Phase-Shifting Adaptive LLC Resonant Converter with Reduced Turn-Off Loss in Wide Voltage Application

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Abstract: In this paper, a hybrid control method with adaptive phase-shifting modulation (PSM) and pulse-frequency modulation (PFM) is proposed to optimize the steady-state performance of an LLC resonant converter in wide voltage-gain range application. For the primary-side switches under the hybrid control method, zero-voltage switching (ZVS) performance is maintained over wide voltage-gain range and the turn-off loss is reduced; therefore, the converter reconciles wide voltage-gain range and high efficiency. Mode characteristics under different phase-shifting-angle-θ and switching-frequency-fs combinations is analyzed, and then the soft-switching characteristic is revealed. By introducing time domain analysis, the turn-on current of the primary-side switches is calculated, and thus the ZVS boundary of different θ and fs combinations is inferred. In addition, to acquire the optimum steady-state operation of the converter, the turn-off current is calculated; by making the converter work near to the minimum turn-off current operating point, the turn-off loss can be reduced greatly. With the principles of achieving ZVS performance and minimizing turn-off current, the phase-shifting angle θ of PSM is designed to be adaptive to the reference output voltage, no additional circuits are needed, and the two control degrees (θ and fs) are simplified to one (fs). The simulation and experiment are developed to verify the feasibility and effectiveness of the hybrid control method; the results show that ZVS performance is maintained at wide voltage-gain range and the turn-off current of the hybrid control method is reduced to that of the single PFM, and thus the turn-off loss is reduced. The efficiency comparison validates the fact that the hybrid control method has less power loss than single PSM and single PFM.

Keywords: LLC resonant converter; wide voltage gain; hybrid control; soft-switching performance; turn-off loss

1. Introduction

The LLC resonant converter is one of the most promising converters in isolated DC-DC applications, such as flat-panel TVs, laptop adapters, servers and so on, because of its attractive features: smooth waveforms, soft-switching characteristic for both primary-side MOSFET and secondary-side diodes [1–6], high efficiency at wide load range and high power density. The resonant network of the LLC resonant converter is composed of three passive devices, so there are two resonant frequency points. Other resonant converters, such as LCC [5–7] and LCLC [8–10] are also used in various industry applications for these similar features.

For LLC resonant converters, one of the most prevalent modulation strategies is pulse-frequency modulation (PFM); by regulating the switching frequency, the impedance of the resonant rank can be adjusted, and as a result, the output voltage is regulated. However, when it comes to the application where the input voltage or output voltage is wide such as electric car charging pile and solar energy applications, the switching frequency will swing within a wide range to meet the voltage-gain requirement, which brings challenges to the magnetic components, gate drivers and EMI performance. In addition, wide voltage gain
usually means small inductance ratio, which result in a large reactive power circulation. Therefore, it is very difficult to optimally design the LLC converter adapted to wide voltage gain with traditional topology and the single-PFM control method. To widen the voltage gain for the LLC resonant converter, many articles in the literature develop improvements from the topology reconfiguration and multi-degree modulation strategies.

The reconfiguration of the resonant tank has attracted much attention for widening the voltage-gain range, including the resonant inductor $L_r$, resonant capacitor $C_r$, magnetic inductor $L_m$, and transformer turns ratios $n$. The modification focuses on variable $L_m$, including adding auxiliary LC structures [11,12] and auxiliary transformer winds [13]. Moreover, the resonant capacitor $C_r$ can be replaced by a switch-controlled capacitor (SCC); the equivalent resonant capacitance is adjusted by high-frequency switching to meet different voltage-gain demand [14,15]. These methods all need additional circuits, which increases the cost the converter and reduces the power density of the converter. Other topology reconfiguration methods are also proposed: paper [16] transits the half-bridge and full-bridge mode in the primary side of the full-bridge LLC converter, and wide voltage gain can be achieved. Paper [17] proposes an interleaved secondary-side modulated LLC resonant converter to achieve wide output range. Paper [18] forms a different inverter-bridge type with six MOSFETs, but only steady states are considered and the cost of the inverter bridge is high. These methods all suffer from the smooth-switching problem, dynamically. Paper [19] proposes a novel topology with an additional PWM switch at the secondary side for narrowing down the frequency range. Paper [20] uses a magnetic amplifier auxiliary post regulator to extend voltage gain, while switching frequency $f_s$ keeps it at a constant value, but the control is complex and the cost is high. A multi-mode hybrid LLC resonant converter with wide output voltage range is proposed in [21]. In the absence of an auxiliary switch, the converter can operate in three different modes by controlling the drive signal of the main switch. A novel LLC converter with single input and double output for wide output voltage applications is proposed in [22]. The H5 bridge is used on the main side and is connected to two separate resonant networks.

The topology reconfiguration methods all require an extra passive or active device, and the cost and control complexity may reduce the benefits of the LLC resonant converter. As a result, the hybrid modulation method as a supplement of PFM is being widely researched. Phase-shifting modulation (PSM) is a promising supplement to the PFM in the LLC resonant converter, and the steady-state analysis of PSM in the LLC converter is studied in [23]. In [24,25], by controlling the phase-shifting angle, the equivalent duty cycle can be adjusted so as to regulate the output, but when the phase-shifting angle is large it may suffer from hard switching to the lagging bridge. Paper [26] applies an FM/PSM mode switching to support a wider voltage gain, but the ZVS constraint is not quantified. Paper [27] introduces a load-adaptive phase-shifting control for a light load of the LLC converter, but the design method only aims at improving the efficiency at light load, and the wide voltage gain and heavy load efficiency is not optimized. Paper [28] proposes a frequency adaptive phase-shifting modulation control for wide input voltage applications; the phase-shifting control scheme is applied in the secondary side, and thus synchronous rectification and its control circuit is needed, which is not cost-effective. Pulse width modulation (PWM) for the LLC converter is also proposed in [29,30] for the primary side and secondary side to regulate the output voltage, but the extra circuit will result in high cost. In [31,32], the burst mode (BM) strategy is employed in light and no-load conditions to achieve a wide output range, but a high-frequency oscillation exists during the OFF state and will result in EMI problems. In [33], a simplified optimal trajectory control based on the adaptive burping mode for the high-frequency LLC converter is proposed and implemented by a low-cost single-chip microcomputer. A new hybrid control method is proposed in [34]. The strategy combines burp mode control and phase shift control of the LLC converter. Burp mode control can adjust the output voltage from zero to the maximum, and phase shift control can reduce the increased resonant current during burping. In order to solve the problem of LLC light-load control, a new LLC resonant converter modulation method is proposed in
Reference [35], which simplifies the control design and maintains high efficiency over the entire load range.

To widen the voltage-gain range and keep the low cost and high efficiency of the LLC converter, a hybrid control method with adaptive PSM and PFM is proposed to optimize the steady-state performance of the LLC resonant converter in a wide voltage-gain range application. For the primary-side switches under the hybrid control, zero-voltage switching (ZVS) performance is maintained over wide voltage range and the turn-off loss is reduced; therefore, the converter reconciles wide voltage-gain range and high efficiency. Mode characteristics under different phase-shifting-angle-$\theta$ and switching-frequency-$f_s$ combinations are analyzed, and then the soft-switching characteristic is revealed. By introducing time domain analysis, the turn-on current of the primary-side switches is calculated, and thus the ZVS boundary of different $\theta$ and $f_s$ combinations is inferred. In addition, to acquire the optimum steady-state operation of the converter, the turn-off current is calculated; by making the converter work near to the minimum turn-off current operating point, the turn-off loss can be greatly reduced. With the principles of achieving ZVS performance and minimizing turn-off current, the phase-shifting angle $\theta$ of PSM is designed to be adaptive to the reference output voltage; no additional circuits are needed and the two control degrees ($\theta$ and $f_s$) are simplified to one ($f_s$). The simulation and a 1 kW/output voltage 200–500 V experiment prototype are built to validate the feasibility and effectiveness of the proposed hybrid control method. Both results show the hybrid control method can maintain ZVS performance at wide output voltage range and wide load range when compared to single PSM; the sum of the turn-off currents of the primary-side leading- and lagging-bridge switches is reduced when compared to the single PFM, and thus turn-off loss is reduced. The efficiency comparison validates the fact that the hybrid control method has less power loss than single PSM and single PFM.

This paper is organized as follows: Section 2 introduces the working principles when applying hybrid PSM and PFM, and then the soft-switching performance characteristic of different working modes is clarified. Section 3 designs an adaptive PSM control method with PFM to regulate the output voltage based on time domain analysis. Sections 4 and 5 provide PSIM simulation and experiment results of the proposed control methods, single PSM and single PFM. Finally, this paper is concluded in Section 6.

2. Working Principles of LLC Converter with Phase-Shifting and Frequency Modulation

In this section, the working principles of the LLC resonant converter when applying hybrid PSM and PFM are discussed, and the soft-switching performance characteristic is inferred. Part A introduces several definitions of the terms relating to the LLC resonant converter and its fundamental harmonic approximation (FHA) analysis results. Part B shows four main operation modes and their soft-switching performance characteristic.

2.1. Definition and FHA Analysis of LLC Resonant Converter

The LLC resonant converter is shown in Figure 1. It is mainly composed of a DC input $V_{in}$, an input filtering capacitor $C_{in}$, a full-bridge inverter, a resonant tank, a high-frequency transformer $T_1$, a full-bridge rectifier, a large output-filter capacitor $C_{out}$ and load $R_o$. The full-bridge inverter consists of four power switches $Q_1$–$Q_4$, and $D_1$–$D_4$ and $C_1$–$C_4$ are the parasitic diode and capacitor of the power switches, separately. The resonant tank consists of resonant inductor $L_r$, resonant capacitor $C_r$ and magnetic inductor $L_m$. $V_{AB}$ is the output voltage of the inverter bridge, $i_{Lr}$ is the resonant current, $v_{Cr}$ is the voltage on $C_r$, $i_m$ is the magnetizing current of $L_m$, $i_{rect}$ is the rectified current, which is composed of the difference between $i_{Lr}$ and $i_m$, $i_o$ is the load current, $V_o$ is the DC output voltage, and $f_r$ is the resonant frequency which is defined in (1).

$$f_r = \frac{1}{2\pi \sqrt{L_r C_r}}$$  (1)
Q₁–Q₄ are all turned on with a 50% duty cycle, and Q₁ is complementary to Q₂, while Q₃ is complementary to Q₄. By adjusting the switching frequency, the impedance of the resonant tank is changed, and the output is regulated by PSM. By shifting the switching angle θ between Q₁ and Q₄, V_{AB} is adjusted to a three-level square wave with a different proportion of zero-voltage level, and then output is regulated by PSM. When θ = 0°, V_{AB} is a two-level square wave without a zero-voltage level; when θ = 180°, V_{AB} equals 0, and the output voltage can be regulated to 0, theoretically.

Figure 2 shows the typical gain characteristic of the LLC resonant converter based on FHA when applying PFM. The working zone can be simply divided into three types: #1 is the boost zone, where the switching frequency fₛ is less than the resonant frequency fᵣ, and the voltage gain is larger than 1. The converter can realize primary-side ZVS and secondary-side ZCS when working at #1. However, reactive power circulation would reduce the benefit when the voltage-gain range is wide at this zone. The buck zone is #2: when the switching frequency fₛ is larger than the resonant frequency fᵣ, and the voltage gain is less than 1, the converter realizes primary-side ZVS but loses secondary-side ZCS when working in this area. The capacitive zone is #3: the whole resonant tank has a capacitive load, ZVS performance is lost, and it is necessary to avoid working in this area for MOSFET.

![LCC resonant converter topology](image)

**Figure 1.** LCC resonant converter topology.

Both #1 and #2 have advantages for realizing ZVS performance: #1 can fully meet the demand for narrow voltage-gain range application, but when it comes to wide gain range, only #1 participation will result in a large reactive power circulation, which undermines the benefits of the converter. To solve this, #2 also participates in the work to regulate the gain, but the frequency range will be large when the voltage range and load range are wide, which is not good for the design of magnetic components and the gate driver. PSM in #2 can change the equivalent input of the resonant tank to reduce voltage gain with a narrow frequency range, but ZVS performance will be hard to achieve if the phase-shifting angle θ is large. To solve these problems, in this paper, an adaptive phase-shifting angle is proposed to ensure the ZVS performance in #2 and reduce the sum of the turn-off currents; the design principle will be discussed in Section 3.

![Gain characteristic based on FHA under different δs](image)

**Figure 2.** Gain characteristic based on FHA under different δs.
2.2. Key Working Waveforms When Applying PSM

In this part, the working zone #2 is mainly been discussed where the working frequency \( f_s \) is larger than the resonant frequency \( f_r \). The equivalent circuit of the converter is shown in Figure 3: \( a \), \( S_0 \) and \( b \) are state parameters which consist of different modes, and Table 1 shows the A–I stage with different state combinations.

![Equivalent circuit with PSM](image)

**Figure 3.** Equivalent circuit with PSM.

<table>
<thead>
<tr>
<th>( a )</th>
<th>(+1)</th>
<th>0</th>
<th>(-1)</th>
</tr>
</thead>
<tbody>
<tr>
<td>( S_0 )</td>
<td>on</td>
<td>off</td>
<td>on</td>
</tr>
<tr>
<td>( b )</td>
<td>(-1)</td>
<td>(+1)</td>
<td>(+1)</td>
</tr>
<tr>
<td>stage</td>
<td>A</td>
<td>B</td>
<td>C</td>
</tr>
</tbody>
</table>

Table 1. Different stage under different state combinations.

One complete switching period of the LLC converter is composed of stage A–I, as paper [21] reveals: the stage combinations ACD, CDE, CDF and CDEF are the most common modes, while other modes only occur at very light load, which is not discussed in this paper. The four types of working modes correspond to different phase-shifting-angle-\( \theta \) and switching-frequency-\( f_s \) combinations at #2 working zone. Type 1 represents the situation when \( f_s \) is large and \( \theta \) is small. In this type, the converter performs like a single PFM; the primary-side ZVS can be guaranteed, and the secondary-side ZCS is lost. This type can be divided into five working modes at a positive half cycle, as Figure 4 shows. When including the dead time, the negative half cycle is similar to the positive, which will not be explained repeatedly.

![Waveforms](image)

**Figure 4.** Key waveforms of type 1 ACD.

Mode 1 [\( t_1 \sim t_2 \)]: In this mode, the inverter-bridge output voltage \( V_{AB} \) equals \( V_m \), \( L_m \) is clamped by \(-nV_o\). \( i_m \) decreases linearly, \( L_r \) resonates with \( C_r \). \( L_r \) rises rapidly, and the secondary-side rectifier diodes \( D_{r2} \) and \( D_{r3} \) conduct to transfer energy to the load side.
Mode 2 [$t_2$–$t_3$]: At $t_2$, $i_m$ drops to the same value as $i_{Lr}$, while $V_{AB}$ still equals $V_{in}$, but the resonant process will result in a difference in current between $i_{Lr}$ and $i_m$. Then $D_{t1}$ and $D_{t4}$ are turned on and $D_{t2}$ and $D_{t3}$ are turned off. Due to the continuous conduction mode of the diodes, $D_{t2}$ and $D_{t3}$ lose ZCS performance. $L_m$ begins to be clamped by $nV_o$ and rises linearly, and $L_r$ continues to resonate with $C_r$.

Mode 3 [$t_3$–$t_4$]: This is the dead-time mode. After $Q_1$ turns off, $i_{Lr}$ is positive, then $C_2$ is discharged and $C_1$ is charged until the $D_2$ freewheels. Thus, this mode provides the conditions for the ZVS performance of $Q_2$.

Mode 4 [$t_4$–$t_5$]: After $Q_2$ turns on, $V_{AB}$ equals 0, $L_r$ and $C_r$ are still resonating, $L_m$ is clamped by $nV_o$, $i_m$ rises linearly, and $D_{t1}$ and $D_{t4}$ conduct to transfer energy to the load side.

Mode 5 [$t_5$–$t_6$]: This is the dead-time mode. After $Q_4$ turns off, $i_{Lr}$ is positive, then $C_3$ is discharged and $C_4$ is charged until the $D_2$ freewheels. Thus, this mode provides the conditions for the ZVS of $Q_3$. Compared to mode 3, at this mode, due to $i_{Lr}$ decreasing rapidly at mode 4, the turn-off current of $Q_4$ is reduced, and thus the turn-off loss is reduced. Owing to the reduction in $i_{Lr}$, the ZVS implementation for $Q_3$ will be harsher, so a minimum value of $i_{Lr}$ needs to be maintained for a reliable ZVS performance of $Q_3$.

Type 2 represents the situation when $f_s$ and $\theta$ are moderate, and the load is light. In this type, $L_m$ will join to the resonance between $L_r$ and $C_r$ at 0 voltage level of $V_{AB}$, so the secondary-side ZCS is obtained. This type can be divided into five working modes at the positive half cycle, as Figure 5 shows, when including the dead time. Due to the limited space of the article, only Mode 4 [$t_4$–$t_5$] is discussed. Other modes are similar to the working condition of the ACD type.

![Figure 5. Key waveforms of type CDE.](image)

Mode 4 [$t_4$–$t_5$]: At $t_4$, $i_{Lr}$ drops to the same value as $i_m$, then $L_m$ begins to join to the resonance between $L_r$ and $C_r$, $i_m$ keeps consistent with $i_{Lr}$, and thus no current transits to the secondary side, and $D_{t1}$ and $D_{t4}$ achieve ZCS performance. Due to the $i_{Lr}$ drops in mode 3 [$t_4$–$t_5$], the turn-off loss of $Q_4$ is reduced and the ZVS implementation for $Q_3$ will be harsher. A minimum value of $i_{Lr}$ needs to be maintained for a reliable ZVS performance of $Q_3$.

Type 3 represents the situation when $f_s$ is low but $\theta$ is large and the load is heavy. Due to this reason, the ZVS performance of the lagging bridge is easily lost, and the hard switching performance will impact the reliability and efficiency seriously. This type can also be divided into five working modes at a positive half cycle as Figure 6 shows, when
including the dead time. Due to the limited space of the article, only mode 4 \([t_4\cdots t_5]\) and mode 5 \([t_5\cdots t_6]\) are discussed; other modes are similar to the CDE type.

![Figure 6. Key waveforms of type CDF.](image)

Mode 4 \([t_4\cdots t_5]\): Due to \(C_r\) being charged by a large current \(i_{mL}\) in \([t_1\cdots t_4]\), when \(i_{mL}\) drops to the same value as \(i_m\) at \(t_4\), \(L_m\) will not join in with the resonance with \(L_r\) and \(C_r\), and the different current between \(i_{mL}\) and \(i_m\) will make \(D_1\) and \(D_4\) turn off and \(D_2\) and \(D_3\) turn on. This means \(D_1\) and \(D_4\) lose ZCS performance. In this mode, \(i_{Lr}\) keeps decreasing and then changes to the negative direction, which will make \(Q_3\) lose ZVS performance.

Mode 5 \([t_5\cdots t_6]\): This is the dead-time mode. After \(Q_4\) turns off, owing to the negative \(i_{Lr}, D_4\) freewheels, so the energy stored in \(C_3\) will not be released, and the ZVS performance of \(Q_3\) is lost.

Type 4 represents the situation which falls between type 3 and type 4. It can be divided into six working modes at a positive half cycle, as Figure 7 shows, when including the dead time. Mode 4 \([t_4\cdots t_5]\) is equal to mode 4 in type 2, and mode 5 \([t_5\cdots t_6]\) is equal to mode 5 in type 3. Thus, it can be inferred that the ZVS performance of \(Q_3\) is hard to achieve, and the ZCS performance of the rectifier diodes is lost.

![Figure 7. Key waveforms of type CDEF.](image)
From the key waveform of the four different types, the soft-switching performance characteristic can be concluded as in Table 2. It can be seen that only the ACD type and the CDE type have the ZVS characteristic for the primary-side switches, and that the ZCS characteristic for the secondary-side diodes only occurs at the CDE type. Thus, for the efficiency and reliability of the converter, only type ACD and type CDE are accepted, while type CDF and type CDEF should be avoided.

Table 2. Soft-switching performance under different working modes.

<table>
<thead>
<tr>
<th>Working Mode</th>
<th>ACD</th>
<th>CDE</th>
<th>CDF</th>
<th>CDEF</th>
</tr>
</thead>
<tbody>
<tr>
<td>Primary-Side ZVS</td>
<td>Achieved</td>
<td>Achieved</td>
<td>Lost</td>
<td>Hard to Achieve</td>
</tr>
<tr>
<td>Secondary-Side ZCS</td>
<td>Lost</td>
<td>Achieved</td>
<td>Lost</td>
<td>Lost</td>
</tr>
</tbody>
</table>

3. Design Principle of the Hybrid Control Method

As Section 2 reveals, ZVS performance only occurs at type ACD and type CDE. Traditional FHA analysis is hard to clarify and it is hard to distinguish the mode characteristic when applying hybrid PSM and PFM. In this section, time domain analysis is provided to calculate the turn-on current and turn-off current; thus, the ZVS boundary conditions can be figured out to determine the suitable hybrid PSM and PFM strategy. To reduce the turn-off loss of the primary-side switches, deducing the minimum turn-off current working point is another determination condition for the hybrid control strategy.

As the equivalent circuit shows in Figure 3, the time-domain equations of the A, C, D, E, and F stages can be listed as (2)–(7). The equations include the state variables i_{Lr}, v_{Cr} and i_m. No matter what the stage is, the ampere–second balance equation of output-filter capacitor C_{out} can be listed as (8).

At stage A, V_{AB} equals V_{in}, L_r and C_r resonate, L_m is clamped by \(-nV_o\), and i_m decreases linearly. The equation is shown as (2).

\[
\begin{align*}
\frac{di_{Lr}}{dt} &= i_{Lr}(t_0) \cos \omega_r(t - t_0) \\
&\quad - \left[ v_{Cr}(t_0) - V_{in} - nV_o \right] \sin \omega_r(t - t_0) / \sqrt{L_r} \\
v_{Cr}(t) &= V_{in} + nV_o + \left[ v_{Cr}(t_0) - V_{in} - nV_o \right] \cos \omega_r(t - t_0) \\
&\quad + \sqrt{L_r/C_r} i_{Lr}(t_0) \sin \omega_r(t - t_0) \\
i_m(t) &= i_m(t_0) - nV_o / L_m(t - t_0)
\end{align*}
\]

where \(t_0\) is the start time of the mode and \(\omega_r\) is shown in (3).

\[
\omega_r = 2\pi f_r
\]

At stage C, V_{AB} equals V_{in}, L_r and C_r resonate, L_m is clamped by \(nV_o\), and i_m increases linearly. The equation is shown as (4).

\[
\begin{align*}
\frac{di_{Lr}}{dt} &= i_{Lr}(t_0) \cos \omega_r(t - t_0) \\
&\quad - \left[ v_{Cr}(t_0) - V_{in} + nV_o \right] \sin \omega_r(t - t_0) / \sqrt{L_r} \\
v_{Cr}(t) &= V_{in} - nV_o + \left[ v_{Cr}(t_0) - V_{in} + nV_o \right] \cos \omega_r(t - t_0) \\
&\quad + \sqrt{L_r/C_r} i_{Lr}(t_0) \sin \omega_r(t - t_0) \\
i_m(t) &= i_m(t_0) + \frac{nV_o}{L_m}(t - t_0)
\end{align*}
\]
At stage D, \( V_{AB} \) equals 0, \( L_r \) and \( C_r \) resonate, \( L_m \) is clamped by \( nV_o \), and \( i_m \) increases linearly. The equation is shown as (5).

\[
\begin{align*}
 i_L(t) &= i_{Lx}(t_0) \cos \omega_r(t - t_0) \\
 &- \frac{1}{\sqrt{1+m}} v_{Cf}(t_0) \sin \omega_r(t - t_0) / \sqrt{L_r} \\
 v_{Cf}(t) &= -nV_o + [v_{Cf}(t_0) + nV_o] \cos \omega_r(t - t_0) \\
 + \sqrt{\frac{T_s}{L_r}} i_{Lx}(t_0) \sin \omega_r(t - t_0) \\
 i_m(t) &= i_{Lx}(t_0) + nV_o / \tau_m(t - t_0)
\end{align*}
\]

At stage E, \( V_{AB} \) equals 0. \( L_m \) joins in with the resonance between \( L_r \) and \( C_r \), and \( L_m \) is no longer clamped by \( nV_o \) or \(-nV_o\). The equation is shown as (6).

\[
\begin{align*}
 i_L(t) &= i_{Lx}(t_0) \cos \frac{1}{\sqrt{1+m}} \omega_r(t - t_0) \\
 &- \frac{1}{\sqrt{1+m}} v_{Cf}(t_0) \sin \frac{1}{\sqrt{1+m}} \omega_r(t - t_0) / \sqrt{L_r} \\
 v_{Cf}(t) &= v_{Cf}(t_0) \cos \frac{1}{\sqrt{1+m}} \omega_r(t - t_0) \\
 + \sqrt{1+m} \sqrt{\frac{T_s}{L_r}} i_{Lx}(t_0) \sin \frac{1}{\sqrt{1+m}} \omega_r(t - t_0) \\
 i_m(t) &= i_{Lx}(t_0)
\end{align*}
\]

where \( m \) is the inductance ratio of \( L_m \) to \( L_r \).

At stage F, \( V_{AB} \) equal to 0. \( L_r \) and \( C_r \) resonate, \( L_m \) is clamped by \(-nV_o\), and \( i_m \) decreases linearly. The equation is shown as (7).

\[
\begin{align*}
 i_L(t) &= i_{Lx}(t_0) \cos \frac{1}{\sqrt{1+m}} \omega_r(t - t_0) \\
 &- \frac{1}{\sqrt{1+m}} v_{Cf}(t_0) \sin \frac{1}{\sqrt{1+m}} \omega_r(t - t_0) / \sqrt{L_r} \\
 v_{Cf}(t) &= nV_o + [v_{Cf}(t_0) - nV_o] \cos \omega_r(t - t_0) \\
 + \sqrt{\frac{T_s}{L_r}} i_{Lx}(t_0) \sin \omega_r(t - t_0) \\
 i_m(t) &= i_{Lx}(t_0) - nV_o / \tau_m(t - t_0)
\end{align*}
\]

By combing the resonance equation and the voltage gain, the turn-on current and turn-off current of the primary-side switches can be figured out under different phase-shifting-angle-\( \theta \) and switching-frequency-\( f_s \) combinations, to design the optimum combination. The design principle is shown in Figure 8. At a specific \( V_{in} \), \( V_o \) and \( P \) working condition, different \( \theta \) and \( f_s \) values can be combined for the steady-state operation of the LLC converter. If the voltage gain is qualified for this working condition, then calculate the turn-on currents of the primary-side switches to estimate the ZVS performance; if the ZVS performance is obtained, then calculate the turn-off currents of the primary-side switches to acquire an optimum \( \theta \) and \( f_s \) combination for minimum turn-off currents. Therefore, the hybrid control is determined based on the ZVS constraint and minimum turn-off currents.

The schematic of the control circuit is shown in Figure 9. The control system adopts a single-loop voltage control. The sampling output voltage \( V_o \) is compared with the output reference voltage \( V_{o,ref} \) to obtain the error value, and then the frequency \( f_s \) or duty cycle \( D \) is obtained by the PI controller. When the frequency \( f_s \) does not reach the peak or valley value, the output duty cycle \( D \) of the phase-shift control unit is only determined by the output reference voltage \( V_{o,ref} \). When the load is light and the frequency reaches the peak value, if the output voltage is still higher than the reference value the phase-shift control unit will be triggered, making the phase-shift angle increase or decrease to \( D \), so as to achieve regulation voltage output. However, when the load is light, the phase-shift angle is
large. Even if the converter is in the CDE combination mode, ZVS is easily lost, due to the small magnetizing current $i_m$.

![Figure 8](image_url)  
**Figure 8.** Hybrid control-strategy design principle.

Table 3 shows the parameters of an LLC converter, where the output voltage $V_o = 200–500$ V, the voltage-gain range is 2.5 times, and $V_o = 400–500$ V is the boost mode of the LLC converter, while $V_o = 200–400$ V is the buck mode. To explain the design principle of the hybrid control method, $V_o = 300$ V, $P = 200$ W, 400 W, 600 W, 800 W and 1 kW are presented for the hybrid control design.

**Figure 9.** Schematic of control circuit.

Table 3. Value of Converter Parameters.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC Input Voltage $V_{in}$</td>
<td>400 V</td>
</tr>
<tr>
<td>DC Output Voltage $V_o$</td>
<td>200–500 V</td>
</tr>
<tr>
<td>Resonant frequency $f_r$</td>
<td>150 kHz</td>
</tr>
<tr>
<td>Switching frequency $f_s$</td>
<td>90–300 kHz</td>
</tr>
<tr>
<td>Rated power $P$</td>
<td>1 kW</td>
</tr>
<tr>
<td>Resonant inductance $L_r$</td>
<td>94 μH</td>
</tr>
<tr>
<td>Resonant capacitance $C_r$</td>
<td>13.3 nF</td>
</tr>
<tr>
<td>Magnetic $I_m$</td>
<td>470 μH</td>
</tr>
<tr>
<td>Transformer ratio $n$</td>
<td>1:1</td>
</tr>
</tbody>
</table>

For the sake of description, define the function of phase-shifting angle $\theta$ and equivalent duty cycle $D$ of $V_{AB}$ as shown in Equation (9), and define normalized frequency $f_n$ as shown in Equation (10).

$$D = 1 - \frac{\theta}{180^\circ} \quad 0^\circ \leq \theta \leq 180^\circ$$

\[ (9) \]
\[ f_n = \frac{f_s}{f_r} \]  \hspace{1cm} (10)

Figure 10a shows different \( f_n \) and duty-cycle-\( D \) combinations at \( V_o = 300 \) V under different loads, and it can be concluded that there are innumerable combinations at a certain voltage gain and load. Figure 10b shows the turn-on current of \( Q_3 \) in the lagging bridge under different the combinations shown in Figure 10a. The minimum current for ZVS performance can be inferred from (11).

\[ i_b = \frac{(C_3 + C_4) V_{in}}{t_{dead}} = \frac{2 \times 120 \times 10^{-6} \times 400}{200 \times 10^{-9}} = 0.48 \text{A} \]  \hspace{1cm} (11)

where \( C_3 = C_4 = 120 \) pF, and \( t_{dead} = 200 \) ns.

(a) \( D \) and \( f_n \) satisfy the voltage gain. 
(b) Turn-on current of \( Q_3 \).

(c) Sum of turn-off current. 
(d) Determine optimum combination.

Figure 10. Hybrid control design for \( V_{in} = 400 \) V, and \( V_o = 300 \) V under different loads.

It can be seen that when the \( f_s \) is low and \( D \) is small, the turn-on current is not large enough to achieve ZVS, and the worst condition is that the polarity is opposite for the ZVS performance. To meet the ZVS constraint, the minimum value of duty cycle \( D \) needs to be constrained. Based on the ZVS constraint, the sum of the turn-off current of the leading-bridge and the lagging-bridge switches can be calculated. As Figure 10c shows, the operating point of the minimum-sum turn-off current is shown in a solid circle, and at that point the switching loss can be greatly reduced compared to the single PFM.

In the actual working condition, two control degrees will make the system complicated. For simplicity and reliability of the control, the combination of \( f_n \) and \( D \) can be treated as a constant \( D \), which varies as the reference output voltage, and thus the constant \( D \) can be set as 0.75, as Figure 10d shows. The actual working point is close to the minimum turn-off current working point, which still maintains a small turn-off loss and leaves enough margin for the ZVS performance.

For simplicity and practicality, the adaptive duty cycle \( D \) can be summarized as a function of the reference output voltage based on the time-domain calculation results. Figure 11 is the fitted curve of the adaptive duty cycle \( D \) and reference \( V_o \), and the fitted equation is shown in (12).
By presetting the duty cycle $D$ according to the reference $V_o$, the two control degrees can be simplified to one degree, and no additional circuits need to be added.

$$D = \begin{cases} 0.002V_o + 0.1909 & V_o \in [200, 400] \\ 1 & V_o \in [400, 500] \end{cases}$$  \hspace{1cm} (12)

4. Simulation Results

Based on the parameter shown in Table 3, simulation results are provided in this section. The proposed hybrid control method is compared to the single PFM or single PSM. The comparative data include the turn-on current of the lagging bridge, the sum of the turn-off current, peak resonant current and RMS resonant current, and the switching frequency $f_s$ under full load and half load.

Figure 12 shows the $V_{AB}$ and $i_{Lr}$ waveforms when applying three control strategies under $V_o = 200$ V and $P = 1$ kW. For hybrid control, it can be seen that the ZVS performance is achieved: the turn-off current is 2.32A (lagging bridge) and 8.38A (leading bridge), and the sum is 10.7 A. For the single-PFM control method, although ZVS performance is achieved, the sum of the turn-off current is 16.54 A, which is 1.55 times that of 10.7 A. For single PSM, the ZVS performance of the lagging bridge is lost, due to a large phase-shifting angle. Table 4 presents all the comparative data: it can be seen the peak value and RMS value of $i_{Lr}$ under hybrid control are almost same as the single PFM. Due to the reduced turn-off current and $f_s$, the turn-off loss is reduced by about 39.7%.

![Figure 12. Simulation results under three control strategies @ $V_o = 200$ V, $P = 1$ kW.](image)

**Table 4. Comparison of Different Control Strategies Under $V_o = 200$ V, $P = 1$ kW.**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Hybrid Control</th>
<th>Single PFM</th>
<th>Single PSM</th>
</tr>
</thead>
<tbody>
<tr>
<td>$D$</td>
<td>0.6</td>
<td>1</td>
<td>0.33</td>
</tr>
<tr>
<td>ZVS</td>
<td>achieved</td>
<td>achieved</td>
<td>lost</td>
</tr>
<tr>
<td>Sum of turn-off current/A</td>
<td>10.70</td>
<td>16.54</td>
<td></td>
</tr>
<tr>
<td>$i_{Lr}$ (peak)/A</td>
<td>8.43</td>
<td>8.30</td>
<td></td>
</tr>
<tr>
<td>$i_{Lr}$ (rms)/A</td>
<td>5.61</td>
<td>5.65</td>
<td>5.81</td>
</tr>
<tr>
<td>$f_s$/kHz</td>
<td>177</td>
<td>190</td>
<td>142</td>
</tr>
<tr>
<td>Turn-off loss–reduction ratio</td>
<td>39.7%</td>
<td>Reference</td>
<td></td>
</tr>
</tbody>
</table>

Due to the limited space in this paper, the experiment results of other $V_o$ and $P$ values are concluded in Tables 5–7, while simulation waveforms are not provided. Table 5 shows...
the simulation results under $V_o = 200 \, \text{V}$, $P = 500 \, \text{W}$; the conclusion is that when applying the hybrid control, ZVS performance is achieved, while it is lost under single PSM, and the sum turn-off current is 5.66 A. For single PFM, though ZVS performance is achieved, the sum of the turn-off current is 9.1 A, which is 1.62 times that of 5.66 A, and the peak value and RMS value of $i_{Lr}$ in the hybrid control are almost the same as the single PFM. Due to the reduced turn-off current and $f_s$, the turn-off loss is reduced by about 44.7%.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Hybrid Control</th>
<th>Single PFM</th>
<th>Single PSM</th>
</tr>
</thead>
<tbody>
<tr>
<td>$D$</td>
<td>0.6</td>
<td>1</td>
<td>0.33</td>
</tr>
<tr>
<td>ZVS achieved</td>
<td>achieved</td>
<td>achieved</td>
<td>lost</td>
</tr>
<tr>
<td>Sum of turn-off current/A</td>
<td>5.66</td>
<td>9.1</td>
<td></td>
</tr>
<tr>
<td>$i_{Lr}$ (peak)/A</td>
<td>4.46</td>
<td>4.58</td>
<td></td>
</tr>
<tr>
<td>$i_{Lr}$ (rms)/A</td>
<td>2.84</td>
<td>2.88</td>
<td>3.27</td>
</tr>
<tr>
<td>$f_s$/kHz</td>
<td>213</td>
<td>240</td>
<td>142</td>
</tr>
<tr>
<td>Turn-off loss–reduction ratio</td>
<td>44.7%</td>
<td>Reference</td>
<td></td>
</tr>
</tbody>
</table>

Table 5. Comparison of Different Control Strategies Under $V_o = 200 \, \text{V}, P = 500 \, \text{W}$.

Table 6 shows the simulation results under $V_o = 300 \, \text{V}, P = 1 \, \text{kW}$. Due to $V_o = 300$, the adaptive $D$ changes to 0.79, according to (12). The conclusion is that when applying the hybrid control, ZVS performance is achieved, while it is lost under single PSM, and the sum of the turn-off current is 5.53 A. For single PFM, though ZVS performance is achieved, the sum of the turn-off current is 9.36 A, which is 1.69 times that of 5.53 A, and the peak value and RMS value of $i_{Lr}$ in the hybrid control are almost the same as the single PFM. Due to the reduced turn-off current and $f_s$, the turn-off loss is reduced by about 43.5%.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Hybrid Control</th>
<th>Single PFM</th>
<th>Single PSM</th>
</tr>
</thead>
<tbody>
<tr>
<td>$D$</td>
<td>0.79</td>
<td>1</td>
<td>0.55</td>
</tr>
<tr>
<td>ZVS achieved</td>
<td>achieved</td>
<td>achieved</td>
<td>lost</td>
</tr>
<tr>
<td>Sum of turn-off current/A</td>
<td>5.53</td>
<td>9.36</td>
<td></td>
</tr>
<tr>
<td>$i_{Lr}$ (peak)/A</td>
<td>5.75</td>
<td>5.48</td>
<td></td>
</tr>
<tr>
<td>$i_{Lr}$ (rms)/A</td>
<td>3.83</td>
<td>3.84</td>
<td>4.29</td>
</tr>
<tr>
<td>$f_s$/kHz</td>
<td>174</td>
<td>182</td>
<td>142</td>
</tr>
<tr>
<td>Turn-off loss–reduction ratio</td>
<td>43.5%</td>
<td>Reference</td>
<td></td>
</tr>
</tbody>
</table>

Table 6. Comparison of Different Control Strategies Under $V_o = 300 \, \text{V}, P = 1 \, \text{kW}$.

Table 7. Comparison of Two Control Strategies Under $V_o = 200 \, \text{V}, P = 1 \, \text{kW}$.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Hybrid Control</th>
<th>Single PFM</th>
<th>Single PSM</th>
</tr>
</thead>
<tbody>
<tr>
<td>$D$</td>
<td>0.67</td>
<td>1</td>
<td>0.43</td>
</tr>
<tr>
<td>ZVS achieved</td>
<td>achieved</td>
<td>achieved</td>
<td>lost</td>
</tr>
<tr>
<td>Sum of turn-off current/A</td>
<td>12.80</td>
<td>17.6</td>
<td></td>
</tr>
<tr>
<td>$i_{Lr}$ (peak)/A</td>
<td>8.6</td>
<td>8.8</td>
<td></td>
</tr>
<tr>
<td>$i_{Lr}$ (rms)/A</td>
<td>5.66</td>
<td>5.80</td>
<td>5.60</td>
</tr>
<tr>
<td>$f_s$/kHz</td>
<td>182</td>
<td>190</td>
<td>142</td>
</tr>
<tr>
<td>Turn-off loss–reduction ratio</td>
<td>30.3%</td>
<td>Reference</td>
<td></td>
</tr>
</tbody>
</table>

5. Experiment Results

In this section, a 1-kW/$V_o = 200–500 \, \text{V}$ experimental prototype is built to validate the theoretical analysis and feasibility of the hybrid control method. The experimental platform is shown in Figure 13, and the parameter of the converter is shown as Table 3. Three groups of experiments are developed including $V_o = 200 \, \text{V}$ ($P = 1 \, \text{kW}/500 \, \text{W}$), and $V_o = 300 \, \text{V}$ ($P = 1 \, \text{kW}$) under the hybrid control method, single PFM and single PSM. To ensure ZVS performance, the adaptive duty cycle $D$ of the hybrid control maintains a margin of about 10% with respect to the theoretical-analysis results.
Figure 13. Experimental platform.

Figure 14 shows the waveforms of $V_o$, $i_{LR}$ and $V_{AB}$ under $V_o = 200$ V and $P = 1$ kW, and the experiment results are concluded in Table 7. The adaptive $D$ is 0.66 according to reference $V_o$, and it can be seen that ZVS performance can be obtained under hybrid control, while it is lost under the single-PSM control method. The sum of the turn-off current under hybrid control is 12.8 A, while it is 17.6 A under single PFM; the turn-off loss under hybrid control can reduce to about 30.3% of that of single PFM when taking $f_s$ into consideration. The peak value and RMS value of $i_{LR}$ are 8.6 A and 5.66 A under hybrid control, while they are 8.8 A and 5.8 A under single PFM. The switching frequency $f_s$ of the two control methods are 182 kHz and 190 kHz, which means the other power losses are almost the same for the two control methods.

Figure 15 shows the waveforms of $V_o$, $i_{LR}$ and $V_{AB}$ under $V_o = 200$ V and $P = 500$ W, and the experiment results are concluded in Table 8. The adaptive $D$ is 0.66 according to reference $V_o$, and it can be seen that ZVS performance can be obtained under hybrid control, while it is lost under the single-PSM control method. The sum of the turn-off current under hybrid control is 7.2 A, while it is 9.2 A under single PFM; the turn-off loss under hybrid control can reduce by about 18.6%, compared to that of the single PFM when taking $f_s$ into consideration. The peak value and RMS value of $i_{LR}$ are 4.6 A and 2.72 A under hybrid control, while they are 4.6 A and 2.92 A under single PFM. The switching frequency $f_s$ of the two control methods are 226 kHz and 248 kHz, which means the other power losses are almost the same for the two control methods.
Table 8. Comparison of Two Control Strategies Under \( V_o = 200 \text{ V} \), \( P = 500 \text{ W} \).

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Hybrid Control</th>
<th>Single PFM</th>
<th>Single PSM</th>
</tr>
</thead>
<tbody>
<tr>
<td>( D )</td>
<td>0.67</td>
<td>1</td>
<td>0.42</td>
</tr>
<tr>
<td>ZVS achieved</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Sum of turn-off current / A</td>
<td>7.2</td>
<td>9.2</td>
<td></td>
</tr>
<tr>
<td>( i_{Lr} ) (peak) / A</td>
<td>4.6</td>
<td>4.6</td>
<td></td>
</tr>
<tr>
<td>( i_{Lr} ) (rms) / A</td>
<td>2.72</td>
<td>2.92</td>
<td>3.17</td>
</tr>
<tr>
<td>( f_s / \text{kHz} )</td>
<td>226</td>
<td>248</td>
<td>142</td>
</tr>
<tr>
<td>Turn-off loss–reduction ratio</td>
<td>18.6%</td>
<td>Reference</td>
<td></td>
</tr>
</tbody>
</table>

Figure 16 shows the waveforms of \( V_o, i_{Lr} \) and \( V_{AB} \) under \( V_o = 300 \text{ V} \) and \( P = 1 \text{ kW} \), and the experiment results are concluded in Table 9. The adaptive \( D \) changes to 0.88; according to the reference \( V_o \), it can be seen that ZVS performance can be obtained under hybrid control, while it is lost under single PSM. The sum of the turn-off current under hybrid control is 9.6 A, while it is 10.4 A under single PFM; the turn-off loss under hybrid control can reduce by about 8.7% compared to that of the single PFM when taking \( f_s \) into consideration. The peak value and RMS value of \( i_{Lr} \) are 5.8 A and 4 A under hybrid control, while they are 5.6 A and 3.96 A under single PFM; the switching frequency \( f_s \) of the two control methods are 183 kHz and 185 kHz, which means the other power losses are almost same for the two control methods. Owing to the adaptive \( D \) (0.88) being large, the hybrid control comes close to the single PFM, and thus the turn-off loss–reduction ratio is small. Therefore, the hybrid control method is more suitable for wide voltage-gain range application where the voltage gain reaches a low value.

Table 9. Comparison of Two Control Strategies Under \( V_o = 300 \text{ V} \), \( P = 1 \text{ kW} \).

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Hybrid Control</th>
<th>Single PFM</th>
<th>Single PSM</th>
</tr>
</thead>
<tbody>
<tr>
<td>( D )</td>
<td>0.88</td>
<td>1</td>
<td>0.56</td>
</tr>
<tr>
<td>ZVS achieved</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Sum of turn-off current / A</td>
<td>9.6</td>
<td>10.4</td>
<td></td>
</tr>
<tr>
<td>( i_{Lr} ) (peak) / A</td>
<td>5.8</td>
<td>5.6</td>
<td></td>
</tr>
<tr>
<td>( i_{Lr} ) (rms) / A</td>
<td>4</td>
<td>3.96</td>
<td>4.33</td>
</tr>
<tr>
<td>( f_s / \text{kHz} )</td>
<td>183</td>
<td>185</td>
<td>142</td>
</tr>
<tr>
<td>Turn-off loss–reduction ratio</td>
<td>8.7%</td>
<td>Reference</td>
<td></td>
</tr>
</tbody>
</table>

For the LLC resonant converter, its main loss comes from the turn-off loss and on-state loss. Through hybrid modulation, the switching frequency can be reduced and the turn-off current can be reduced. When the operation conditions are consistent, lower loss and higher efficiency can be achieved than with the single-modulation technique. To verify the effectiveness of hybrid control for reducing the power loss, the efficiency comparison is developed under hybrid control, single PFM and single PFM at \( V_o = 200 \text{ V} \) and \( 300 \text{ V} \). As Figure 17 shows, the efficiency of the hybrid control is higher than PFM and PSM. At \( V_o = 200 \text{ V} \), the efficiency improvement is between 0.75 and 1.4%, compared to the single PFM. At \( V_o = 300 \text{ V} \), due to the adaptive \( D \) being large, the operating characteristic of the hybrid control is nearly that of the single PFM, and the improvement of efficiency is
slight. For PSM, due to the loss of ZVS performance, the efficiency reduces to below 90%, at $V_o = 200 \, \text{V}$.

![Figure 17](image.png)

**Figure 17.** Efficiency comparison.

6. Conclusions

In this paper, a hybrid control method with adaptive PSM and PFM is proposed to maintain ZVS performance and reduce turn-off loss for the LLC resonant converter operating at wide voltage-gain range. Mode characteristics and soft-switching performance characteristics of different modes when applying PSM are illustrated. Based on time domain analysis, the turn-on current of the primary-side switches is calculated, and thus the ZVS boundary of different phase-shifting-angle-$\theta$ and switching-frequency-$f_s$ combinations is inferred. In addition, to acquire the optimum steady-state operation of the converter, the turn-off current is calculated; by making the converter work near to the minimum turn-off current operating point, the turn-off loss can be greatly reduced. With the principles of the ZVS boundary and minimum turn-off current, the phase-shifting angle $\theta$ of PSM is designed to be adaptive to the reference output voltage, no additional circuits are needed, and the two control degrees ($\theta$ and $f_s$) are simplified to one ($f_s$). The simulation and a 1 kW/output voltage 200–500 V experimental prototype are built to validate the feasibility and effectiveness of the proposed hybrid control method. Both results show the hybrid control method can maintain ZVS performance at wide output voltage range and wide load range when compared to single PSM, and the sum of the turn-off currents of the primary-side leading- and lagging-bridge switches is reduced when compared to single PFM; thus, turn-off loss is reduced. The efficiency comparison validates the fact that the hybrid control method has less power loss than single PSM and single PFM. Through the proposed hybrid-modulation technology, efficiency can be improved, and for data centers and charging-pile power consumption sites, a small efficiency increase can bring huge benefits.

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**Conflicts of Interest:** The authors declare no conflicts of interest.

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