Nonisolated Multiport Converters Based on Integration of PWM Converter and Phase-Shift Switched Capacitor Converter

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Abstract—Photovoltaic (PV) systems having rechargeable batteries are prone to be complex and costly because multiple converters are necessary to individually regulate a load, PV panel, and battery. This paper proposes novel nonisolated multiport converters (MPCs) integrating a bidirectional PWM converter and phase-shift switched capacitor converter (PS-SCC) for standalone PV systems. A PWM converter and PS-SCC are integrated with reducing the total switch count, realizing the simplified system and circuit. In the proposed MPCs, two control freedoms of duty cycle and phase shift angle are manipulated to individually regulate the load, PV panel, and/or battery. The detailed operation analysis was performed to mathematically derive gain characteristics and ZVS operation boundaries. For the battery discharging mode, in which the PV panel is not available and the MPC behaves as a single-input-singleoutput converter with two control freedoms available, the optimized control scheme achieving the lowest RMS current is also proposed to maximize power conversion efficiencies. Various kinds of experimental verification tests using a 200-W prototype were performed to verify the theoretical analysis and to demonstrate the performance of the proposed MPC.

Keywords—Bidirectional pulse width modulation (PWM) converter, multiport converter (MPC), nonisolated dc-dc converter, phase-shift switched capacitor converter (PS-SCC)

I. INTRODUCTION

Recent power systems are prone to be complex and costly as they comprise multiple power sources and loads. Photovoltaic (PV) systems, for example, consist of not only PV panels but also rechargeable batteries to buffer weather-dependent unstable power generation of panels. Hybrid electric vehicles also contain multiple power sources including a generator and multiple batteries for various loads. In such multi-power-source systems, multiple converters in proportion to the number of power sources are required to regulate power sources individually, as shown in Fig. 1(a).

To reduce the converter count in such systems, various kinds of multiport converters (MPCs) that integrate multiple converters into a single unit have been proposed and developed, as illustrated in Fig. 1(b). MPCs are roughly classified into three categories: the isolated, partially-isolated, and nonisolated topologies. Among the most typical isolated MPC topologies is a triple active bridge (TAB) converter [1]–[3] that is an extended version of traditional dual active bridge (DAB) converters. The number of input and output ports can be extended by adding transformer windings as well as inverter bridges. The TAB topologies, however, are prone to complexity due to the large switch count because each inverter bridge requires two or four switches for half- and full-bridge topologies.

The partially-isolated MPCs, on the other hand, can reduce the switch count by sharing switches between isolated and nonisolated converters [4]–[14]. These MPCs are derived from the combination of a bidirectional PWM converter and an isolated converter, such as full- or half-bridge converters [4]– [7] and resonant converters [8], [9]. Among various promising topologies are the DAB-based MPCs [10]–[14] that achieve zero voltage switching (ZVS) in wide operation ranges, realizing efficient and flexible power conversion thanks to the reduced switching loss and their inherent bidirectional power conversion capability. Although partially-isolated MPCs are an appealing topology from the viewpoint of component count, a bulky transformer is indispensable regardless of isolation requirement. For nonisolated applications, nonisolated MPCs are undoubtedly suitable because of the lack of bulky and



Fig. 1. (a) Conventional system with multiple converters. (b) MPC-based system.

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expensive transformers, which also lead to substantial power losses.

Various types of nonisolated MPCs have been reported [15]– [24]. In MPCs operating with a time division manner [15], [16], all input or output ports share an on-duty cycle in a single switching cycle, hence resulting in decreased effective duty cycle, increased RMS currents, and deteriorated power conversion efficiencies. MPCs with a shared bus can reduce the number of passive components to some extent [17]–[19], but most active switches remain unshared. Topology reported in [20]–[24], on the other hand, can reduce both passive and active component counts, allowing reduced costs and simplified topology. These topologies, however, pose major issues such as narrowed operation ranges [20]–[22], unshared ground [23], and increased circuit volume due to the requirement of numerous inductors [24].

Switched capacitor converters (SCCs) are widely known as high power-density converters for nonisolated applications. SCCs chiefly rely on capacitors rather than inductors as an energy storage medium in circuits, realizing miniaturized circuit design because an energy density of discrete capacitors is 100-1000 times greater than that of similarly-scaled inductors [25]-[27]. Power conversion efficiencies of SCCs, however, are known to decrease when the required load voltage is lower than the theoretically attainable voltage [29]. To cope with this issue, hybrid SCCs employing an additional inductor to realize efficient voltage regulation capability have been proposed. With a single additional inductor, SCCs can be modified to be hybrid SCCs that can be regulated with PWM [27], [28], PFM [29], [30], or phase-shift (PS) control [31]. Despite the additional inductors, power densities of hybrid SCCs are reportedly greater than those of ordinary inductorbased converters [25]. In addition to the enhanced power densities, the hybrid SCCs with PS control (hereafter called PS-SCCs) achieve ZVS and flexible power flow, making them an attractive candidate for nonisolated MPC topologies.

This paper proposes nonisolated MPCs based on PS-SCCs for standalone PV systems. A traditional bidirectional PWM converter and PS-SCC are integrated with sharing active switches, achieving simplified circuit. The remaining of this paper is organized as follows. Section II presents the derivation and major features of the proposed MPCs. Section III introduces three operation scenarios and control schemes in the proposed MPCs. The detailed operation analyses will be performed in Sections IV and V. A design example for a 200-W experimental prototype will be presented in Section VI, followed by the experimental verification in Section VII. The proposed and conventional MPCs will be compared from various aspects in Section VIII.

II. PROPOSED NONISOLATED MPCS

A. Key Elements

The combination of a traditional bidirectional PWM converter and PS-SCC, shown in Figs. 2(a) and (b), derives a proposed nonisolated MPC. Another key circuit element is a nonisolated DAB converter [see Fig. 2(c)], which can be derived from the PS-SCC of Fig. 2(b). To be specific, by breaking the source pin of Q_3 in the PS-SCC [marked as "a" in

Fig. 2(b)] and connecting it to the ground, the PS-SCC can be transformed into the nonisolated DAB converter. Fundamental operation principle and major features of the nonisolated DAB converter are identical to those of the PS-SCC, though their suitable voltage conversion ratios differ—voltage conversion ratios of M = 2.0 and 1.0 are the best conditions for the PS-SCC [31] and DAB converter [32], respectively, from the viewpoint of power conversion efficiency.

A resonant SCC topology [30] is very similar to the PS-SCC in Fig. 2(b) but is considered not suitable for the proposed MPC in standalone PV systems. Although the inductor L can be smaller thanks to resonant operations, relatively narrow voltage regulation ranges of resonant topologies are a major drawback. Since voltages of rechargeable batteries and PV panels vary significantly, PS-SCCs with wider regulation ranges are a preferable topology—for applications where voltage regulation ranges are not of importance, resonant SCCs would be an appealing candidate from the viewpoint of circuit miniaturization.

B. Derivation of Proposed MPCs

By sharing two switches (Q_1 and Q_2) of the bidirectional PWM converter and PS-SCC or nonisolated DAB converter, the proposed SCC-MPC and DAB-MPC can be derived, as shown in Figs. 3(a) and (b). Switches, Q_1 and Q_2 , are shared by two circuits, hence reducing the total switch count and realizing the simplified topology.

Although the PS-SCC and nonisolated DAB converter are integrated with the bidirectional PWM converter, their original features and suitable voltage conversion ratios are essentially retained. Hence, a suitable MPC topology should be selected with considering applications and requirements. Our target application in this paper, for example, is a standalone PV system with the PV panel voltage $V_{in} = 30$ V, the battery voltage $V_{bat} = 12-16$ V, and the load voltage $V_{out} = 48$ V. Therefore, the SCC-MPC is preferable because the target system corresponds to M = 1.6 that is closer to M = 2.0 than M = 1.0. The following sections focus mainly on the SCC-MPC.



Fig. 2. Key elements for the proposed MPCs. (a) Bidirectional PWM converter. (b) Phase-shift switched capacitor converter (PS-SCC). (c) Nonisolated dual active bridge (DAB) converter.



Fig. 3. Proposed nonisolated MPCs. (a) SCC-MPC. (b) DAB-MPC.

Although the PS-SCC shown in Fig. 2(b) is the main focus in this paper, various kinds of SCCs have been reported [33]– [36] and can be applied as PS-SCC topologies. Based on these PS-SCC topologies, a variety of PS-SCC-based MPCs can also be derived. Details about these SCCs and their MPC versions will be discussed in the next subsection.

C. Family of PS-SCCs and their MPC Versions

Various PS-SCCs can be derived by adding inductors to conventional SCCs, as listed in Table I. It should be noted that Table I shows topologies containing two L-C circuits for differentiation because these SCCs with only one L-C circuit are equivalently identical. Structures of these PS-SCCs can be extended by adding L-C circuits and switches to obtain higher voltage conversion ratios.

The MPC versions of these PS-SCCs can be derived by adding an LC filter of L_{bat} and C_{bat} at the switching nodes "b" in Table I. LC filters can be connected to any switching nodes other than the node "b," but voltage conversion ratios of PWM converters naturally differ. Hence, an optimal MPC topology as well as a switching node should be selected depending on applications and requirements.

The four kinds of SCC-MPCs are compared in terms of component counts, voltage gain, advantages, and disadvantages, as summarized in Table II, in which *n* represents the number of L-C circuits. Conventional non-PS-SCCs have been quantitatively analyzed and compared from various aspects in the past works [34], [36], and their major features would be unchanged even for PS versions. The ladder-SCC, the most widely-used SCC topology, is the foundation of the proposed MPC shown in Fig. 3(a). Voltage stresses of all switches and

capacitors are nearly identical in the ladder-SCC, allowing simple circuit design and good modularity. In the Dickson SCC, on the other hand, capacitors' voltage stress varies depending on positions. A capacitor count can be reduced with the seriesparallel SCC, but switches must be properly selected with considering individual voltage stresses. The Fibonacci SCC is a suitable topology for applications needing high step-up or down voltage conversion, whereas both switches and capacitors are exposed to different voltage stresses depending on positions, resulting in increased design difficulty. The target application in this paper is a 200-W standalone PV system with the voltage gain M = 1.6, as mentioned in Section II-B. Given the target specification, the ladder-type SCC-MPC is considered to be the best topology from the viewpoint of design simplicity.

D. Features

Similar to conventional MPCs, the proposed MPCs achieve system simplification thanks to the integration of two separate converters into a single unit. Furthermore, the switch count is halved, thus allowing the circuit-level simplification. All input and output ports are capable of bidirectional power flows because of the integration of two kinds of bidirectional converters, though our target application in this paper is the standalone PV system, in which the input power source is unidirectional.

To control all input and output ports individually, the proposed MPCs employ two control schemes of PWM and PS controls. The battery voltage V_{bat} or battery current I_{bat} (see Fig. 3) is regulated by PWM control adjusting duty cycle *d* of Q₂ and Q₄ for constant-current–constant-voltage (CC–CV) charging. Meanwhile, the PS control plays the role of the load power regulation by manipulating PS angle φ between the leading (Q₁–Q₂) and lagging legs (Q₃–Q₄).

ZVS operations are feasible depending on load conditions, similar to conventional PS converters. However, since the proposed MPCs are transformerless and their voltage conversion ratios cannot be adjusted by turns ratios, ZVS ranges tend to be narrower than those of traditional transformerbased PS converters.

III. OPERATION SCENARIOS AND CONTROL SCHEME

A. Operation Scenarios

Although all input and output ports of the proposed MPCs are capable of bidirectional power flow, our target application in this paper is a standalone PV system where a PV panel, a unidirectional power source, is tied to one of the ports. A rechargeable battery and non-regenerative load are connected to the remaining ports.

For the MPCs having three input and output ports, controlling two of them automatically determines the remaining one, as expressed by the simple equation;

$$P_{in} = P_{out} + P_{bat} \,, \tag{1}$$

where P_{in} , P_{out} , and P_{bat} are the input, output, and battery powers, respectively, as designated in Fig. 1(b). Depending on the power balance among three ports, the MPC operates in one of the following three scenarios: the CC–CV battery charging mode, battery discharging mode, and MPPT mode, as illustrated in Fig. 4.

CC–CV Battery Charging Mode $(P_{in} > P_{out})$ [Fig. 4(a)]: The PV panel is capable of supplying not only P_{out} but also P_{bat} , and

Table I. Family of PS-SCCs and SCC-MPCs.



the surplus power (i.e., $P_{in}-P_{out}$) is greater than the acceptable charging power of the battery. Hence, a charging current or voltage is regulated to be constant. In other words, P_{bat} is regulated based on the CC–CV charging scheme in this operation mode. The MPC regulates P_{bat} and P_{out} with PWM and PS controls, respectively, whereas P_{in} is unregulated in this mode. The detailed operation analysis for the battery charging mode will be performed in Section IV.

Battery Discharging Mode ($P_{in} = 0$) [Fig. 4(b)]: When the input power source is no longer available (e.g., PV panels at night), the battery alone supplies the whole load power. Hence, the MPC in this mode behaves as a single-input-single-output converter. However, there are two control freedoms of duty

cycle *d* and PS angle φ for one output port. The operation in the battery discharging mode will be detailed in Section V-A. In addition, the optimal control strategy manipulating both *d* and φ to maximize the power conversion efficiency in the battery discharging mode will be proposed in Section V-B.

MPPT Mode ($P_{in} > P_{out}$ or $P_{in} < P_{out}$) [Fig. 4(c)]: The PV panel is regulated with PWM control employing an MPPT algorithm, regardless of the load power demand, while P_{out} is regulated by PS control. Thus, P_{in} and P_{out} are regulated with PWM and PS controls, respectively, whereas P_{bat} is unregulated in the MPPT mode. When P_{in} surpasses P_{out} (i.e., $P_{in} > P_{out}$), P_{bat} is positive, and the surplus power is allocated for battery charging. The power flow in this case is identical to that in the

| Topology | Component Count | | | Voltage | A dvantage | Disa dyanta sa | |
|---------------------|-----------------|---------------|-------------|--|--|---|--|
| | Q | С | L | Gain | Auvantage | Disacivantage | |
| Ladder | 2(<i>n</i> +1) | 2 <i>n</i> +2 | <i>n</i> +1 | $M \approx n+1$ | Unform voltage stresses of swiches and capacitors | High capacitor count | |
| Dickson | 2(<i>n</i> +1) | 2 <i>n</i> +3 | <i>n</i> +1 | $M \approx n+1$ | Unform switch voltage stresses | High capacitor count | |
| Series- Parallel | 3 <i>n</i> +1 | n+3 | <i>n</i> +1 | $M \approx n+1$ | Low capacitor count | Non-uniform switch voltage stresses | |
| Fibonacci | 3 <i>n</i> +1 | n+3 | <i>n</i> +1 | $M \approx 2, 3, 5, 8$ for $n = 1, 2, 3, 4$ | High voltage gain Low capacitor count | Non-uniform voltage stresses of switches and capacitors | |

Table II. Comparison among SCC-MPCs.



Fig. 4. Power flows in (a) CC–CV battery charging mode, (b) battery discharging mode, and (c) MPPT mode.

battery charging mode. The difference from the CC–CV charging mode is that the surplus power is within the battery's acceptable charging power. If P_{in} falls below P_{out} (i.e., $P_{in} < P_{out}$), P_{bat} becomes negative, and the battery discharges to support the PV panel. Since the power flow in this operation mode is similar to that in the battery charging and discharging modes, the detailed analysis of this mode is omitted to save page length.

B. Control Scheme

A control block diagram is shown in Fig. 5. In all operation scenarios, the output voltage V_{out} is regulated by adjusting the PS angle φ . In the CC–CV battery charging and MPPT modes, *d* is manipulated to regulate either P_{in} or V_{bat} (or I_{bat}), depending on the operation scenarios, as discussed in Section III-A. Control loops for P_{in} in the MPPT mode and V_{bat} or I_{bat} in the CC–CV charging mode are seamlessly switched by the minimum function in accordance with whether the surplus



Fig. 5. Control block diagram.

power $(P_{in} - P_{out})$ is greater than the battery's acceptable charging power. Any MPPT algorithms can be employed for the proposed MPCs, and a traditional hill climbing-based MPPT will be used for the experimental verification in Section VII-F.

In the battery discharging mode, in which two control freedoms of d and φ can be available to regulate V_{out} , the optimal relationship between d and φ is determined based on (19), which needs to be predetermined based on the operation analysis. Similar to other operation scenarios, φ is manipulated to regulate V_{out} , while d is adjusted based on the optimal control scheme of (19) to minimize Joule losses of the converter, as will be detailed in Section V-B.

IV. OPERATION ANALYSIS FOR CC–CV BATTERY CHARGING MODE

The detailed operation analysis is performed only for the SCC-MPC [see Fig. 3(a)] to save the page length, but other topologies can be analyzed in a similar manner. The operation analysis is based on the following assumptions; all the components are ideal, the capacitance of C is large enough so that it can be regarded as a constant voltage source, and dead-time periods are negligibly short.

A. Battery and Capacitor Voltages

Before detailing the operation analysis, equations for the battery voltage V_{bat} and the voltage of the capacitor C, V_C , are derived in this subsection. Average voltages at the switching nodes of the leading leg (Q₁–Q₂) and lagging leg (Q₃–Q₄) are dV_{in} and $(1-d)V_{in} + dV_{out}$, respectively. Meanwhile, average voltages of the inductors L_{bat} and L must be zero under steady-state conditions, and hence, V_{bat} and V_C can be yielded as

$$V_{bat} = dV_{in} \tag{2}$$

$$V_C = (1 - 2d) V_{in} + dV_{out} \tag{3}$$

Equation (2) is identical to a voltage conversion ratio of traditional PWM buck converters and does not include the PS angle φ , suggesting that V_{bat} can be independently regulated by PWM control.

B. Mode Analysis

The key operation waveforms and current flow directions are shown in Figs. 6 and 7, respectively. φ_d is the PS duty cycle (= $\varphi/360^\circ$), *T* is the switching period, and $i_{Q1}-i_{Q4}$ are the switch currents. It should be noted that operation waveforms and current flow directions in these figures are independent on



Fig. 6. Key operation waveforms in CC-CV battery charging mode.

either I_{bat} or V_{bat} is regulated in the CC or CV charging modes. The battery port is regulated in both the CC charging or CV charging modes, and therefore the analysis in this section can be applied to the entire period of the CC–CV charging mode.

The current of L, i_L , at t = 0, T_1 , T_2 , and T_3 is derived from the volt-sec (flux) balance on L and the charge balance on C, as

$$\begin{cases} i_{L}(0) = \frac{(1-d) \{ 2(\varphi_{d} - d) V_{in} - (2\varphi_{d} - d) V_{out} \}}{2fL} \\ i_{L}(T_{1}) = \frac{d \{ -2(1-d-\varphi_{d}) V_{in} + (1-d) V_{out} \}}{2fL} \\ i_{L}(T_{2}) = \frac{d \{ 2(1-d-\varphi_{d}) V_{in} - (1-d-2\varphi_{d}) V_{out} \}}{2fL} \\ i_{L}(T_{3}) = \frac{(1-d) \{ -2(\varphi_{d} - d) V_{in} - dV_{out} \}}{2fL} \end{cases}$$

$$(4)$$

where f (= 1/T) is the switching frequency. Similarly, the current of L_{bat}, i_{Lbat} , at t = 0 and T_2 is derived from the volt-sec balance on the inductor L_{bat}, as

$$\begin{cases} i_{Lbat}(0) = I_{bat} + \frac{d(1-d)V_{in}}{2fL_{bat}} \\ i_{Lbat}(T_2) = I_{bat} - \frac{d(1-d)V_{in}}{2fL_{bat}}, \end{cases}$$
(5)

where I_{bat} is the battery charging current.

Mode 1 ($0 \le t < T_1$) [Fig. 7(a)]: Q₁ and Q₄ are conducting. The voltages across L and L_{bat}, v_L and v_{Lbat} , are $V_{out}-V_C$ and $-V_{bat}$, respectively. Therefore, i_L in Mode 1 is expressed as

$$i_{L}(t) = i_{L}(0) + \frac{V_{out} - V_{C}}{L}t = i_{L}(0) + \frac{-(1 - 2d)V_{in} + (1 - d)V_{out}}{L}t$$
(6)

C is discharged and charged during i_L is negative and positive, respectively. Since v_L is positive in this mode, L is charged and its current i_L linearly increases. Likewise, i_{Lbat} in Mode 1 (as well as Mode 2) is yielded as

$$i_{Lbat}(t) = i_{Lbat}(0) - \frac{dV_{in}}{L_{bat}}t$$
 (7)

 v_{Lbat} is negative in Modes 1–2, and i_{Lbat} linearly decreases as L_{bat} is discharged.

Mode 2 $(T_1 \le t < T_2)$ [Fig. 7(b)]: This mode begins as Q₃ and Q₄ are turned-on and -off, respectively. v_{Lbat} is still $-V_{bat}$, and i_{Lbat} still decreases. v_L is $V_{in}-V_C$, and hence, i_L in Mode 2 is

$$i_{L}(t) = i_{L}(T_{1}) + \frac{V_{in} - V_{C}}{L} \left(t - T_{1}\right) = i_{L}(T_{1}) + \frac{d\left(2V_{in} - V_{out}\right)}{L} \left(t - T_{1}\right)$$
(8)

Accordingly, i_L increases or decreases depending on the polarity of $(2V_{in}-V_{out})$. For $2V_{in}-V_{out} > 0$, i_L increases as L is charged, and vice versa for $2V_{in}-V_{out} < 0$. On the other hand, C is always charged during entire this mode.

Mode 3 ($T_2 \le t < T_3$) [Fig. 7(c)]: Q₁ and Q₂ are turned-off and -on, respectively, and the L-C circuit is short-circuited. v_L is negative as $-V_C$, and L starts discharging. i_L in Mode 3 is yielded as

$$i_{L}(t) = i_{L}(T_{2}) - \frac{V_{C}}{L} \left(t - T_{2} \right) = i_{L}(T_{2}) - \frac{\left(1 - 2d \right) V_{in} + dV_{out}}{L} \left(t - T_{2} \right)$$
(9)

Meanwhile, v_{Lbat} is positive as $V_{in}-V_{bat}$, and L_{bat} is charged. i_{Lbat} in Modes 3–4 is given by

$$i_{Lbat}(t) = i_{Lbat}(T_2) + \frac{(1-d)V_{in}}{L_{bat}}(t-T_2) \cdot$$
(10)

Mode 4 ($T_3 \le t < T$) [Fig. 7(d)]: Q₃ and Q₄ are turned-off and -on, respectively. v_{Lbat} is still $V_{in}-V_{bat}$, and L_{bat} is still being charged as i_{Lbat} increases. v_L is $-V_{in}+V_{out}-V_C$, and thus, i_L in Mode 4 is

$$i_{L}(t) = i_{L}(T_{3}) + \frac{-V_{in} + V_{out} - V_{C}}{L} \left(t - T_{3}\right) = i_{L}(T_{3}) - \frac{\left(1 - d\right)\left(2V_{in} - V_{out}\right)}{L} \left(t - T_{3}\right)$$
(11)

This equation suggests that L is discharged for $2V_{in}-V_{out} > 0$, and vice versa for $2V_{in}-V_{out} < 0$. Meanwhile, C is discharged during this mode.

C. Output Power

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The output current I_{out} can be determined from i_L in Modes 1 and 4, during which Q_4 conducts and the power is provided to the load, yielding I_{out} and output power P_{out} as

$$I_{out} = -\frac{1}{T} \left(\int_{0}^{T_{1}} i_{L}(t) dt + \int_{T_{3}}^{T} i_{L}(t) dt \right) = \frac{V_{in}}{2fL} \cdot \varphi_{d} \left\{ 2d(1-d) - \varphi_{d} \right\}$$
(12)

$$P_{out} = I_{out}V_{out} = \frac{V_{in}V_{out}}{2fL} \cdot \varphi_d \left\{ 2d\left(1-d\right) - \varphi_d \right\}$$
(13)

 P_{out} in (13) contains not only φ_d but also *d*, suggesting the interdependence between PWM and PS controls.

 P_{out} is normalized by $V_{in}V_{out}/2fL$ to be dimensionless and is plotted as functions of d and φ_d , as shown in Fig. 8. The normalized P_{out} peaks to be 0.0625 at d = 0.5 and $\varphi_d = 0.25$, and it decreases as d moves away from 0.5. Hence, L needs to be



Fig. 7. Operational modes in CC–CV battery charging mode: (a) Mode 1, (b) Mode 2, (c) Mode 3, (d) Mode 4.

determined with considering voltages of V_{in} , V_{out} , and variation ranges of d and φ_d in target applications. A design example of a 200-W prototype for standalone PV systems will be presented in Section VI.

D. RMS Current of Inductor

The RMS current of L, $I_{L_{RMS}}$, is derived from (4), as shown at the bottom of this page.

 $I_{L_{RMS}}$ is important for not only calculating a Joule loss but also determining the optimized control in the battery discharging mode, as will be detailed in Section V-B.

E. ZVS Conditions



Fig. 8. Normalized P_{out} as functions of d and φ_d .

Similar to traditional PS converters [13], parasitic capacitances and body diodes of switches (not shown in figures for the sake of clarity) allow all switches are turned off at zero voltage, achieving ZVS turn-off. Meanwhile, ZVS turn-on conditions depend on current directions at the turn-on moment. The proposed MPCs achieve ZVS turn-on under the condition that body diodes conduct before the switches are turned-on. In other words, the currents flowing through Q_1 – Q_4 must be negative before turning-on. Note that the currents flowing through Q_1 and Q_2 are the sum of i_L and i_{Lbat} . Therefore, the ZVS constraints are given by

$$\begin{cases} Q_{1}: i_{L}(0) - i_{Lbat}(0) < 0\\ Q_{2}: i_{L}(T_{2}) - i_{Lbat}(T_{2}) > 0\\ Q_{3}: i_{L}(T_{1}) > 0\\ Q_{4}: i_{L}(T_{3}) < 0 \end{cases}$$

$$\begin{cases} Q_{1}: k > \frac{(1-d)\{(1-M)(2\varphi_{d}-d) - d(1+l)\}}{\varphi_{d}\{2d(1-d) - \varphi_{d}\}}\\ Q_{2}: k < \frac{d\{(1-M)(1-2\varphi_{d}-d) + (1-d)(1+l)\}}{\varphi_{d}\{2d(1-d) - \varphi_{d}\}},\\ Q_{3}: \varphi_{d} > (1-d)(2-M)/2\\ Q_{4}: \varphi_{d} > d(2-M)/2 \end{cases}$$

$$(15)$$

where k is the current ratio of the battery to load $(k = I_{bat}/I_{out})$, k is the ratio of L to L_{bat} $(l = L/L_{bat})$, and M is the voltage conversion ratio $(M = V_{out}/V_{in})$.

The ZVS boundaries of (15) are shown in Fig. 9. Around M = 2.0, the ZVS operation is feasible at any φ_d . As M moves away from 2.0, ZVS ranges narrow especially when φ_d is small in the light-load region—small φ_d corresponds to low P_{out} as indicated by (13) and shown in Fig. 8. It should be noted that Q₁ and Q₂ determine the ZVS conditions in the range of M > 2.0, while the boundaries at M < 2.0 were due to Q₃ and Q₄. According to Fig. 9(a), the ZVS range is dependent on d. Figure 9(b) suggests that the ZVS range at M > 2.0 (i.e., constraints by Q₁ and Q₂) narrows as k increases. The larger the value of l, the wider will be the ZVS range at M > 2.0, as shown in Fig. 9(c). Thus, ZVS ranges are dependent on M and operation conditions, and ZVS operations are prone to be lost in the light-load region. In the design example in Section VI, the proposed MPC is designed to achieve ZVS operations in the certain heavy-load region with

$$I_{L_{RMS}} = \sqrt{\frac{1}{T}} \int_{0}^{T} i_{L}^{2}(t) dt = \frac{1}{2\sqrt{3}fL} \sqrt{4\varphi_{d}^{2}V_{in}(V_{in} - V_{out})\{\varphi_{d} - 3d(1-d)\} + d^{2}(1-d)^{2}(2V_{in} - V_{out})^{2}}$$
(14)



Fig. 9. ZVS boundaries: (a) k = 2.0, l = 0.1, and d = 0.4-0.6, (b) d = 0.4, l = 0.1, and k = 1.0-3.0, (c) d = 0.4, k = 2.0, and l = 0.1-0.5.

considering target applications, and ZVS operations in the lightload region are compromised.

V. OPERATION ANALYSIS FOR BATTERY DISCHARGING MODE

A. Operation Mode

The input power from the PV panel is zero in the battery discharging mode. The input port can be regarded as an open circuit providing no power, and hence, the SCC-MPC in this mode is equivalent to a cascaded converter comprising the PWM boost converter and PS-SCC with V_{bat} as an input power source. The key operation waveforms and current flows in the battery discharging mode are shown in Figs. 10 and 11, respectively.

The output power in this mode is derived by combining (2) and (13), as

$$P_{out} = \frac{V_{bat}V_{out}}{2fL} \cdot \frac{\varphi_d}{d} \left\{ 2d\left(1-d\right) - \varphi_d \right\}$$
(16)



Fig. 10. Key operation waveforms in battery discharging mode.

There are two control freedoms of d and φ_d in (16), whereas the SCC-MPC in this mode is equivalent to a single-input–single-output converter. In other words, the two control freedoms can be manipulated for the output regulation. With the aim of minimizing the loss in the heavy-load region, an optimal control scheme in the battery discharging mode is proposed in the next subsection.

B. Optimal Control Scheme

In general, switching loss, iron loss, and Joule loss are major loss factors in switching converters. As discussed in Section IV-E, ZVS ranges of the proposed MPC are dependent on d and φ_d , and ZVS operations might be infeasible especially in the lightload region. In the heavy-load region, on the other hand, ZVS ranges widen, and Joule losses generally become dominant. Hence, the Joule loss minimization is a key to enhance efficiency performance in the heavy-load region.

As illustrated in Figs. 10 and 11, i_L always flows through L, C, and either Q₃ or Q₄ [i.e., $i_L(t) = i_{Q3}(t) - i_{Q4}(t)$, see the 3rd panel from the bottom in Fig. 10], and therefore, Joule losses of these components can be simply determined with the inductor's RMS current, $I_{L,RMS}$. As for Q₁ and Q₂, i_L together with i_{Lbat} flows through them, hence $i_L(t) - i_{Lbat}(t) = i_{Q1}(t) - i_{Q2}(t)$ (see the 4th panel from the bottom in Fig. 10). Since i_{Lbat} can be assumed to be a dc current of I_{bat} , the RMS currents of $i_{Q1}(t) - i_{Q2}(t)$ are

 $\sqrt{I_{L_{RMS}}^2 + I_{Lbat}^2}$. I_{Lbat} is simply equal to P_{bat}/V_{bat} , and therefore, its Joule loss is independent on *d* and φ_d . Thus, the minimization

of $I_{L_{RMS}}$ is equivalent to that of Joule losses of L, C, and all switches. Furthermore, since an iron loss of L at a given frequency increases with a peak-to-peak flux density or a peak-to-peak current, the minimization of $I_{L_{RMS}}$ also translates to minimize the iron loss.

The two control freedoms of d and φ_d are determined to minimize $I_{L_{RMS}}$ in order to maximize the power conversion efficiency. The substitution of (2) into (14) yields $I_{L_{RMS}}$ in the battery discharging mode, shown at the bottom in this page.



Fig. 11. Operational modes in battery discharging mode: (a) Mode 1, (b) Mode 2, (c) Mode 3, (d) Mode 4.

We will derive a relational equation that minimizes I_{L_RMS} of (17) with two variables of φ_d and d. However, this equation is impractically complex to solve algebraically. Hence, in the following, the relational equation is derived numerically.

Solving (16) produces d, as

$$d = \frac{1}{2} \left(1 - \frac{A}{2\varphi_d} + \sqrt{\frac{A^2}{2\varphi_d^2} - \frac{A}{\varphi_d} + 1 - 2\varphi_d} \right)$$
(18)

where $A = 2fLP_{out}/V_{bat}V_{out}$. Substitution of (18) into (17) yields the relationship between $I_{L_{RMS}}$ and φ_d at a given value of P_{out} , as shown in Fig. 12(a). These characteristics suggest that, at any P_{out} , an optimal φ_d minimizing $I_{L_{RMS}}$ exists.



Fig. 12. Derivation of optimal φ_d and d at $V_{bat} = 16 \text{ V}$, $V_{out} = 48 \text{ V}$, f = 100 kHz, $L = 3.3 \mu\text{H}$: (a) $I_{L_{RMS}}$ as a function of φ_d , (b) optimal φ_d and d as a function of P_{out} .

The optimal φ_d and *d* as a function of P_{out} are plotted in Fig. 12(b). Both characteristics are almost linear, and their approximated functions can be obtained on the basis of linear approximation, as designated in Fig. 12(b). By arranging these two functions, the relational equation for the optimal φ_d and *d* at $V_{bat} = 16$ V, $V_{out} = 48$ V, f = 100 kHz, and L = 3.3 µH (the same condition as the experiment) can be yielded, as

$$d = -1.21\varphi_d + 0.70 \tag{19}$$

By operating the MPC in the battery discharging mode so that d and φ_d obey (19), the RMS current as well as the Joule loss are minimized, achieving the maximized power conversion efficiency. This optimal control scheme of (19) is implemented with the control block diagram (see Fig. 5) and will be experimentally demonstrated in Section VII-C.

VI. DESIGN EXAMPLE

This section presents a design example of a 200-W experimental prototype. The design target is a standalone PV system with $P_{out} = 100$ W, $P_{bat} = 100$ W, $V_{in} = 30$ V, $V_{out} = 48$ V (i.e., M = 1.6), and $V_{bat} = 12-16$ V at f = 100 kHz. All the circuit elements are assumed ideal to simplify the design procedure.

A. Inductor for Bidirectional PWM Converter

According to the voltage conversion ratio of (2), d varies between 0.40 and 0.54. The inductance L_{bat} for the bidirectional PWM converter is designed considering the largest current ripple ratio.

$$I_{L_{RMS}} = \frac{1}{2\sqrt{3}fL} \sqrt{4\varphi_d^2 \frac{V_{bat}}{d} \left(\frac{V_{bat}}{d} - V_{out}\right)} \left\{\varphi_d - 3d(1-d)\right\} + d^2(1-d)^2 \left(\frac{2V_{bat}}{d} - V_{out}\right)^2$$
(17)

At $V_{bat} = 12$ V, I_{bat} becomes the largest value of 8.33 A. In general, the inductance L_{bat} is designed so that its ripple ratio is around 30%, and therefore,

$$\frac{\Delta I_{Lbat}}{I_{bat}} = \frac{(1-d)V_{bat}T}{I_{bat}L} \approx 30\% \to L$$

$$\approx \frac{0.6 \times 12 \times 10^{-5}}{0.3 \times 8.33} = 28.8 \,\mu\text{H}$$
(20)

where ΔI_{Lbat} is the ripple current of L_{bat}. An inductor with 33 μ H was selected based on (20).

B. Inductor and Capacitor for PS-SCC

As shown in Fig. 9, ZVS operation is not feasible with small φ_d in the light-load region—small φ_d corresponds to low P_{out} as indicated by (13) and shown in Fig. 8. ZVS range might be extended with large $l = L/L_{bat}$, as shown in Fig. 9(c), but circulation currents tend to increase with L and φ_d , eventually resulting in increased Joule loss [37]. Theoretical voltage stresses of switches are lower than $V_{in} = 30$ V for the target application, and therefore, Joule losses would be dominant rather than switching losses.

In this design example, *L* is designed to achieve ZVS operations in the heavy-load region of $P_{out} > 80$ W. According to Fig. 9(a), ZVS operation is feasible with $\varphi_d > 0.10$ in the range of d = 0.40-0.54. For P_{out} to be 80 W with $\varphi_d > 0.10$, *L* was determined to be 3.3 µH based on (13). Once *L* is designed, (13) also yields the maximum $\varphi_d = 0.15$ at $P_{out} = 100$ W and d = 0.40.

The capacitor C is designed large enough so that the resonance between C and L does not influence the PS operation. The capacitance C was determined to be 80 μ F so that the resonant frequency of $1/2\pi\sqrt{LC}$ is approximately one-tenth of f = 100 kHz.

C. Voltage and Current Stresses of Switches

Switches experience different voltage and current stresses depending on operation mods, and therefore, switches need to be selected with considering the largest stress. The currents flowing through Q₁ and Q₂ are the sum of i_L and i_{Lbat} , whereas those of Q₃ and Q₄ are only i_L , as shown in Figs. 7 and 11. Hence, the switch current stresses can be determined from i_L and i_{Lbat} . The voltage stresses of Q₁–Q₂ and Q₃–Q₄ are equal to V_{in} (or the voltage of C_{in}) and $V_{out}-V_{in}$, respectively.

The maximum voltage and current stresses of switches in the CC–CV battery charging mode and battery discharging mode were theoretically derived using the designed parameters in the

| Table III. 1 | Maximum | voltage and | current | values ir | 1 each o | peration | mode. |
|--------------|---------|-------------|---------|-----------|----------|----------|-------|
|--------------|---------|-------------|---------|-----------|----------|----------|-------|

| | CC–CV chargin | Battery g mode | Battery discharging mode | | |
|----------------|------------------|----------------|--------------------------|----------------|--|
| Switch | Voltage [V] | Current [A] | Voltage [V] | Current [A] | |
| Q ₁ | 30.0 | 15.0 | 29.2 | 15.2 | |
| Q ₂ | 30.0 | 15.0 | 29.2 | 14.7 | |
| Q ₃ | 18.0 | 8.07 | 25.1 | 7.82 | |
| Q ₄ | 18.0 | 7.63 | 25.1 | 7.20 | |

Sections VI-A and -B, as listed in Table III. The largest stresses of each switch are highlighted with grey. The MPPT mode, an intermediate mode between the battery charging and discharging modes, was excluded in this table because voltage and current stresses in the MPPT mode are lower than those in other modes. In the battery charging mode, voltage stresses are simply equal to $V_{in} = 30$ V or $V_{out}-V_{in} = 18$ V. In the battery discharging mode, on the other hand, voltage stresses vary because the voltage of C_{in} is dependent on *d*. Current stresses changed with operation modes. In summary, Q₁ and Q₂ are exposed to higher voltage and current stresses, and larger conduction losses (or Joule losses) are expected from these switches.

VII. EXPERIMENTAL RESULTS

A. Prototype and Measured Waveforms

Based on the design procedure presented in Section VI, a 200-W prototype ($P_{out} = 100$ W and $P_{bat} = 100$ W) was built, as shown in Fig. 13, and its component values are listed in Table IV. The prototype was designed for the standalone PV system with $V_{in} = 30$ V, $V_{bat} = 12-16$ V, and $V_{out} = 48$ V. Gate drivers were powered by external auxiliary power supplies. A control card TMS320F28335 (Texas Instruments) was used for feedback control and to generate gating signals at f = 100 kHz.

The measured key operation waveforms at the full load with $V_{bat} = 16$ V are shown in Fig. 14. These waveforms agreed well with the theoretical ones, verifying the operation of the prototype.

Screenshots of the switches' drain-source and gate-source voltages, v_{ds} and v_{gs} , in the battery charging mode at $P_{out} = 100$ W and $P_{bat} = 100$ W are shown in Fig. 15. The measured v_{ds} dropped to zero before corresponding v_{gs} was applied, verifying the ZVS operation for all switches.



Fig. 13. Photograph of 200-W prototype.

| Table IV. Component values. | | | | | | |
|-----------------------------|--|--|--|--|--|--|
| Component | Value | | | | | |
| $Q_1 - Q_4$ | FDD390N15A, $R_{on} = 33.5 \text{ m}\Omega$ | | | | | |
| L | 3.3 µH, $R_{ac} = 120 \text{ m}\Omega$ (at 100 kHz) | | | | | |
| L _{bat} | 33 μ H, R_{dc} = 30 m Ω | | | | | |
| С | Ceramic capacitor, 80 $\mu F,1~m\Omega$ | | | | | |
| C_{in} | Aluminum electrolytic capacitor, 204 $\mu F,5m\Omega$ | | | | | |
| C _{out} | Aluminum electrolytic capacitor, 534 $\mu F,$ 5 m Ω | | | | | |
| C _{bat} | Aluminum electrolytic capacitor, 136 $\mu F,10~m\Omega$ | | | | | |
| Gate driver | IRS2186SPBF, dead-time 100 ns | | | | | |

g. 15. Thotograph of 200-w prototype.

B. Output Characteristics

Measured output characteristics in the CC–CV battery charging mode are compared with theoretical ones [see (13) and (2)] in Fig. 16. The measured P_{out} characteristics, as shown in Fig. 16(a), were dependent on both d and φ_d , as explained in



Fig. 14. Measured key waveforms in (a) battery charging mode at $P_{out} = 100$ W and $P_{bat} = 100$ W, and (b) battery discharging mode at $P_{out} = 100$ W.



Fig. 15. ZVS waveforms in battery charging mode at $P_{out} = 100$ W and $P_{bat} = 100$ W: (a) Q_1 , (b) Q_2 , (c) Q_3 , (d) Q_4 .

Section IV-C. Although slight disagreement due to neglected parameters, such as dead-time periods and parasitic components, was observed, the measured and theoretical characteristics satisfactorily agreed.

Figure 16(b) shows and compares the measured and theoretical V_{bat} characteristics as a function of *d*. V_{bat} was independent on φ_d , verifying that the PS control does not affect the V_{bat} regulation, as discussed in Section IV-A.

The measured output characteristics in the battery discharging mode are shown in Fig. 17. The characteristics with the fixed values of *d* exhibited that P_{out} was dependent on both *d* and φ_d , as indicated by (16). With the optimal control of (19), P_{out} linearly increased with φ_d , agreeing with the characteristic shown in Fig. 12(b).

C. Power Conversion Efficiencies

The measured and calculated power conversion efficiencies in the CC–CV battery charging mode at fixed P_{bat} and P_{out} are shown in Figs. 18(a) and (b), respectively—the theoretical loss model was derived but is not shown in this paper for the sake of



Fig. 16. Measured and theoretical output characteristics in battery charging mode: (a) P_{out} as a function of φ_d , (b) V_{bat} as a function of d.



Fig. 17. Measured output characteristics in battery discharging mode.



Fig. 18. Measured and calculated power conversion efficiencies in battery charging mode at (a) fixed P_{bat} and (b) fixed P_{out} .



Fig. 19. Measured power conversion efficiencies in battery discharging mode with optimized and fixed-*d* controls.

page length. Losses due to gate driving are excluded from these results as gate drivers were powered by external auxiliary power supplies. The measured and calculated efficiencies showed good agreement, verifying the theoretical loss model, based on which loss breakdowns will be discussed in Section VII-D. The measured efficiencies deteriorated in the light-load region due to the iron loss. In the medium- to heavy-load regions, on the other hand, efficiencies were greater than 95%. The efficiency at the full load of 200 W was as high as 95.7%.

The measured power conversion efficiencies in the battery discharging mode are shown in Fig. 19. With the fixed-*d* control, efficiencies dropped in the light- and heavy-load regions. The optimal control, on the other hand, improved efficiencies in the entire region, demonstrating the efficacy of the proposed optimal control scheme. The efficiency at the full load of 100 W was as high as 91.7%. The efficiencies under the heavy-load conditions in the battery discharging mode were inferior to those in the battery charging mode because the proposed MPC



Fig. 20. Estimated loss breakdowns in battery charging mode at (a) fixed $P_{bat} =$ 100 W and (b) fixed $P_{out} =$ 100 W.

in the battery discharging mode is equivalent to a two-stage converter comprising a boost converter and PS-SCC, as discussed in Section V-A.

D. Loss Analysis

The estimated loss breakdowns based on the theoretical loss model in the CC–CV battery charging mode are shown in Fig. 20. Iron losses were calculated based on Steinmetz's equation [38]. The switching loss was assumed zero within the ZVS region specified in Fig. 9. Switching losses occurred when $P_{out} = 20$ W and 60 W [see Fig. 20(a)] because the prototype was designed to achieve ZVS operations under the heavy-load condition of $P_{out} >$ 80 W, as discussed in Section VI-B. However, the portion of the switching losses was very minor in comparison with Joule losses. The Joule loss of L was the most dominant factor in the entire range because of its relatively large resistance due to the skin effect at 100 kHz ($R_{ac} = 120 \text{ m}\Omega$). Meanwhile, the Joule loss of Q₂ also took a significant portion because both i_L and i_{Lbat} flowed through Q₂.

E. Transient Response Characteristics

To investigate the influence of the interdependence between PS and PWM controls in the CC–CV battery charging mode (see Section IV-C), transient response characteristics were measured with applying step changes in I_{out} and I_{bat} , as shown in Fig. 21. V_{out} and V_{bat} were regulated to be 48 V and 16 V by PS and PWM controls, respectively, while I_{out} or I_{bat} abruptly increased from 50 W to 100 W using an electronic load operating in a constant-current mode. Measured transient response characteristics are shown in Fig. 21. V_{out} slightly



Fig. 21. Measured transient response characteristics: (a) step change in I_{outb} (b) step change in I_{bar} .

dropped by 1.2 V in response to the step change in I_{out} [see Fig. 21(a)], while V_{bat} was unaffected. Similarly, the step change in I_{bat} slightly affected V_{bat} only [see Fig 21(b)], and V_{out} was independent on I_{bat} . In summary, both V_{bat} and V_{out} were tightly regulated during the transient and were nearly unaffected by step changes in I_{out} and I_{bat} , suggesting the insignificant interdependence between V_{bat} and V_{out} regulation of the proposed converter.

F. Power Balance Test with MPPT Control

A power balance test with MPPT control was performed to verify the operation in the MPPT mode. Instead of an actual PV panel, a solar array simulator was used as the input power source. A hill-climbing method was employed as the MPPT algorithm with a duty cycle perturbation Δd of 1% and a sampling interval of 1.0 s. The input and output voltages were regulated by PWM and PS controls, respectively, as shown in Fig. 5. The maximum power of the PV panel was 60 W, while P_{out} swung between 40 W and 80 W.

Figure 22 shows the measured P_{in} , P_{out} , and P_{bat} in the power balance test. During Periods 1 and 3 ($P_{in} > P_{out}$), P_{bat} was positive, and the battery was charged. As P_{out} surpassed P_{in} in Period 2, the battery discharged as P_{bat} was negative. The PV panel kept generating its maximum power of 60 W thanks to the MPPT control, while the battery charged and discharged depending on the power balance between P_{in} and P_{out} . The test result demonstrated that the proposed SCC-MPC could smoothly switch the charging and discharging of the battery in the MPPT mode.

G. Mode Transitions among CC–CV Battery Charging, Battery Discharging, and MPPT Modes

The mode transition test was performed using an electronic double-layer capacitor (EDLC) module with a capacitance of



Fig. 22. Experimental results of power balance test with MPPT control.

50 F as a rechargeable battery, as shown in Fig. 23(a). The PV panel (i.e., the solar array simulator) with a maximum power of 60 W was used as the input source, and the EDLC module was charged with the CC–CV charging scheme of 1.0 A–16 V. The load power P_{out} was fixed to be 43 W at $V_{out} = 48$ V and $I_{out} = 0.9$ A, while the PV panel was disabled in the middle of the test in order to force the MPC to operate in the battery discharging mode.

The resultant voltage, current, and power profiles are shown in Fig. 23(b). At the beginning of the test, the MPC operated in the CC charging mode. As the EDLC module was charged with a constant current of 1.0 A, V_{bat} increased almost linearly until P_{in} reached the maximum power of 60 W. The maximum power of the PV panel was tracked in the MPPT mode until V_{bat} reached the CV level of 16 V. Since then, the operation shifted to the CV charging mode, and I_{bat} gradually declined. After I_{bat} was tapered to zero, the PV panel was disabled, and the EDLC module started discharging to supply P_{out} . V_{in} in the battery discharging mode corresponded to the voltage of C_{in}. After the 30-s discharging, the PV panel was enabled, and the EDLC module was charged again in the MPPT mode. In summary, all the operation modes were switched with the proposed MPC and control block, while V_{out} was regulated to be 48 V.



Fig. 23. (a) Experimental setup for mode transition test. (b) Resultant voltage, current and power profiles.

| Topology | Component Count | | | | Control Full Load | | Conversion Ratios | | Operational |
|----------|-----------------|---|---|----|-------------------|------------|---------------------|-----------------------|---|
| | Q | D | L | C | Scheme | Efficiency | 1 | Z | Constraint |
| [18] | 6 | 0 | 2 | 3 | PWM | 94% | $\frac{d_1}{1-d_1}$ | $rac{d_2}{1-d_1}$ | $d_2 > d_1$ |
| [20] | 3 | 0 | 2 | 3 | PWM | > 93% | $\frac{1}{1-d_1}$ | $\frac{1}{1-d_2}$ | $d_2 > d_1$ |
| [21] | 3 | 1 | 2 | 4 | PWM | 97% | $\frac{1}{2-d_1}$ | $\frac{d_2}{2-d_1}$ | $d_1 > d_2$ |
| [24] | 3 | 1 | 3 | 4 | PWM | 97% | $\frac{1}{1-d_1}$ | $\frac{1-d_2}{1-d_1}$ | $d_1 > d_2$ |
| [22] | 6 | 4 | 3 | 10 | PWM PFM | 94% | $\frac{1+d_1}{3}$ | See [22] | $\frac{f_s}{2f_r} \le d \le 1 - \frac{f_s}{2f_r}$ |
| Proposed | 4 | 0 | 2 | 4 | PWM PS | 95.7% | d | Eq. (13) | — |

Table V. Comparison among nonisolated MPCs.

VIII. COMPARISON WITH CONVENTIONAL NONISOLATED MPCs

The proposed MPC and conventional nonisolated MPCs are compared from various aspects, as shown in Table V. Most conventional nonisolated MPCs are based on PWM converters [18], [20], [21], [24], and their conversion ratios are dependent on two duty cycles of d_1 and d_2 . In these PWM MPCs, however, two control freedoms of d_1 and d_2 impose operational constraints. The conventional PWM MPCs can operate only when the constraints are satisfied, and these constraints substantially narrow conversion ranges. An MPC employing PWM and PFM control schemes has also been proposed [22], but its regulation range is inherently narrow due to its resonant operation.

The proposed MPC is comparable with conventional nonisolated MPCs from the viewpoints of component counts and full load efficiency. The noticeable difference is that the proposed MPC can operate without an operational constraint. Although the output power P_{out} is dependent on duty cycle *d* [see (13)], the MPC can work at any *d*.

IX. CONCLUSIONS

The nonisolated SCC-MPCs integrating PWM converter and PS-SCC have been proposed. A bidirectional PWM converter and PS-SCC are integrated into a single unit with reducing the total switch count, achieving the simplified circuit. Previouslyreported SCCs, such as a ladder, Dickson, series-parallel, and Fibonacci SCCs, can be used as a PS-SCC, and various types of SCC-MPC topologies can be derived based on the proposed integration procedure.

The detailed operation analysis was performed to mathematically derive the gain characteristics and ZVS boundaries in the battery charging and discharging modes. The optimal control scheme for the battery discharging mode, in which two control freedoms of duty cycle *d* and phase-shift angle φ_d are available to regulate the output, was also proposed. The optimal *d* and φ_d are determined depending on the output power so as to minimize the RMS current of the inductor and to maximize the power conversion efficiency.

The experimental verification tests using the 200-W prototype were performed, and the results verified the theoretical operation analysis as well as the proposed optimal

control scheme for the battery discharging mode. The power balance test with the MPPT control and mode transition test demonstrated that the battery charging, discharging, and MPPT could be smoothly switched with the SCC-MPC and control block.

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