A Novel and Simple Hybrid DC-DC Converter of Resonant Forward and PWM Flyback

Han Peng*، Mengtian Yu†، Jin Ke‡، Ming Xu†
Email: hanpeng@fsp-powerland.com، yumengtian@nuaa.edu.cn، jinke@nuaa.edu.cn، mingxu@fsp-powerland.com
*FSP-Powerland Inc., Nanjing China, 210014. Currently with HuaZhong University of Science and Technology, Wuhan China
† Nanjing University of Aeronautics and Astronautics, Nanjing China, 210016
‡FSP-Powerland Inc., Nanjing China 210014

Abstract— A novel single switch, hybrid DC-DC converter is proposed in this paper, where an unregulated resonant forward converter is cascaded with a PWM regulated flyback converter. The converter output is formed by resonant forward circuit utilizing transformer leakage inductance and a resonant capacitor. It transfers the energy fully to the output and is only dependent with transformer turns ratio. Zero current turn off is realized at secondary forward diode without reverse recovery loss. Both forward part and flyback part share the same primary transformer winding. The output of the flyback part, instead of delivering magnetizing flux energy to the load directly, is moved to the primary side and connected in series with input source. With fixed transformer turns ratio, a PWM regulated flyback part is employed to regulate the output voltage. The proposed topology maximizes the capability of energy delivering and is able to ensure both high efficiency and simple control methodology with a wide operating range and high output current. To verify the concept, a high frequency 400V to 20V/65W DC-DC converter for adapter power supply at 200 KHz switching frequency is designed and tested with the proposed architecture.

Keywords—Resonant forward, PWM flyback, cascaded, zero current switching, high efficiency

I. INTRODUCTION

In high-input, low-output DC-DC conversion systems, single power transistor topologies, as flyback [1-2] and forward converter [3-5], have been widely used with the benefits of simple architecture, low cost, and high reliability. Flyback converter, where the energy is transferred through mutual inductance, has relatively low efficiency due to high switching loss [1-2] with increased output power. Forward converter, which utilizes output filter for energy transferring, requires magnetizing energy reset circuit to avoid transformer saturation. It increases the voltage stress over the main power switch and limits the output power capability of forward converter to sub-500W [3-5]. The hybrid forward-flyback topology that merges the merits of each has been studied by many researchers over the past several decades [6-14]. By utilizing both magnetizing and output filter inductor as energy delivering elements, output energy in each cycle becomes larger with reduced the number of switching effectively [9]. However, poor efficiency is still a major impediment to its widespread application. Soft switching techniques as well as voltage clamping techniques are required to improve the switching loss over power transistors [8, 10, 14]. Forward/flyback series output operation makes such topology more attractive for low-input high-output applications with low output current [12-13]. Series connected output capacitors will significantly lower the effective output capacitance. Therefore, huge capacitors are required for low output ripple.

Based on the concept of $\Sigma$-Sigma converter [15-17], a hybrid resonant-forward/PWM flyback converter is proposed in this paper targeting at high-input voltage, low-output voltage with high output current, shown in Fig. 1(c). Different from traditional forward-flyback converter, the secondary stage of flyback converter is moved to the primary side and connected in series with input supply. The output of proposed converter output is formed by forward converter with LC resonant network which can realize ZCS at secondary diode [18-20]. Unregulated forward part processes the majority of...
output power, and PWM regulated flyback part delivers the energy stored in the magnetizing inductance to the output. The proposed architecture maximizes the capability for efficient energy delivering.  A 400V to 20V/65W power converter is constructed and tested to evaluate the proposed architecture.

The organization of the paper is as follows. Section II discussed operation principles of the proposed series-input-single-output converter. Section III discussed the output characteristics followed by design guidelines explored in Section IV. The prototype measurement results are presented in Section V and conclusions are drawn in Section VI.

II. Operation Modes and Principles

As shown in Fig.1(c), the proposed hybrid converter is composed by series connected forward and flyback converter to fully utilizing the benefits of both topologies. Both forward part and flyback part share the same primary winding \( N_1 \) and power transistor \( Q \). The output of flyback converter is connected in series with primary winding so that the effective converter input becomes: \( V_{in\_equiv} = V_{in} + V_c \), where \( V_c \) is the output of the flyback part. The output stage of the proposed topology is formed only by resonant forward with resonant tank composed by \( L_r \) and \( C_r \), where \( L_r \) can be transformer leakage inductance. Output LC filter (\( L_f \) and \( C_f \)) is inserted to reduce output ripple. Fig. 2 provides the critical operation waveforms for both continuous conduction mode (CCM) and discontinuous conduction mode (DCM) distinguished by flux continuity in flyback part. In the following analysis, the parasitics of power transistor and diodes are ignored and the output of the proposed converter is treated as constant current source. The equivalent circuits at different operation stages are shown in Fig. 3.

\[ t_0 \rightarrow t_1 \]: Shown in Fig. 3(a), switch \( Q \) is turned on at time \( t_0 \) and \( D_2 \) conducts. \( L_r \) and \( C_r \) start to resonate with a sinusoidal waveform at \( i_{Lr} \). Magnetizing current starts to increase linearly. At time \( t_1 \), resonant current \( i_{Lr} \) is reduced to zero and the resonant process ends. Diode \( D_2 \) is turned off at zero current with no reverse recovery loss.

\[ t_1 \rightarrow t_2 \]: As depicted in Fig. 3(b), transformer magnetizing current \( i_{Lm} \) continues to increase in a linear way with switch keeps on. As \( D_2 \) turns off, energy stored in the resonant capacitor \( C_r \) starts to deliver to the output linearly.

\[ t_2 \rightarrow t_3 \]: Shown in Fig. 3(c), when \( Q \) turns off at \( t_2 \), the circuit starts to operate in flyback mode. Energy stored in magnetizing inductance (\( L_m \)) is discharging linearly through \( D_1 \) into the capacitor \( C_f \). \( C_f \) keeps discharging until the next switching cycle starts.

If the magnetizing current drops to zero before the start of next cycle, the converter will operate in DCM mode. DCM shares the same operation states from \( t_0 \rightarrow t_1 \) as CCM operation and has additional time state as:

\[ t_0 \rightarrow t_4 \]: As plotted in Fig. 3(d), magnetizing current keeps at zero and the switch \( Q \) is still off. \( C_f \) keeps discharging until the next switching cycle starts. Drain-source voltage across \( Q \) starts to be discharged before \( Q \) turns on. Therefore, valley

---

**Figure 2.** Theoretical waveforms of proposed hybrid forward-flyback converter in CCM (a) and DCM (b) operation.

**Figure 3.** Operation states of proposed converter: (a-c) is for CCM operation; (d) is only for DCM operation.
voltage turning on will be realized to achieve less switching loss.

The flyback part in the proposed topology not only provides the path for resetting magnetizing current, but also transfers the energy stored in the magnetizing inductance to the output by cascaded input stage. Therefore, the proposed hybrid topology utilizes the full transformer energy.

### III. OUTPUT CHARACTERISTIC

The proposed hybrid resonant-forward PWM-flyback converter is formed by cascaded resonant forward converter and flyback converter. Both forward and flyback parts share the same primary transformer winding. Instead of delivering energy to the output, the secondary stage of flyback part is connected in series with primary winding. The output stage is only formed by resonant forward part. To fully understand the converter output characteristics, output stage and input stage are studies respectively.

#### A. Output Stage

The equivalent circuit of the converter output is shown in Fig. 4. $C_r$ is resonantly charged by the input source during interval $[t_0-t_1]$, and linearly discharged to the output for the rest of the switching cycles as depicted in Fig. 2.

Assume that resonant current $i_{Lr}$ starts at zero at $t_0$ and the initial voltage across $C_r$ is $V_{C,0}$, the voltage and current of the resonant network during charging interval can be derived as:

$$v_{Cr}(t) = V_2 - (V_2 - V_{C,0}) \cos \omega_r(t - t_0) - Z_r i_{Lr} \sin \omega_r(t - t_0)$$

$$i_{Lr}(t) = I_o \cos \omega_r(t - t_0) + \frac{V_2 - V_{C,0}}{Z_r} \sin \omega_r(t - t_0)$$

where $\omega_r = \sqrt{L_r/C_r}$, $Z_r = \sqrt{L_r/C_r}$, and $V_2$ is the voltage across secondary winding $N_2$.

Current at $L_r$ turns to zero at the end the resonating interval, as: $i_{Lr}(t_1) = i_{Lr}(t_0) = 0$. Hence, the resonant time interval is derived as:

$$t_r = t_1 - t_0 = \frac{\pi + 2 \arctan \frac{I_o Z_r}{V_2 - V_{C,0}}}{\omega_r}$$

By inserting Eq. 3 into Eq. 1, voltage across resonant capacitor $C_r$ at the end of resonating interval $t_1$ can be expressed as:

$$v_{Cr}(t_1) = 2V_2 - V_{C,0}$$

Therefore, the average voltage $V_o$ during resonating interval is $V_2$.

The equivalent output stage in the linear discharging interval is plotted in Fig. 4(b) and the voltage across $C_r$ can be expressed as:

$$v_{C_r}(t) = v_{C_r}(t_1) - \frac{I_o}{C_r}(t-t_1)$$

In steady state operation, based on charge balance equation, we will have $v_{C_r}(t_1) = v_{C_r}(t_0) = V_{C,0}$. So, the average voltage across $C_r$ during one steady state switching cycle is $V_2$. Since the average output voltage is equal to the average $C_r$ voltage, we will have:

$$V_o = V_{C_r} = V_2 = V_1 \frac{N_2}{N_1}$$

The output stage characteristic of resonant current ($i_{Lr}$) in terms of resonant voltage ($V_{C_r}$) is depicted in Fig. 5. With the center point of $[V_2, L_r]$, the resonant stage starts at $[V_{C,0}, 0]$ and ends at $[2V_2 - V_{C,0}, 0]$. The long radius in the oval curve is $r_1 = \sqrt{I_o^2 + (V_2 - V_{C,0}/Z_r)^2}$ and the short radius is $r_2 = r_1 \cdot Z_r$.

The linear cycle follows the blue curve. The output of converter shows a fixed relationship with primary side voltage and turns ratio between $N_1$ and $N_2$. Therefore, the resonant forward part in the proposed converter is irrelevant to the duty cycle of $Q$.

#### B. Input Stage

Although the output relies only on turns ratio and voltage across primary winding, $V_1$ is composed by the summary of supply voltage $V_a$ and flyback output $V_c$, as $V_1 = V_a + V_c$. The series connected flyback converter is used to reset the magnetizing current so that the proposed topology makes full use of transformer energy. The equivalent circuits of primary side are plotted in Fig. 6 and the output of flyback part will be discussed under CCM and DCM operation respectively.

1) Under CCM operation

![Connecting diagram](http://www.itrans24.com/landing1.html)
When $Q$ turns on, energy is transferred to the secondary side and $D_1$ is off.

![Figure 6. Equivalent circuit of input stage](image)

When $Q$ is off, magnetizing current starts to charge $C_c$ through $N_3$ and $D_1$ turns on.

$$i_C = -(i_{t_{m1}} + i_{t_{m2}}) = -(i_{t_{m1}} + N_3 i_{t_{m2}}/N_1)$$

(12)

Let's assume $T_R$ is the time interval for magnetizing current to reset, which corresponds to $t_3-t_4$ in Fig. 2b. When $Q$ is off, we will have: $V_{in} = -N_1 V_C / N_1$ and $i_C = N_1 i_{t_{m1}} / N_3$.

$T_R$ can be derived based on volt-second balance at $L_m$ as:

$$T_R = N_3 (V_{in} + V_C) T_{ON}/N_1 V_C$$

Hence, $V_C$ at DCM operation is:

$$V_C = \frac{N_1 T_{ON}^2 V_{in}^2}{2 N_2^2 L_m I_o - N_1^2 T_{ON}^2 V_{in}}$$

(13)

Inserting Eqn. 11 and Eqn. 13 into Eqn. 6, output voltage of the proposed hybrid converter at both CCM and DCM operation can be expressed as:

$$V_o = \begin{cases} \frac{(1-D) N_2}{(1-D) N_1 - DN_3} V_{in} & \text{for CCM} \\ \frac{2 N_2^2 L_m I_o}{2 N_1 N_2 L_m I_o - N_1^2 TD^2 V_{in}} & \text{for DCM} \end{cases}$$

(14)

The output gain characteristic at different load currents is plotted in Fig. 7. The black solid line shows the boundary between flyback part CCM and DCM operation.

2) **Under DCM operation**

An extra zero magnetizing current time interval $[t_1-t_4]$ exists in DCM operation. When $Q$ is on, the voltage at primary winding is the same as Eqn. 7 and the current is:

$$V_{in} + V_C = V_i = L_i \frac{di_1}{dt_1}$$

(7)

$$\Delta_l = (i_{t_{max}} - i_{t_{min}}) = \frac{V_{in} + V_C}{L_1} T_{ON}$$

(8)

When $Q$ is off, magnetizing current starts to charge $C_c$ through $N_1$ and $D_1$ turns on.

$$V_C = V_3 = -L_3 \frac{di_3}{dt_3}$$

(9)

$$\Delta_3 = (i_{t_{max}} - i_{t_{min}}) = \frac{V_C}{L_3} T_{OFF}$$

(10)

where: $T_{ON} + T_{OFF} = T_{SW}$. According to flux-balancing relationship of $\Delta_l N_1 = \Delta_3 N_3$, flyback output in CCM mode can be derived as:

$$V_C = \frac{DN_3}{(1-D) N_1 - DN_3} V_{in}$$

(11)
reset to zero quicker. Hence, the flyback part falls into DCM operation at higher duty cycles. With \( N_1 / N' \) equals to 0.24, the converter needs to operate at switching frequency higher than 280 KHz and duty cycle larger than 0.35 to ensure full resonance. However, with \( N_1 / N' \) equals to 0.2, the output stage will never work under full resonance. Therefore, the design of transformer and resonant network should consider the effect of switching frequency, output voltage and duty cycle comprehensively.

\[
V_{\text{DSS}} = V_{\text{in}} \left( 1 + \frac{N_1}{N_3} \right) V_C
\] (15)

\( V_{\text{DSS}} \) is proportional related with flyback output \( (V_c) \) and \( N_1 / N' \). Although flyback part is employed to reset the magnetizing current and deliver its energy to the output, the majority of output energy is supported by resonant forward part. To avoid \( V_c \) being too large to increase the voltage stress, the duty ratio of the proposed converter is usually less than 0.5. The relationship of \( V_c \), \( N_1 / N' \) and duty cycle in CCM operation is drawn in Fig. 9. \( V_c \) increases linearly with \( N_1 / N' \) when \( N_1 / N' \) is small and \( V_{\text{DSS}}/V_{\text{in}} \) is quite constant over a wide range of \( N_1 / N' \). For example, to have \( V_c \) equals to 0.01 \( V_{\text{in}} \), the converter can be designed with duty ratio of 0.1 and \( N_1 / N' = 0.09 \). It can also be designed with duty ratio of 0.5 and \( N_1 / N' = 0.01 \). The maximum drain-source voltage will vary from 1.12×\( V_{\text{in}} \) to 2.02×\( V_{\text{in}} \) with duty cycle changes from 0.1 to 0.5. Therefore, voltage rating is the primary index for selecting power transistor, which is closely related with transformer turns ratio and duty cycle. In high-input, low-output applications, primary side of the converter usually has relatively low current, making conduction loss negligible in this topology. So, power transistors with high switching capability will be preferred.

**B. Flyback output diode and forward output diode selection**

The forward current of both flyback output diode \( (D_1) \) and forward output diode \( (D_2) \) are:

\[
I_{D1_{fblk}} = \frac{N_1(V_{\text{in}} + V_c) D T}{N_3 L_m}
\] (16)

\[
I_{D2_{fwd}} = I_o + \sqrt{I_o^2 + \left( V_{2-V_c} \right)^2} / Z_{r}
\] (17)

In the resonant output, the peak diode current depends on the feature impedance of the resonant network and is much higher than average output current. To reduce power consumption, schottky diode with low forward drop is preferred for secondary output diode.

**C. Transformer design**

Transformer turns ratio can be calculated through Eqn. 14 at certain input, output voltage and switching frequency. Magnetizing inductance determines the operation mode of flyback part with the boundary current as:

\[
I_o = I_o = \frac{N_1^2 V_{\text{in}} T D (1 - D)}{2N_2 N_3 L_m}
\] (18)

When output current is higher than \( I_o \), the converter works under CCM operation and vice versa. With larger \( L_m \), the converter is more likely to work at CCM operation. In this design, the leakage inductance of the transformer is employed as resonant inductor.
To verify the concept, a 400V-to-20V/65W resonant forward PWM flyback converter is designed with the key design specifications summarized in Tab. I.

<table>
<thead>
<tr>
<th>TABLE I. KEY SPECIFICATIONS AND COMPONENTS OF THE PROPOSED CONVERTER</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Parameters</strong></td>
</tr>
<tr>
<td>Input voltage</td>
</tr>
<tr>
<td>Output voltage</td>
</tr>
<tr>
<td>Rated power</td>
</tr>
<tr>
<td>Primary winding turns</td>
</tr>
<tr>
<td>Forward winding turns</td>
</tr>
<tr>
<td>Flyback winding turns</td>
</tr>
<tr>
<td>Switching frequency</td>
</tr>
<tr>
<td>Resonant inductance</td>
</tr>
<tr>
<td>Resonant capacitance</td>
</tr>
<tr>
<td>Forward output capacitance</td>
</tr>
<tr>
<td>Flyback output capacitance</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Components</th>
<th><strong>Symbols</strong></th>
<th><strong>Part numbers</strong></th>
</tr>
</thead>
<tbody>
<tr>
<td>Main switch</td>
<td>$Q$</td>
<td>STB6NK90Z</td>
</tr>
<tr>
<td>Flyback diode</td>
<td>$D_1$</td>
<td>B3200B</td>
</tr>
<tr>
<td>Forward diode</td>
<td>$D_2$</td>
<td>SD860S</td>
</tr>
<tr>
<td>Transformer core</td>
<td>$T$</td>
<td>PQ26/20</td>
</tr>
</tbody>
</table>

The designed prototype board of hybrid resonant forward and PWM flyback converter is shown in Fig. 9 with the design parameters shown in Tab. 1. The leakage inductance of the designed transformer is 0.25μH and the resonant frequency is 464 KHz. Fig. 10 shows the experimental waveforms of gate-source voltage ($v_{GS}$), resonant inductance current ($i_{Lr}$), resonant capacitor voltage ($v_{Cr}$) and diode current ($i_{D1}$) in CCM and DCM at 20V, 65W output respectively. In this experiment, 250 KHz switching frequency is selected for CCM and 200 KHz is used for DCM. The converter operates with full resonant cycles and therefore, zero current turned off is realized at secondary diode.

Converter falls into DCM operation with reduced switching frequency and duty cycle. Fig. 11 shows the gate-source voltage ($v_{GS}$), resonant inductance current ($i_{Lr}$), primary winding current ($i_o$) and drain source voltage of main switch ($V_{DS}$). At flyback DCM operation, main power transistor is turned on at lower drain-source voltage, but undergoes high voltage stress due to higher flyback voltage ($V_c$).

The proposed converter operating under complete resonance and non-complete resonance is compared in Fig. 12 with same outputs of 10.3V and 1A. The on time of the main switch is reduced from 1.7 μs to 1.3 μs, while switching frequency is increased from 200 KHz to 250 KHz to maintain same outputs. Peak resonant current and $V_{DS}$ of main switch is much higher at non-complete resonant condition than at complete resonant condition.

Depicted in Fig. 13, converter output efficiency is measured at different output power with both Si power switch and SiC power switch respectively. With Si power transistor, the peak output efficiency is 91.7% at 65W. While, the peak efficiency reaches to 95% for SiC power transistor. This is due to smaller gate charge, output capacitor and on-resistor of SiC power switch.
The resonant forward output is independent of the proposed topology makes the most of transformer energy and magnetizing current reset and output voltage regulation. The output of flyback circuit as output to realize main power transmission with winding and power transistor. It utilizes resonant forward flyback part in the proposed topology share the same primary winding.

VI. CONCLUSIONS

A novel hybrid resonant-forward, PWM-flyback DC-DC converter is presented in this paper. The forward part and flyback part in the proposed topology share the same primary winding and power transistor. It utilizes resonant forward circuit as output to realize main power transmission with minimum power consumptions. The output of flyback circuit is connected in series with the primary side to realize the magnetizing current reset and output voltage regulation. The proposed topology makes the most of transformer energy and combines the advantages of both forward and flyback converter. The resonant forward output is independent of the load and duty cycle and simple regulation can be achieved by flyback PWM control. Zero current turned off is achieved at secondary diode. All of the above make the proposed topology a good candidate for efficient high frequency switching applications, especially for high-input, low-output and high current conditions. A 400V to 20V/65W power conversion prototype is built to demonstrate the proposed architectures. The peak measured efficiency at initial test is 92% and the optimized efficiency reaches to 95%.

REFERENCES


