Comprehensive Analysis and Gain Derivation of Phase-Shifted Dual-Input LLC Converter

ABDULLAH ALHATLANI, (Member, IEEE),
SUMANA GHOSH, (Graduate Student Member, IEEE),
AND ISSA BATARSEH, (Fellow, IEEE)

1Department of Electrical Engineering, Imam Mohammad Ibn Saud Islamic University (IMSIU), Riyadh 11564, Saudi Arabia
2Department of Electrical and Computer Engineering, University of Central Florida, Orlando, FL 32816, USA

Corresponding author: Abdullah Alhatlani (a.n.hatlani@gmail.com)

This work was supported in part by Grant NSF-ECCS-1810733 and in part by the University of Central Florida (UCF) College of Graduate Studies.

ABSTRACT This paper presents a comprehensive time-domain analysis of the dual-input LLC converter. By utilizing PSM and FSM, the converter power stage interfaces two photovoltaic (PV) panels to achieve maximum power point tracking (MPPT). The detailed mode-by-mode operation and steady-state analysis are derived using three parameters: the switching frequency, phase shift, and load weight. Furthermore, the DC gain equations of each PV panel and their characteristic curves are derived. In addition, the waveform boundaries and ZVS analysis are discussed as well. Moreover, an extended perturbation & observation (P&O) algorithm that guarantees achieving MPPT for each panel is proposed. Finally, the theoretical analysis has been tested and verified by a 450-W prototype of dual-input LLC converter.

INDEX TERMS Characteristic curve, DC gain, dual-input, frequency-shift modulation, LLC resonant converter, phase-shift modulation, maximum power point tracking.

I. INTRODUCTION

Solar energy is becoming very popular as a green energy source because of its redundancy and cost-effectiveness [1]. Total capacity of grid-connected PV power systems has grown dramatically from 300 MW in 2000 to 627 GW in 2019. At the end of 2019, the total solar PV capacity increased to 139 GW, for an estimated total of 760 GW with a huge growth in global market [2].

In a grid-tied PV system, the inverter can be either a central inverter, a string inverter, or a micro-inverter. Compared with the other two configurations, micro-inverters provide many advantages like capability against partial shadow, flexibility, convenience for enlarging capacity, replacement, and safety [3]. Because of its well-understood operation and simple control functionality, a two-stage grid-tied micro inverter is the most commonly used topology.

In this paper, the focus is on the DC-DC converter stage of the micro-inverter. Various multi-port DC-DC converter models and control methods have been discussed in [4], [5], and [6]. Among these converters, the most efficient and useful one is the LLC resonant converter. Many different models of the full-bridge LLC converter were developed to achieve a wide input-voltage range [7], [8]. LLC converters have various advantages like low electromagnetic interference, high power density, high efficiency, and most importantly soft switching facility [9], [10]. In [11], a detailed analysis of different operating modes was presented, which gave a clear idea of resonant current and dc gain characteristics for a single input LLC converter. The converter can operate at resonance in normal input voltage range, and hence zero-voltage switching (ZVS) for primary side switches and zero-current switching (ZCS) in secondary side diodes can be realized. The operation of the converter can be separated with respect to the switching frequency. When the switching frequency \( f_s \) is below the resonant frequency \( f_r \), the converter can be used in a step-up operation while it can be used for a step-down operation when \( f_s \) is above \( f_r \). Nevertheless, the maximum efficiency occurs when \( f_s \) operates at resonant [12]. In [13], the effect of phase shift on single-input LLC converter at a fixed frequency and its impact on ZVS was studied. On the other hand, the impact of phase shift on the gain curve for the conventional LLC converter was proposed in [14].
In [15], the dual-input LLC converter shown in Fig. 1 was proposed. Since the converter interfaces two PV panels with the load, two degrees of freedom are needed for a regulation [16]. While the general technique to regulate the load power by using frequency-shift modulation (FSM), the phase-shift modulation (PSM) was utilized to achieve MPPT for each of the two PVs in [15]. It showed that a negative phase shift is needed when PV1 voltage (\(V_{PV1}\)) is greater than PV2 voltage (\(V_{PV2}\)), otherwise, a positive phase shift should be applied to maintain the MPPT. Moreover, PSM was utilized first to balance the input power before regulating the load by FSM since it has the dominant effect on the output voltage. However, the gain characteristics were not presented since the conventional Fundamental Harmonic Approximation (FHA) does not consider the phase shift nor voltage difference between the PV panels. Furthermore, MPPT cannot be guaranteed since the applied control strategy relied only on the input power.

In [17], the gain characteristics for the dual-input LLC converter were derived in the normalized form. In addition, the steady-state modes of operation are demonstrated. Moreover, it discussed the possible waveforms for the tank components. Finally, the theoretical analysis was verified by simulation results.

The main difference between the single input full bridge LLC converter and this Dual-input LLC is in the way they interface the input ports. The major drawback of the PV microinverter system is the cost. In a single input full bridge LLC converter, the resonant tank transfers the energy from a single source. Hence two full bridge converters are needed for transferring the energy from two sources. Whereas this dual PV architecture reduces the component count as well as the cost by utilizing a single resonant tank for interfacing two PV sources. It is advantageous to design multi-input converters without increasing the number of circuit components, as they would increase the power density of the overall system. Also, all the ports can be controlled with a centralized controller, which eliminates the need for additional controller for each PV interface.

Due to the intermittency nature of the PV panels, implementation of MPPT algorithms has become an important part of the control for solar panels. For a PV microinverter, many hybrid control techniques are proposed in the literature to track the Maximum Power Point (MPP) [18], [19], [20]. Different multiple phase-shift modulation techniques are used in some applications [21], [22]. The proposed topology can be controlled via any two parameters: duty cycle, switching frequency and phase shift based on MPPT control methods used. In this paper the classical MPPT methods like Perturbation and Observation (P&O) has been utilized to develop an improved MPPT algorithm.

This paper is an expanded version of the paper proposed in [17]. The main contributions of this this paper are discussed in following sections. This paper contributes a detailed analysis of the conditions that form the switching cycle waveforms. Moreover, the waveform boundaries and their different conditions are demonstrated. Also, the PSM impact on the ZVS range for the proposed dual-input LLC converter is discussed. Furthermore, it proposes an improved P&O technique that guarantees achieving MPPT for both panels. Finally, the experimental results are presented to verify the theoretical analysis.

The paper is structured as follows: Section II presents the modes of operations and the switching-cycle waveforms are discussed in Section III. Then, a mode-by-mode normalized analysis for the general waveform is illustrated in Section IV and followed by deriving the DC gain equations and characteristic curves in Section V. Section VI illustrates the waveform boundaries along with the ZVS analysis, and the MPPT control algorithm is presented in Section VII. The experimental results are shown in Section VIII and the conclusions are given in Section IX.

### II. STEADY-STATE MODES OF OPERATION

There are four general modes of operation for the proposed dual-input LLC converter when it operates in the steady-state. Three of these modes occur when either \(V_{PV1}\) or \(-V_{PV2}\) is applied, so they are similar to the modes discussed in the conventional LLC converter [11]. On the other hand, the fourth mode occurs due to PSM.

First, Positive clamped mode (P-mode) appears when the magnetizing inductor (\(L_m\)) is positively clamped to the output as shown in Fig. 2(a), and the input voltage (\(V_{in}\)) is either \(V_{PV1}\) or \(-V_{PV2}\) for the positive or negative half-cycle, respectively. In this mode, the resonant inductor (\(L_r\)) and capacitor (\(C_r\)) resonate under the resonance frequency (\(f_r = \frac{1}{2\pi \sqrt{L_r C_r}}\)).

The normalized magnetizing inductor voltage (\(v_{nr}\)) equals 1 while the normalized resonant capacitor voltage (\(v_{nr}\)) equals \(v_r/nV_o\), normalized resonant inductor current (\(i_{nr}\)) equals \((i_rZ_r/nV_o)\) and magnetizing inductor current (\(i_{nm}\)) can be expressed as

\[
\begin{align*}
\frac{v_{nr}(\theta)}{M_{nr}} &= \frac{1}{M_{nr}} - 1 + i_{nr}(\theta_0) \sin(\theta - \theta_0) \\
&- \left[\frac{1}{M_{nr}} - 1 - v_{nr}(\theta_0)\right] \cos(\theta - \theta_0) \\
i_{nr}(\theta) &= i_{nr}(\theta_0) \cos(\theta - \theta_0) \\
&+ \left[\frac{1}{M_{nr}} - 1 - v_{nr}(\theta_0)\right] \sin(\theta - \theta_0) \\
i_{nm}(\theta) &= \frac{\theta - \theta_0}{l} + i_{nm}(\theta_0)
\end{align*}
\]
where \( \theta = 2\pi f_s t \) is the normalized time, \( M_{in} = nV_o/V_{in} \) is the DC gain, \( l = L_m/L_r \) is the inductor ratio, \( n = N_1/N_2 \) is the turns ratio, \( Z_r = \sqrt{L_r/C_r} \) is the characteristic impedance, and \( \theta_0 \) is the normalized initial time of the mode.

The second mode is the Negative clamped (N-mode), which is similar to the previous one except for the negative clamped of \( L_m \) to the output, as in Fig. 2(b). This reverses the polarity of \( nV_o \) in all the normalized equations, so \( v_{nm} = -1 \) while \( v_{nr}, i_{nr} \) and \( i_{nm} \) are similar to the equations in (1) except that the term \((-1)\) becomes \((+1)\).

Then, there is the cut-Off mode (O-mode) in which \( i_{nr} = i_{nm} \) since \( L_m \) resonates with \( L_r \) and \( C_r \) at the frequency \( f_m = 1/(2\pi \sqrt{(L_r + L_m)/C_r}) \), and there is no power flow to the output, as in Fig. 2(c). As in P and N-mode, \( V_{in} \) is either \( V_{PV1} \) or \( V_{PV2} \) in this mode. This leads to the equations in (2).

As a result, S-mode occurs, which is similar to O-mode except that \( V_{in} \) is either \((V_{PV1} - V_{PV2})\) or zero. This means there is no power flow to the output in this mode as well and \( L_m \) resonates with \( L_r \) and \( C_r \) at \( f_m \). Accordingly, \( i_{sr} = i_{sm} \) through the whole period of this mode. Unlike O-mode, the tank current is always decreasing, during the positive half cycle of the steady state, due to the low input voltage in S-mode.

If a \(-\phi\) is applied, \( PV_2 \) would be connected to the tank before the end of the positive half cycle while \( PV_1 \) is still connected. Accordingly, \( V_{in} \) in Fig. 2(c) would become \((V_{PV1} - V_{PV2})\) forming a negative phase shift mode (\( S^- \)-mode). This leads to the equations shown in (3) for \( S^- \)-mode,

\[
\begin{align*}
v_{nrS^-} (\theta) &= \frac{1}{M_1} - \frac{1}{M_2} + i_{nr} (\theta_0) \sqrt{1+l} \sin \left( \frac{\theta - \theta_0}{\sqrt{1+l}} \right) \\
&\quad - \left[ \frac{1}{M_1} - \frac{1}{M_2} - v_{nr} (\theta_0) \right] \cos \left( \frac{\theta - \theta_0}{\sqrt{1+l}} \right) \\
i_{nrS^-} (\theta) &= i_{nr} (\theta_0) \cos \left( \frac{\theta - \theta_0}{\sqrt{1+l}} \right) + \left[ \frac{1}{M_1} - \frac{1}{M_2} - v_{nr} (\theta_0) \right] \sin \left( \frac{\theta - \theta_0}{\sqrt{1+l}} \right) \\
v_{nmS^-} (\theta) &= -i_{nr} (\theta_0) \frac{l}{\sqrt{1+l}} \sin \left( \frac{\theta - \theta_0}{\sqrt{1+l}} \right) \\
&\quad + \frac{l}{l+1} \left[ \frac{1}{M_1} - \frac{1}{M_2} - v_{nr} (\theta_0) \right] \cos \left( \frac{\theta - \theta_0}{\sqrt{1+l}} \right)
\end{align*}
\]

where \( M_1 \) and \( M_2 \) are the DC gain of \( PV_1 \) and \( PV_2 \), respectively.

In contrast, \( PV_2 \) would be connected to the tank after disconnecting \( PV_1 \) when a \(+\phi\) is applied. This leads to a free-wheeling period for the tank with zero input voltage forming a positive phase shift mode (\( S^+ \)-mode). Accordingly, the equations of this mode are like \( S^- \), but without \((1/M_1 - 1/M_2)\) term.

### III. SWITCHING CYCLE WAVEFORMS

In steady state, the resonant tank components waveforms shape is a combination of up to three modes from P, N, O, and S-mode illustrated in the previous section. For example, the tank starts with P-mode followed by O-mode forming a PO waveform when it operates below resonance with no phase shift. Similarly, other waveforms appear under different conditions. The control factors of the waveform shape in the proposed dual-input LLC are the switching frequency \( f_s \), phase shift \( \phi \), and load weight. When \( \phi = 0 \), both PVs produce the same power, the converter acts as a single full-bridge LLC converter. Accordingly, the half switching cycle shape would be either NOP, NP, P, PO, PON, OPO, or PN as illustrated in [11]. However, new waveform shapes occur when \( \phi \neq 0 \), and they are demonstrated in this section. Furthermore, all the possible waveforms under the different conditions are summarized in Table. 1.

---

**FIGURE 2.** The equivalent circuits of the four different modes: (a) P-mode, (b) N-mode, and (c) O-mode and S-mode.
TABLE 1. Half switching cycle waveforms conditions.

| Waveform Shape | \( f_s \) | \( \varphi \) | Load Weight |
|----------------|--------|--------|------------|
| NP            | \( < f_r \) | 0      | normal     |
| P             | \( > f_r \) | \( +\varphi \) or \(-\varphi\) | too heavy  |
| PS            | \( f_r \) | \( +\varphi \) or \(-\varphi\) | normal     |
| OPO           | \( < f_r \) | 0      | normal     |
| PON           | \( < f_r \) | \( +\varphi \) or \(-\varphi\) | light      |
| PN            | \( < f_r \) | \( +\varphi \) or \(-\varphi\) | too heavy  |

A. PS WAVEFORM

When the two PVs are unbalanced and the normalized frequency \( f_{ns} = f_s/f_r = 1 \), PS waveform appears due to the applied \( \varphi \). At the beginning of the positive half cycle, \( i_{tr} = i_{am}, V_{PV1} \) is connected to the tank and the power is delivered to the load. As a result, \( L_m \) is positively clamped to the output with \( v_{nm} = 1 \) while \( L_r \) resonate with \( C_r \). Accordingly, \( i_{am} \) increases linearly while \( i_{tr} \) increases in a sinusoidal shape forming a P-mode. At \((\pi f_{ns} - |\varphi_n|)\), where \( \varphi_n = \varphi/f_{ns} \) is the normalized phase shift, the input voltage is reduced to either \((V_{PV1} - V_{PV2})\) or zero in case of \(-\varphi\) or \(+\varphi\), respectively, which is not enough to deliver the power to the output. Accordingly, \( i_{tr} \) rapidly decreases till \( i_{am} = i_{nm} \), usually during the dead time, and they both begin to nonlinearly decrease as \( L_m \) resonates with \( L_r \) and \( C_r \) entering S-mode. Hence, \( v_{nm} \) nonlinearly decreases below 0 due to the low input voltage to the tank. S-mode ends at the end of the positive half cycle as shown in Fig. 3(a). Similarly, P-mode occurs at the beginning of the negative half cycle when \(-V_{PV2}\) is connected to the tank. Then, it would be followed by S-mode.

B. POS WAVEFORM

This waveform occurs under the normal-load condition when \( f_{ns} < 1 \) as the resonance period of \( L_r \) and \( C_r \) is shorter than the half cycle. Consequently, the waveform begins with P-mode as in A, but \( i_{tr} \) and \( i_{am} \) converge before S-mode begins. Therefore, O-mode occurs between P and S-mode while \( V_{PV1} \) is still connected to the tank, as shown in Fig. 3(b). O-mode starts when \( i_{tr} = i_{am} \) at \( \theta_r/2 = \pi \), where \( \theta_r \) is the normalized resonance time, and there is no power flow to the load. As \( v_{nm} \) decreases below 1 in this mode, \( i_{tr} = i_{am} \) nonlinearly increase since the input voltage is \( V_{PV1} \) while \( L_m \) is resonating with \( L_r \) and \( C_r \). After that, S-mode begins at \((\pi f_{ns} - |\varphi_n|)\). Accordingly, this waveform cannot appear if the condition in (4) is not satisfied, otherwise, it would become a PS waveform.

\[
\pi f_{ns} - |\varphi_n| > \pi \tag{4}
\]

C. PNO WAVEFORM

The phase shift has a different impact on the steady state waveform when \( f_{ns} > 1 \). As illustrated in [11], NP waveform appears under normal-load condition above resonance. However, applying \( \varphi \) results in shifting the modes to shape a PNO waveform as shown in Fig. 3(c). First, P-mode occurs as \( V_{PV1} \) is connected to the tank and \( v_{nm} = 1 \). At \((\pi f_{ns} - |\varphi_n|)\), the input voltage is reduced to either \((V_{PV1} - V_{PV2})\) or zero in case of \(-\varphi\) or \(+\varphi\), respectively, and \( v_{nm} \) switches from 1 to -1 forming N-mode. During this mode, \( i_{nr} \) decreases in a sinusoidal shape while \( i_{am} \) linearly increases. When \( i_{nr} = i_{nm} \) at the end of N-mode, \( L_m \) resonates with \( L_r \) and O-mode begins. Due to the low input voltage during this mode, \( v_{nm} \) begins to nonlinearly decrease below 0 as in Fig. 3(c). However, NP appears under too heavy-load conditions even with applying \( \varphi \).

D. PON \( n \) WAVEFORM

As the load gets heavier or \( f_s \) goes farther below resonance, \( v_{nm} \) gets closer to -1 during S-mode. When \( v_{nm} = -1 \) before the end of S-mode, the POS waveform becomes PON \( n \) as shown in Fig. 3(d). Accordingly, P-mode has different initial values for \( i_{tr} \) and \( i_{am} \) and the shorter period since \( i_{tr} \) and \( i_{am} \) diverge from \( i_{nm} \) during the N-mode before the switching cycle ends. After P-mode, O-mode starts when \( i_{tr} = i_{nm} \) as \( v_{nm} \) decreases below 0 due to the heavy load. Accordingly, \( i_{tr} = i_{nm} \) nonlinearly decrease during this mode. Finally, N-mode starts at \((\pi f_{ns} - \varphi_n)\), so this mode is called N \( n \) since the converter operates in N-mode during the phase shift period.

E. OPOS WAVEFORM

Under light-load condition when \( f_{ns} < 1 \), the switching cycle starts with O-mode since \( v_{nm} \) cannot reach 1 at the beginning. When \( v_{nm} = 1 \), P-mode begins and \( i_{am} \) diverge from \( i_{nm} \) as in POS. As a result, an OPOS waveform appears as shown in Fig. 3(e). Unlike OPO waveform in the conventional LLC converter, OPOS operates only below resonance. On the other hand, PS waveform occurs above resonance under light-load.

IV. MODE BY MODE NORMALIZED ANALYSIS

In order to derive the gain equation, a mode by mode analysis is needed. The POS switching cycle waveform is analyzed since it covers most of the possible conditions for the above unity gain region, where \( f_s \leq f_r \), besides, it is the only waveform with closed-form solution. By assuming \( V_{PV1} = V_{PV2} \), a negative phase shift is applied to balance the input power. Following the switching cycle of POS shown in Fig. 4, the steady-state mode by mode normalized analysis for one cycle can be divided into six modes, which can be illustrated as follow:

Mode 1 (\( \theta_0, \theta_1 \)): In this mode, the switches S1 and S4 are ON while S2 and S3 are OFF, as shown in Fig. 5. Therefore, PV1 supplies the power to the output, and \( L_m \) is positively clamped to the output forming a P-mode. This interval ends...
when \( i_{nr} = i_{nm} \) at \( \theta_1 = \theta_0 + \pi \), where \( \theta_0 \) is the initial normalized time, and the tank components normalized equations can be defined as in (5).

\[
\begin{align*}
v_{nr}(\theta) &= \frac{1}{M_1} - 1 + i_{nr}(\theta_0) \sin (\theta - \theta_0) \\
&\quad - \left[ \frac{1}{M_1} - 1 - v_{nr}(\theta_0) \right] \cos (\theta - \theta_0) \\
i_{nr}(\theta) &= i_{nr}(\theta_0) \cos (\theta - \theta_0) \\
&\quad + \left[ \frac{1}{M_1} - 1 - v_{nr}(\theta_0) \right] \sin (\theta - \theta_0) \\
i_{nm}(\theta) &= \frac{\theta - \theta_0}{i} + i_{am}(\theta_0) \\
\end{align*}
\]

Mode 2 (\( \theta_1, \theta_2 \)): In this mode, the switches remain in the same state as the previous mode, but there is no output power flow as long as \( i_{nr} = i_{am} \), shown in Fig. 6. This means the converter is in the cut-off mode forming O-mode, and the tank components normalized equations can be defined as in (6).

\[
\begin{align*}
v_{nr}(\theta) &= \frac{1}{M_1} + i_{nr}(\theta_1) \sqrt{1 + l} \sin \left( \frac{\theta - \theta_1}{\sqrt{1 + l}} \right) \\
&\quad - \left[ \frac{1}{M_1} - v_{nr}(\theta_1) \right] \cos \left( \frac{\theta - \theta_1}{\sqrt{1 + l}} \right) \\
i_{nr}(\theta) &= i_{nr}(\theta_1) \cos \left( \frac{\theta - \theta_1}{\sqrt{1 + l}} \right) \\
&\quad + \frac{1}{\sqrt{1 + l}} \left[ \frac{1}{M_1} - v_{nr}(\theta_1) \right] \sin \left( \frac{\theta - \theta_1}{\sqrt{1 + l}} \right) \\
i_{nm}(\theta) &= \frac{\theta_1 - \theta_0}{i} + i_{am}(\theta_0) \\
\end{align*}
\]
where the initial values can be derived from the equations of the previous mode as

\[
\begin{align*}
\nu_{nr} (\theta_1) &= \nu_{nr} (\theta_0) \cos (\theta_1 - \theta_0) \\
&+ \left[ \frac{1}{M_1} - 1 - \frac{v_r (\theta_0)}{n V_o} \right] \sin (\theta_1 - \theta_0) \\
\nu_{nr} (\theta_1) &= \frac{1}{M_1} - 1 + \nu_{nr} (\theta_0) \sin (\theta_1 - \theta_0) \\
&- \left[ \frac{1}{M_1} - 1 - \nu_{nr} (\theta_0) \right] \cos (\theta_1 - \theta_0)
\end{align*}
\]

Mode 3 (\(\theta_2, \pi / f_{ns}\)): This mode occurs due to the phase shift, so the period between \(\theta_2\) and the normalized half-cycle \((\pi / f_{ns})\) equals to the normalized phase shift \(\phi_{ns}\). In this mode, the switches \(S_1\) and \(S_3\) are ON while \(S_2\) and \(S_4\) are OFF, so the tank is fed with both PVs as shown in Fig. 7. However, there is still no output power flow in this mode, which means this interval represents the S' mode, and the tank normalized equations can be defined as in (7).

\[
\begin{align*}
\nu_{nr} (\theta) &= \frac{1}{M_1} - 1 - \nu_{nr} (\theta_2) \sqrt{1 + l} \sin \left( \frac{\theta - \theta_2}{\sqrt{1 + l}} \right) \\
&- \left[ \frac{1}{M_1} - 1 - \nu_{nr} (\theta_2) \right] \cos \left( \frac{\theta - \theta_2}{\sqrt{1 + l}} \right) \\
i_{nr} (\theta) &= i_{nr} (\theta_2) \cos \left( \frac{\theta - \theta_2}{\sqrt{1 + l}} \right) \\
&+ \left[ \frac{1}{M_1} - 1 - \nu_{nr} (\theta_2) \right] \sin \left( \frac{\theta - \theta_2}{\sqrt{1 + l}} \right)
\end{align*}
\]

where the initial values can be derived from the equations of the previous mode as

\[
\begin{align*}
i_{nr} (\theta_2) &= i_{nr} (\theta_0) \cos \left( \frac{\theta_2 - \theta_1}{\sqrt{1 + l}} \right) \\
&+ \frac{1}{\sqrt{1 + l}} \left[ \frac{1}{M_1} - \nu_{nr} (\theta_1) \right] \sin \left( \frac{\theta_2 - \theta_1}{\sqrt{1 + l}} \right) \\
v_{nr} (\theta_2) &= \frac{1}{M_1} - 1 + i_{nr} (\theta_1) \sqrt{1 + l} \sin \left( \frac{\theta_2 - \theta_1}{\sqrt{1 + l}} \right) \\
&- \left[ \frac{1}{M_1} - \nu_{nr} (\theta_1) \right] \cos \left( \frac{\theta_2 - \theta_1}{\sqrt{1 + l}} \right)
\end{align*}
\]

Mode 4 (\(\pi / f_{ns}, \theta_3\)): In this mode, the negative half-cycle begins, so it is similar to mode 1. Nevertheless, the load is fed with PV2 since the switches \(S_1\) and \(S_4\) are OFF while \(S_2\) and \(S_3\) are ON, as shown in Fig. 8. Therefore, \(L_m\) is again positively clamped to the output forming a P-mode. This interval ends when \(i_{nr} = i_{nm}\) again at \(\theta_3 = \pi / f_{ns} + \pi\), and the tank normalized equations can be defined as in (8).

\[
\begin{align*}
v_{nr} (\theta) &= -\frac{1}{M_2} + 1 - i_{nr} (\theta_0) \sin \left( \frac{\theta - \pi}{f_{ns}} \right) \\
&+ \left[ \frac{1}{M_2} - 1 + v_{nr} (\theta / f_{ns}) \right] \cos \left( \frac{\theta - \pi}{f_{ns}} \right) \\
i_{nr} (\theta) &= -i_{nr} (\theta_0) \cos \left( \frac{\theta - \pi}{f_{ns}} \right) \\
&- \left[ \frac{1}{M_2} - 1 - v_{nr} (\theta / f_{ns}) \right] \sin \left( \frac{\theta - \pi}{f_{ns}} \right) \\
i_{nm} (\theta) &= \frac{\theta - \theta_0}{l} - i_{nm} (\theta_0)
\end{align*}
\]

where

\[
v_{nr} \left( \frac{\pi}{f_{ns}} \right) = -v_{nr} (\theta_0) + \frac{1}{M_1} - \frac{1}{M_2}
\]

Mode 5 (\(\theta_3, \theta_4\)): This mode is the mirror of mode 2 in the negative half cycle, so the switches remain in the same state as the previous mode, shown in Fig. 9. This means O-Mode is formed again since the tank is fed only with PV2 and there is no output power flow. This mode remains till the effect of \(\psi\) appears at \(\theta_4\) and its tank components normalized equations can be defined as in (9).

\[
\begin{align*}
v_{nr} (\theta) &= -\frac{1}{M_2} - i_{nr} (\theta_1) \sqrt{1 + l} \sin \left( \frac{\theta - \theta_3}{\sqrt{1 + l}} \right) \\
&+ \left[ \frac{1}{M_2} + v_{nr} (\theta_1) \right] \cos \left( \frac{\theta - \theta_3}{\sqrt{1 + l}} \right) \\
i_{nr} (\theta) &= -i_{nr} (\theta_1) \cos \left( \frac{\theta - \theta_3}{\sqrt{1 + l}} \right) \\
&- \left[ \frac{1}{M_2} + v_{nr} (\theta_1) \right] \sin \left( \frac{\theta - \theta_3}{\sqrt{1 + l}} \right)
\end{align*}
\]
where

\[ v_{nr} (\theta_2) = -v_{nr} (\theta_1) + \frac{1}{M_1} - \frac{1}{M_2} \]

Mode 6 (\(2\pi/f_{ns}\)): This is the final mode of a whole switching cycle. As in mode 3, the \(\theta_4\) and the normalized full-cycle angle (\(2\pi/f_{ns}\)) are separated by \(\varphi_n\). In this mode, the switches \(S_1\) and \(S_3\) are OFF while \(S_2\) and \(S_4\) are ON, so the tank is freewheeling as shown in Fig. 10. This interval represents the \(S^+\)-Mode, and the tank components normalized equations can be defined as in (10).

\[ v_{nr} (\theta) = -i_{nr} (\theta_2) \sqrt{1 + l} \sin \left(\frac{\theta - \theta_4}{\sqrt{1 + l}}\right) + v_{nr} (\theta_4) \cos \left(\frac{\theta - \theta_4}{\sqrt{1 + l}}\right) + v_{nr} (\theta_1) \cos \left(\frac{\theta - \theta_4}{\sqrt{1 + l}}\right) - v_{nr} (\theta_2) \sin \left(\frac{\theta - \theta_4}{\sqrt{1 + l}}\right) \]

\[ i_{nr} (\theta) = -i_{nr} (\theta_2) \cos \left(\frac{\theta - \theta_4}{\sqrt{1 + l}}\right) - v_{nr} (\theta_4) \sin \left(\frac{\theta - \theta_4}{\sqrt{1 + l}}\right) \times \sin \left(\frac{\theta - \theta_4}{\sqrt{1 + l}}\right) \]

where

\[ v_{nr} (\theta_4) = -v_{nr} (\theta_2) + \frac{1}{M_1} - \frac{1}{M_2} \]

V. GAIN EQUATIONS AND CHARACTERISTICS

By utilizing the continuity and symmetry between \(i_{nr}\) and \(i_{nm}\) in mode 1 and 2, \(i_{nr} (\theta_0) = -i_{nm}(\theta_1)\), the initial value \(i_{nr} (\theta_0)\) can be derived as in (11).

\[ i_{nr} (\theta_0) = -\frac{\pi}{2l} \]

Conversely, from the normalized output current \((I_{no})\) equation in (12), the initial value \(v_{nr} (\theta_0)\) can be derived as in (13).

\[ I_{no} = \frac{f_{ns}}{\pi} \int_0^{\pi} [i_{nr} (\theta) - i_{nm}(\theta)]d\theta = Q_o \]

\[ v_{nr} (\theta_0) = \left(1 - \frac{1}{M_1} - \frac{1}{M_2} - \frac{\pi}{2f_{ns}} Q_o\right) \]

where \(Q_o = \pi^2 Z_o / (8n^2 R_o)\) is the quality factor and \(R_o\) is the load resistance. By using the relations in (5)-(13), the gain equations for both PVs are defined as in (14).

\[ M_1 = \left[1 + \frac{\pi}{2f_{ns}} Q_o \left(1 - \cos \left(\frac{\pi}{f_{ns}}\right)\right) - \frac{\pi}{2l} \sqrt{1 + l} \sin \left(\frac{\pi}{f_{ns}}\right) + \cos \left(\frac{\pi}{f_{ns}}\right)\right]^{-1} \times \frac{1}{l + k \cos \left(\varphi_n / \sqrt{1 + l}\right)} \]

\[ M_2 = kM_1 \]

\[ M_{avg} = \frac{2}{1/M_1 + 1/M_2} \]

where \((k = M_2 / M_1)\) is the gain ratio between the two PVs and \(M_{avg}\) is the average gain of the tank.

Accordingly, the characteristic curve of the proposed dual-input LLC converter is derived for multiple quality factor values in Fig. 11 and several inductor ratios in Fig. 12(a). The curves show that the gain exponentially increases as \(f_{ns}\) decreases until it reaches the peak at the margin between the inductive and capacitive region, and it decreases rapidly as \(f_{ns}\) enters the capacitive region. Furthermore, the curves show that the boost gain is higher for the lower \(Q_o\) and \(l\) values while the frequency range is wider for higher \(Q_o\) and \(l\) values. This proves that the derived gain equations meet the features of the LLC converter. In addition, the effect of \(\varphi\) on the gain curve is verified in Fig. 12(b). It is clear that \(\varphi\) has no significant impact on the gain curve shape, but it shifts it toward a lower gain as the \(|\varphi|\) increases. Furthermore, the phase shift effect increases as the frequency move farther below the resonant frequency in the inductive region. As a result, (14) can be considered as the general gain equation for all regions of operation.

VI. WAVEFORMS BOUNDARIES AND ZVS ANALYSIS

A. WAVEFORM BOUNDARIES

As illustrated in Section III, the steady-state waveform changes depending on \(f_{ns}\), \(\varphi\), and the load condition. In this section, the boundaries between those waveforms are defined.

For PONs/PONs boundary, PONs occurs when \(v_{nm,\pi}\) is negatively clamped to the output before the negative half cycle begins. This means POS limit can be defined as in (15).

\[ v_{nm} S \left(\frac{\pi}{f_{ns}}\right) = -1 \]

Since the quality factor is inversely proportional to the load resistance, it is used to represent the load condition for the waveforms' boundaries. As a result, PONs starts at

\[ Q_{PONs} = \frac{2f_{ns}}{\pi} + \frac{l + 1}{l} - \frac{\cos \left(\frac{\pi}{f_{ns}}\right)}{M_2} - \frac{\pi}{2l} \sqrt{1 + l} \sin \left(\frac{\pi}{f_{ns}}\right) \]

\[ \frac{\pi}{2l} \cos \left(\frac{\pi}{f_{ns}}\right) \]

(16)
A. Alhatlani et al.: Comprehensive Analysis and Gain Derivation of Phase-Shifted Dual-Input LLC Converter

**FIGURE 11.** Dual-input LLC converter characteristic curve for multiple $Q_0$ with $I = 4$, $k = 1.2$ and $\phi = -30^0$: (a) $M_1$ vs $M_2$ vs $f_{ns}$, and (b) $M_{avg}$ vs $f_{ns}$.

OPOS appears when the load is too light to enable $v_{num}$ of POS to reach 1 at the beginning of the P-mode. Accordingly, the POS/OPOS boundary can be defined from (2) as in (17).

\[ v_{num}(\theta_0) = 1 \]  

This means OPOS starts at

\[ Q_{OPOS} = \frac{2f_{ns}}{\pi I} \]  

PN waveform is independent of $\phi$ since it appears under too heavy-load condition. Therefore, its limit can be defined from the PN/PO boundary when $v_{num}$ of PO is negatively clamped before the negative half cycle begins. This means PO limit is

\[ v_{num} \left( \frac{\pi}{f_{ns}} \right) = -1 \]  

Accordingly, PN starts at

\[ Q_{PN} = \frac{2f_{ns}}{\pi} + \frac{l+1}{l} - \frac{\pi}{2} \sqrt{1+l} \sin \left( \frac{\pi}{2} \sqrt{1+l} \right) \]  

Finally, the gain boundaries are derived in Fig. 13, which shows that POS waveform covers most of the inductive region of the converter. Therefore, it is considered as the general waveform of the proposed dual-input LLC converter.

**FIGURE 12.** Dual-input input LLC converter characteristic curve with $Q_0 = 0.4$ and $k = 1.2$: (a) different $I$ values at $\phi = -30^0$, and (b) various $\phi$ at $I = 4$.

**FIGURE 13.** Dual-input LLC converter waveform boundaries through the gain curve with $I = 4.8$, $\phi = -30^0$ and $k = 1.3$.

**B. ZVS ANALYSIS**

For any LLC converter, two conditions should be satisfied to achieve ZVS. First, the converter operates in the inductive region, on the right side of the peak gain. Second, the dead
the inductive and capacitive regions.

\[ i_{nr}(\theta_2) \geq 0 \]  
\[ i_{nr}(\theta_1) \leq 0 \]  

In contrast, when \( \phi \) is applied, \( S_4 \) and \( S_3 \) turn ON after \( S_1 \) and \( S_2 \), respectively, when \( -\phi \) is applied. Accordingly, \( S_1 \) and \( S_2 \) have the same ZVS limits as in (24) and (23), respectively, whereas \( S_4 \) and \( S_3 \) achieve ZVS when they satisfy (21) and (22), respectively. This indicates the ZVS region for \( S_3 \) and \( S_4 \) is smaller than \( S_1 \) and \( S_2 \) in case of positive phase shift.

As shown in Fig. 11-12, the peak gain gets closer to the resonance frequency as the load weight increases, or the inductor ratio decreases. However, the peak gain never occurs above resonance even when phase shift modulation is applied [13]. Therefore, ZVS is always achieved for \( f_{as} > 1 \), yet hard switching might occur for \( f_{as} < 1 \) [10]. Since POS waveform always starts with \( i_{ns}(\theta_0) = i_{nr}(\theta_0) \), ZVS is always achieved when it appears. Consequently, the POS analysis used in Section IV cannot be utilized to derive the ZVS limits. Therefore, the mode-by-mode normalized analysis for PONs is used instead, with taking (21)-(24) into consideration. Nevertheless, PON waveform does not have a closed-form solution, so the ZVS limits are derived by applying the numerical solutions using MATLAB.

The PSM impact on the ZVS limits for the proposed converter under positive \( \phi \) (\( V_{PV1} > V_{PV2} \)) is shown in Fig. 14. Since controlling the PV panels does not require high \( |\phi| \), the effect of \( \phi \) from 0 to \(-\pi/4\) on the ZVS range is tested. As can be seen in Fig. 14(a), the ZVS range, the area below the shape, decreases for \( S_1/S_2 \) as \( |\phi| \) increases or \( f_{as} \) decreases. On the other hand, the ZVS range increases with \( |\phi| \) for \( S_3/S_4 \) as shown in Fig. 14(b). Similarly, \( S_1/S_2 \) ZVS range increases with \( |\phi| \) in case of a positive \( \phi \), since it becomes the leading leg, while \( S_3/S_4 \) ZVS range decreases. It is noteworthy that since the PSM is used to control the PV panels, there is no need to test the impact of a positive \( \phi \) when \( V_{PV1} > V_{PV2} \) because it would harm the delivered power from the panels, and vice versa in case of \( V_{PV1} < V_{PV2} \).

VII. MPPT CONTROL ALGORITHM

Since two PV panels are connected to a single LLC tank in the proposed topology, both control variables are used to track the MPP for both panels. Therefore, an improved Perturbation and Observation (P&O) method is presented in this paper. According to the panels status, either \( \phi \) or \( f_s \) is modified in every iteration \( x \) till MPPT is achieved as can be seen in Fig. 15.

When the panel operates on the right of MPP in the PV power curve, the slope is negative. In contrast, the slope is positive on the left side of MPP. In the proposed topology, there are four different cases demand perturbation as shown in the flowchart in Fig. 15. In the first case, a \( -\phi \) is required when \( PV_1 \) and \( PV_2 \) operate on the right and left side of their respective MPP, respectively. Therefore, a negative phase shift step \( -\Delta \phi \) is applied in this case. The second scenario
A. Alhatlani et al.: Comprehensive Analysis and Gain Derivation of Phase-Shifted Dual-Input LLC Converter

is the opposite scenario where a positive phase shift $+\Delta \phi$ is applied. Thirdly, a negative switching frequency step $-\Delta f_s$ is applied when both panels operate on the right side of their respective MPPs. The last scenario is the contrast of the third one where a positive switching frequency step $+\Delta f_s$ is applied. This means, $\phi$ is modified when the PVs operate in opposite sides while $f_s$ is tuned when the panels operate on the same side of MPP. Finally, there is no change applied when the slope is zero since MPPT is achieved in that case. Following this algorithm, MPPT is achieved for both panels of the proposed converter.

VIII. EXPERIMENTAL RESULTS

A 450-W, 300V output voltage prototype of the proposed dual-input LLC converter was built and tested to verify the theoretical analysis as shown in Fig. 16. Two (Agilent-E4350B) solar array simulators were used to represent the two PV panels. To set the resonant frequency on $f_r = 100 \text{ kHz}$ and the inductor ratio on $l = 4.1$, the resonant tank components were designed as follow: $L_r = 3 \mu\text{H}$, $C_r = 0.85 \mu\text{F}$ and $L_m = 12.4 \mu\text{H}$. Since the nominal input for each panel is 52V, the transformer turns ratio was designed as $n = 10/58$ to reach a 300V output when the converter operates at resonance.

The experimental waveforms of the resonant current ($i_r$) and magnetizing voltage ($v_m$) for different conditions are shown in Fig. 17-19. As illustrated in Section IV, the switches $S_1$ and $S_2$ are complementary as $S_3$ and $S_4$, so $S_1$ and $S_3$ waveforms are used to determine the phase shift ($\phi$). First,
is too light to be fully clamped to the load at the beginning of the half cycle.

Finally, the gain characteristics curve is drawn using the experimental results to verify the derived gain equations in (14). Fig. 20 shows the experimental and theoretical gain curves for three different power rates including 150-W, 300-W, and 450-W. As can be seen in the figure, the experimental results almost follow the gain equations for a wide range below and above resonance with a less than 10% error. This means the derived gain equations give an efficient indication of the tank behavior for the proposed dual-input LLC converter.

IX. CONCLUSION

This paper presents a comprehensive steady-state analysis of the dual-input LLC converter interfacing two PV panels for grid-tied applications. PSM and FSM control technique are utilized to achieve MPPT for both panels. The contribution of this paper is studying the impact of phase shift on the dual-input LLC converter with variable frequency. First, all the modes of operation are illustrated in details. Secondly, the conditions that form the steady state switching cycle are demonstrated in terms of the switching frequency, phase shift, and load weight. Another contribution is deriving the DC gain equations and their characteristic curves over $f_{ns}$ by analyzing the steady-state equations in the normalized form of the proposed converter. Furthermore, a mode by mode analysis for a whole switching cycle of the resonant tank considering the phase shift is illustrated. In addition, the effect of different parameters including the inductor ratio, quality factor, and phase shift on the gain curves is studied. After that, the boundaries between the different waveforms are determined along with the ZVS analysis. Then, an extended P&O algorithm that tracks the MPP for each panel is presented and illustrated.

Finally, a 450-W prototype of the proposed dual-input LLC converter is built to verify the theoretical analysis. The experimental waveforms are similar to the provided analysis and the experimental gain curves follow the derived gain equations. Accordingly, the derived gain equations can be utilized to indicate the dual-input LLC converter gain behavior. These gain characteristics along with the parameter dependency will be very helpful to the future researchers to choose design parameters.

ACKNOWLEDGMENT

The authors would like to thank Dr. Nasser Kutkut, the CEO/Founder of Smart Charging Technologies, for his contribution in this research.
REFERENCES

[1] P. Zhang, Y. Wang, W. Xiao, and W. Li, “Reliability evaluation of grid-connected photovoltaic power systems,” IEEE Trans. Sustain. Energy, vol. 3, no. 3, pp. 379–389, Jul. 2012, doi: 10.1109/TSTE.2012.2186644.

[2] REN21. (2021). Renewables 2020 Global Status Report. [Online]. Available: https://www.ren21.net/gsr-2020/

[3] S. Harb, M. Kedia, H. Zhang, and R. S. Balog, “Microinverter and string inverter grid-connected photovoltaic system—A comprehensive study,” in Proc. Conf. Rec. IEEE Photovolt. Spec. Conf., Jun. 2013, pp. 2885–2890, doi: 10.1109/SPV.2013.6745072.

[4] H. Al-Atrash, F. Tian, and I. Batarseh, “Tri-modal half-bridge converter topology for three-port interface,” IEEE Trans. Power Electron., vol. 22, no. 1, pp. 341–345, Jan. 2007.

[5] Z. Qian, O. Abdel-Rahman, H. Al-Atrash, and I. Batarseh, “Modeling and control of three-port DC/DC converter interface for satellite applications,” IEEE Trans. Power Electron., vol. 25, no. 3, pp. 637–649, Mar. 2010.

[6] B. N. Alajmi, M. I. Marei, and I. Abdelsalam, “A multiport DC–DC converter based on two-quadrant inverter topology for PV systems,” IEEE Trans. Power Electron., vol. 36, no. 1, pp. 522–532, Jan. 2021, doi: 10.1109/TPEL.2020.3002504.

[7] W. Sun, Y. Xing, H. Wu, and J. Ding, “Modified high-efficiency LLC converters with two split resonant branches for wide input-voltage range applications,” IEEE Trans. Power Electron., vol. 33, no. 9, pp. 7867–7879, Sep. 2018, doi: 10.1109/TPEL.2017.2773484.

[8] H. Hu, X. Fang, F. Chen, Z. J. Shen, and I. Batarseh, “A modified high-efficiency LLC converter with two transformers for wide input-voltage range applications,” IEEE Trans. Power Electron., vol. 28, no. 4, pp. 1946–1960, Apr. 2013, doi: 10.1109/TPEL.2012.2210959.

[9] C. Bhuvaneswari and R. S. R. Babu, “A review on LLC resonant converter,” in Proc. Int. Conf. Comput. Power Energy Inf. Commun. (ICCPEC), Apr. 2016, pp. 620–625, doi: 10.1109/ICCPEC.2016.7557268.

[10] H. Wen, J. Gong, C.-S. Yeh, Y. Han, and J. Lai, “An investigation on fully zero-voltage-switching condition for high-frequency GaN based LLC converter in solid-state transformer application,” in Proc. IEEE Appl. Power Electron. Conf. Expo. (APEC), Mar. 2019, pp. 797–801, doi: 10.1109/APEC.2019.8721789.

[11] X. Fang, H. Hu, Z. J. Shen, and I. Batarseh, “Operation mode analysis and peak gain approximation of the LLC resonant converter,” IEEE Trans. Power Electron., vol. 27, no. 4, pp. 1985–1995, Apr. 2012, doi: 10.1109/TPEL.2011.2168545.

[12] M. Xingkui, H. Qisheng, K. Qingbo, X. Yudi, Z. Zhe, and M. A. E. Andersen, “Grid-connected photovoltaic micro-inverter with a new hybrid control LLC resonant converter,” in Proc. 42nd Annu. Conf. IEEE Ind. Electron. Soc., Oct. 2016, pp. 2319–2324, doi: 10.1109/IECON.2016.7793632.

[13] B. Guo, Y. Zhang, J. Zhang, X. Wang, and J. Ding, “Hybrid control strategy of phase-shifted full-bridge LLC converter based on digital direct phase-shift control,” China Electrotech. Soc., vol. 33, no. 19, pp. 4583–4593, 2018, doi: 10.15955/j.cnki.1008-6753.aces.171141.

[14] M. Rashidi, “Design and implementation of a LLC resonant solid-state transformer,” IEEE Trans. Ind. Appl., vol. 56, no. 4, pp. 3855–3864, Mar. 2020, doi: 10.1109/TIA.2020.2982847.

[15] S. M. Tayebi, H. Hu, S. Abdel-Rahman, and I. Batarseh, “Dual-input single-resonant tank LLC converter with phase shift control for PV applications,” IEEE Trans. Ind. Appl., vol. 55, no. 2, pp. 1729–1739, Mar. 2019, doi: 10.1109/TIA.2018.2883015.

[16] C. Zhao, S. D. Round, and J. W. Kolar, “An isolated three-port bidirectional DC–DC converter with decoupled power flow management,” IEEE Trans. Power Electron., vol. 23, no. 5, pp. 2443–2453, Sep. 2008, doi: 10.1109/TPEL.2008.2002056.

[17] A. Alhatlani, S. Ghosh, I. Batarseh, and N. Kutkut, “Exact steady-state analysis of phase-shifted dual-input LLC converter,” in Proc. IEEE Energy Convers. Congr. Expo. (ECCE), Sep. 2019, pp. 1394–1400.

[18] Q. Zhang, C. Hu, L. Chen, A. AmiriAhmadi, N. Kutkut, Z. J. Shen, and I. Batarseh, “A center point iteration MPPT method with application on the frequency-modulated LLC microinverter,” IEEE Trans. Power Electron., vol. 29, no. 3, pp. 1262–1274, Mar. 2014, doi: 10.1109/TPEL.2013.2262806.

[19] X. Chen and I. Batarseh, “A fixed switching frequency dual-input LLC converter with PWM controlled semi-active rectifiers for PV applications,” in Proc. IEEE Appl. Power Electron. Conf. Expo. (APEC), Jun. 2021, pp. 320–326, doi: 10.1109/APEC42165.2021.9487039.

[20] N. Alamir, M. A. Ismail, and M. Orabi, “New MPPT technique using phase-shift modulation for LLC resonant micro-inverter,” in Proc. 19th Int. Middle East Power Syst. Conf. (MEPCON), Dec. 2017, pp. 1465–1470, doi: 10.1109/MEPCON.2017.8301376.

[21] S. Ghosh, M. Safayetullah, M. T. Elraii, and I. Batarseh, “A novel four-port LLC converter for dual PV and battery integration,” in Proc. 47th Annu. Conf. IEEE Ind. Electron. Soc., Oct. 2021, pp. 1–6, doi: 10.1109/IECON48115.2021.9589562.

[22] S. Ghosh, R. Rezaii, A. Alhatlani, and I. Batarseh, “Analysis and control of grid-tied quad-PV LLC converter with MPPT,” in Proc. IEEE Energy Convers. Congr. Expo. (ECCE), Oct. 2020, pp. 1912–1918, doi: 10.1109/ECCE44975.2020.9235601.

ABDULLAH ALHATLANI (Member, IEEE) received the B.S. degree in electrical engineering from King Saud University, Riyadh, Saudi Arabia, in 2011, and the M.S. and Ph.D. degrees in electrical engineering from the University of Central Florida (UCF), Orlando, FL, USA, in 2017 and 2021, respectively. He is currently an Assistant Professor in electrical engineering with the Department of Electrical Engineering, Imam Mohammad Ibn Saud Islamic University (IMSIU), Riyadh. His research interests include design and control multi-port converters, DC/DC power conversion for solar systems, and PV-Battery systems.

SUMANA GHOSH (Graduate Student Member, IEEE) received the bachelor’s degree from the West Bengal University of Technology, West Bengal, India, in 2014, and the master’s degree in electrical engineering from the National Institute of Technology, Durgapur, India, in 2018. She is currently pursuing the joint Ph.D. degree with the Department of Electrical Engineering and the Florida Power Electronics Centre, University of Central Florida.

Her research interests include the DC/DC resonant converters, advanced control method, multi-input DC/DC converter, and PFC for data server application.

ISSA BATARSEH (Fellow, IEEE) received the B.S., M.S., and Ph.D. degrees in electrical engineering from the University of Illinois, Chicago, IL, USA, in 1983, 1985, and 1990, respectively. He is currently a Professor in electrical engineering with the Department of Electrical and Computer Engineering, University of Central Florida (UCF), Orlando, FL, USA. From 1989 to 1990, he was a Visiting Assistant Professor with Purdue University, Calumet, IN, USA, before joining UCF, in 1991. His research interests include power electronics and energy conversion systems for smart-grid and renewable energy applications. His team at the Florida Power Electronics Center (FPEC) has been leading the design, development, commercialization of smart microinverters, and smart EV and industrial chargers. He has published more than 90 refereed journals and 320 conference papers in addition to 37 U.S. patents. Based on his research work, he has founded two start-up companies in power electronics: petra systems and ApECOR. He authored a textbook titled Power Electronic Circuits. He is a Fellow Member of the AAAS and a member of the National Academy of Inventors (NAI) and has been inducted into the Florida Inventors Hall of Fame. He has served as the Chair of IEEE Power Electronics Specialists Conference (PESC 2007) and was the Chair of the IEEE Power Engineering Chapter, the IEEE Orlando Section, and served on PELS AdCom Committee. He is a Registered Professional Engineer (PE) in FL.

VOLUME 10, 2022 116807