due to both are driven by the common switch $S_1$. Substituting (33) into (31), the boundary normalized input-inductor time constant $\tau_{\text{LinB}}$ can be derived as

$$\tau_{\text{LinB}} = \frac{L_{\text{inB}}}{R \cdot T_s} = \frac{D \cdot (1 - D)^4}{2(nD + 1)^2}$$

(34)

The curve of $\tau_{\text{LinB}}$ is plotted in Fig. 10. Once the $\tau_{\text{Lin}}$ is higher than $\tau_{\text{LinB}}$, the input-inductor is operated in CCM.

IV. EXPERIMENTAL RESULTS

A 280 W prototype sample is presented to demonstrate the practicability of the proposed converter. The electrical specification is $V_{in} = 20$ V, $V_o = 400$ V, $f_s = 40$ kHz, $P_o = 280$ W (the full load resistance $R \approx 570$ Ω). The requirement of major components as $C_1 = 1000$ μF, $C_{O1}$ is the same as $C_{O2} = 220$ μF, the main switch $S_1$ is a MOSFET IXFK180N15P, the diodes $D_1$ and $D_2$ both are MBR30100CT, $D_3$ is BYR29-600 and MBR20200CT is selected to $D_4$.

Based on specifications, the voltage gain is up to 20 and $n = 4.4$ substituting into (16), the duty ratio $D$ is about 0.58. Then, substituting $D$ and $n$ into (31) and (34), the boundary normalized magnetizing-inductor time constant $\tau_{\text{LmB}}$ is obtained as 0.002667, the boundary normalized input-inductor time constant $\tau_{\text{LinB}}$ is obtained as 0.000715. To define the proposed converter is operated in CCM, the boundary condition is designed at 40% full load, the load resistance $R$ approximated to 1400 Ω. When both $\tau_{\text{Lm}}$ and $\tau_{\text{Lin}}$ are larger than $\tau_{\text{LmB}}$ and $\tau_{\text{LinB}}$ respectively, the proposed converter is operated in CCM. The $L_m$ and $L_{in}$ are found as,

$$\frac{L_{\text{inB}} \cdot f_s}{R} > 0.002667 \Rightarrow L_{\text{in}} > 93.4 \mu H$$

(35)

$$\frac{L_{\text{inB}} \cdot f_s}{R} > 0.000715 \Rightarrow L_{\text{m}} > 25 \mu H$$

(36)

the actual inductance on input-inductor $L_{\text{in}}$ and the magnetizing-inductor $L_{m}$ of couple-inductor are measured as 29 μH and 94 μH, respectively.

The practical operating condition is $V_{in} = 20$ V, $V_o = 400$ V, $P_o = 280$ W and derived by 40 kHz gate signal, all experimental waveforms are measured and shown in Fig. 11. As Fig. 11(a) the waveform of input current $I_{in}$ is continuously, and the waveforms of current $i_{DS}$ and voltage $v_{DS}$ crossed switch $S_1$. The Fig. 11(b) is shown the current waveforms through diodes $D_1$, $D_2$, $D_3$ and $D_4$. Fig. 11(c) is shown the waveforms of
voltage are across capacitors $C_1$, $C_{O1}$, $C_{O2}$ and load. From these experimental waveforms are agreed with the operating principle and the steady-state analysis.

In terms of diversely applications, the nominal output voltage of a single fuel-cell stack or solar-cell module about 24 V to 36 V by 150W to 250W power capacity. The proposed converter input voltage range is from 20 V to 40V which is completely fulfilled the utilization of regular microsources. This experiment is also verified the performances of converter efficiency at high-line (40 V) and low-line (20 V) input voltages by different loadings as shown in Fig. 12. Once the low-line input voltage, the highest efficiency is up to 92.1% at 30% of full load; in case of full load condition, the efficiency still reached to 89.3%. When the high-line input voltage, that efficiency is up to 92.5% at full load condition, while output load decreased to about 30% of full load, the maximum efficiency is 93.3%.

V. CONCLUSIONS

A boost converter and a flyback converter are successfully combined as a quadratic boost converter driven by a single switch and achieved high step-up voltage gain, the voltage gain is up to 20 times than input. The energy of leakage inductor of the coupled-inductor can be recycled, which is effectively constrained the voltage stress of the main switch $S_1$ and benefits the low on-state resistance $R_{DS(ON)}$ can be selected. The upmost efficiency 93.3% is measured at high line input, the full-load efficiency still remains at 89.3%. As long as adding active snubber, auxiliary resonant circuit, synchronous rectifiers, or switched-capacitor-based resonant circuits and so on, that all are able to achieve soft switching on the main switch to reaching higher efficiency [27]-[30].

REFERENCES


Fig. 1. The basic schematic of microgrid consisted of diversely microsources and power converters.

Fig. 2. Circuit configuration of the proposed converter.

Fig. 3. Simplified circuit model of the proposed converter.

Fig. 4. Some typical waveforms of the proposed converter at both $L_{m}$ and $L_{in}$ are CCM operation.
Fig. 5. Current flow path of operating modes during one switching period at CCM operation. (a) Mode I. (b) Mode II. (c) Mode III. (d) Mode IV. (e) Mode V.

Fig. 6. Voltage-gain versus duty ratio at CCM operation under $n = 4.4$ and diverse $k$.

Fig. 7. Voltage gain versus duty-ratio of the proposed converter, the converters in [16], [18] and [19] at CCM operation under $n = 4.4$ and $k = 1$. 

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Fig. 8. Boundary condition of $r_{Lin}$ of the proposed converter under $n = 4.4$.

Fig. 9. The ratio curves of magnetizing-inductor and input-inductor versus duty ratios for different turn ratios.

Fig. 10. Boundary condition of $r_{Lin}$ of the proposed converter under $n = 4.4$.

Fig. 11. Experimental waveforms are deriv by the condition of $f_S = 40kHz$, $V_{in} = 20$ V and output 280 W. (a) $v_{GS}$, $i_{DS}$, $v_{DS}$ and $i_{DS}$. (b) $i_{D1}$, $i_{D2}$, $i_{D3}$, and $i_{D4}$. (c) $V_{C1}$, $V_{C01}$, $V_{C02}$ and $V_O$.
Fig. 12. Measured efficiency of the proposed converter.