High-Efficiency Battery Charging With Switching Resonant Converter

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Abstract—The high-frequency resonant converter has numerous well-known advantages over the traditional hard-switching converters. The most important advantage is that it offers a lower switching loss and a higher power density. Additionally, the soft-switching current waveform characterizes a lower electromagnetic interference (EMI). This study presents the circuit configuration with the least components to realize a highly efficient solar energy battery charger with a zero-voltage-switching resonant converter. The optimal values of the resonant components are determined by applying the characteristic curve and the electric functions derived from the circuit configuration. The experiment demonstrates the switching on and off of the main switches in a solar energy battery charger with a zero-voltage-switching resonant converter, wherein the switches are all operated using zero-voltage switching. The circuit efficiency in the overall charging process exceeds 80%.

Index Terms—Battery charger, photovoltaic array, resonant converter, soft-switching, zero-voltage-switching (ZVS).

I. INTRODUCTION

RAPID technological changes have led to power electronic products playing a crucial role in daily life. Energy storage equipment is a commonly used form of power electronic products. However, the conventionally adopted battery chargers produce power losses that incur power dissipation during charging. Therefore, the charging method is especially important. Various charging approaches engender various charging efficiencies and also indirectly influence the life of a battery charger. This study analyzes the charging losses and power dissipation of a buck zero-voltage-switching resonant solar battery charger, and then, improves the charging efficiency. Additionally, Taiwan is located in a subtropical zone that is close to the equator, and southern Taiwan, particularly, is full of sunshine during summer. Consequently, the energy collected on photovoltaic arrays is utilized as the source of a battery charger for saving energy. However, the output voltage of photovoltaic arrays varies with the sunshine. Unstable output voltage shifts the operating point of a zero-voltage-switching resonant converter. Consequently, this study designs a novel approach in which the output terminal of photovoltaic arrays is in series with a closed-loop buck converter to stabilize the input voltage of the zero-voltage-switching resonant converter, and prevent the operating point of the charger from varying with the sunshine. Fig. 1 illustrates the whole system block diagram of the zero-voltage-switching resonant converter for photovoltaic arrays.

Batteries are extensively used in various applications, including telecommunication power supply, electric vehicles, uninterruptible power supplies, photovoltaic systems, portable electronics products, and others. Photovoltaic arrays are presently employed to supply electricity to remote or inaccessible areas. A battery charger is crucial in a photovoltaic array and the charge mode markedly affects battery life and capacity. Topologies with high switching frequencies are used to reduce the charging current ripple and extend the battery life. However, as the switching frequency further increases, switching losses and electromagnetic interference (EMI) noise arise. Hence, to solve this problem, the switching frequency is increased by reducing the switching losses through a so-called zero-voltage-switching (ZVS) circuit.

The traditional battery charger, which extracts power from an ac-line source, requires a thyristor ac/dc converter rectifier, with an equivalent series resistance, to control the power flow to charge the battery system. Such a charging circuit necessarily draws a high ripple charging current. This shows notoriously low efficiency, and is associated with a large volume. Accordingly, as the concern about the quality of a charger grows, a charging circuit for reducing the ripple and extending the battery life becomes more important in designing the battery storage systems. Several charging circuits have been proposed to overcome the disadvantages of the traditional battery charger. The linear power supply is the simplest. A 60-Hz transformer is required to deliver the output within the desired voltage range. However, the linear power supply is op-erated at the line frequency, which makes it large both in size and weight. Besides, the system conversion efficiency is low because the transistor operates in the active region. Hence, when higher power is required, the use of an over-sized and oversized line-frequency transformer makes this approach impractical.

The high-frequency operation of the conventional converter topologies depends on a considerable reduction in switching losses to minimize size and weight. Many soft-switching techniques have been proposed in recent years to solve these problems. Unfortunately, switching losses in these new circuits can be reduced only at the expense of substantially increased...
voltage and current stresses in the power switches, resulting in a considerable increase in the conduction loss. However, flexible control techniques can be used to realize switching-mode converters with small circuit components of high efficiency and low cost. In switch-mode topologies, controllable switches are operated under hard-switching conditions, resulting in increased switching losses, switching noise, and electromagnetic interference. In an attempt to overcome these shortcomings, much effort has been made to find a less expensive filtering approach and topology of the charging circuit, to be able to offer a competitive price in the consumer market. This work considers the application of zero-voltage-switching to a charger to minimize the switching losses, switching noise, and high-frequency electromagnetic interference. In the developed approach, a resonant tank is inserted between the input and the battery. With the zero-voltage-switching topology, a charger with high efficiency can be achieved without an additional power switch or sophisticated control circuit.

II. COMPARISONS BETWEEN THE RESONANT CONVERTER AND THE TRADITIONAL PULSE WIDTH MODULATOR (PWM) CONVERTER

In the conventionally adopted power converters, the switching frequency must be increased to lower the volume and weight in order to enhance the power density. Paradoxically, this also leads to an increase in the switching losses, switching stress, and EMI. The snubber circuit is parallel to the switch in the conventional solution for reducing the $dv/dt$ surge resulting from the switching-off, while the switching-on buffer circuit is in series with the switch to reduce the $di/dt$ surge resulting from the switching-on. Although snubber circuits can reduce the switching losses, switching stress, and EMI, the wholeloss remains unchanged and efficiency becomes a significant concern. The resonant converter is developed by using the series or parallel combination of inductor and capacitor to generate the resonance situation. The switch is switched to achieve zero-voltage or zero-current switching in the resonant situation, to effectively solve the switching losses, switching stress, and EMI.

A. Traditional PWM Power Converter

Switching losses are produced by the current linearly increases owing to the switching-on operation in the active region, when the transistor is generally adopted as the switch of the converter. The traditional PWM power converter is designed to use the power transistor as the switching switch, and be operated in a switching-on and switching-off model to control the power transistor duty cycle to achieve the buck/boost topology. Four main circuit configurations have been developed: 1) buck converter, 2) boost converter, 3) buck-boost converter, and 4) C'uk converter. Although the traditional PWM power converter enhances the low efficiency shortcoming of the conventionally adopted linear power converter, the converter switching frequency must be increased to reduce the volume and weight, simultaneously increasing the switching losses, switching stress, and EMI.

1) Switching Loss: The switch must withstand the input terminal voltage, and during switching-off, the switch current must immediately reduce to zero. In practice, the increase of voltage $V_c(t)$, and the decrease of current $i_c(t)$, both exhibit a slope form, the switch voltage and switch current overlap in the form of a cross, owing to the existence of these slopes. The generated power $P_{loss}$ at the intersection is the switching loss. Switching losses increase with the increase in switching frequency. Fig. 2 shows the switching losses and EMI in a traditional PWM power converter. $V_c(t)$ is the voltage drop between the collector and emitter of transistors, and $i_c(t)$ is the current flow from the collector.

2) EMI: The $dv/dt$ is generated in the voltage waveform at the instance of switching-off, because of the stray inductance during switching-off. The $di/dt$ is generated at the instance of switching-on, owing to bad reverse recovery characteristics during the switching-on period. These two situations are the source of EMI. EMI also appears increasingly serious with increasing switching frequency.

3) Switching Stress: Fig. 3 displays the switching path in the $v_c$–$i_c$ plane in a switched power converter. The switch must withstand the switching stress within a safe operating area (SOA), because of the high voltage and large current in the switch during switching. Power semiconductor elements withstand large switching stress and certainly have some effect on performance.

B. Resonant Power Converter

The resonant power converter utilizes the resonance theory by incorporating a LC resonant circuit from the conventionally adopted PWM converter. The switching-on and switching-off conditions, both operate in zero-voltage to reduce the switching...
In a single switching cycle, the circuit operates in the following four modes.

1) **Mode I: Linear stage (between \( t_0 \) and \( t_1 \)):** Prior to \( t_0 \), the power switch \( Q \) is on, and conducts a drain current that equals the output current \( I_o \), and the freewheeling diode \( D_m \) is off. Fig. 6 depicts the equivalent circuit. At \( t_0 \), \( Q \) is turned off. The current through the resonant inductor \( L_r \) does not change instantaneously, and so the current is diverted around the power switch through the resonant capacitor \( C_r \). The current of the resonant inductor equals the output current \( I_o \) and the capacitor voltage \( v_{cr} \), which increases, as given by

\[
\frac{d}{dt} v_{cr} = \frac{1}{C_r} \int_0^t I_o \, dt = \frac{I_o}{2} t. \tag{1}
\]

Voltage across freewheeling diode \( D_m \) is determined by

\[
v(t) = V_{in} - v_{cr}(t) = V_{in} - \frac{I_o}{2} t. \tag{2}
\]

\( v_i \) declines to zero at time \( t_1 \), when \( D_m \) is turned on by soft-switching. The constant output current linearly increases the voltage across the resonant capacitor, until the input voltage is reached

\[
V_{in} = C_r I_o. \tag{3}
\]

We can obtain (4) by substituting (3) into (2)

\[
v_i(t) = V_{in} - \frac{1}{2} \frac{I_o}{t_1} t. \tag{4}
\]

Model I is completed when \( t = t_1 \), namely \( v_{cr}(t_1) = V_{in} \). The time interval \( T_1 \) in Model I is obtained using (5). Moreover, Model II is initiated when \( v_i \) decreases to zero

\[
T = \frac{V_{in} C_r}{I_o}. \tag{5}
\]

2) **Mode II: Resonant stage (between \( t_2 \) and \( t_3 \)):** After \( t_1 \), the freewheeling diode \( D_m \) becomes forward-biased, and \( C_r \) and \( L_r \) resonate. The instantaneous voltage across \( C_r \) and the resonant inductor current can be evaluated, respectively as

\[
i_{L_r}(t) = I_o \cos[\omega_o(t - t_1)] \frac{1}{t_1} \tag{6}
\]

\( i_{L_r} \) can be divided into four modes, whose associate
equivalent circuits are displayed in Fig. 6. The parameters are defined as follows. Characteristic impedance: \( Z_0 = L_r / C_r \); resonant angular frequency: \( \omega_0 = 1 / L_r C_r \); resonant frequency: \( f_r = \omega_0 / 2\pi \); switching period: \( T_s \).
The resonant inductor current \( i_{lr} (t) \) is linearly returned from its negative peak of minus \( I_o \) to its positive value of positive \( I_o \). Consequently, \( i_{lr} (t) \) increases linearly and \( i_{om} \) decreases linearly.

This model is completed at \( t = t_3 \) when \( v_c(t_3) = 0 \) and \( i_{lr}(t_3) = I_o \).

The commutation interval in this stage is expressed by

\[
T_{III} = \frac{L_r I_o}{\omega_o Z_o I_o} \left[ 1 - \cos \omega (t_2 - t_1) \right].
\]

Notably, the voltage across the switch \( Q \) is zero, when the power switch is turned on. It enables the turn-on switching loss to be avoided and the total efficiency of the converter to be increased accordingly.

4) Mode IV: Freewheeling stage (between \( t_3 \) and \( t_4 \)): When \( i_i(t) \) reaches \( I_o \) at \( t_3 \), the freewheeling diode \( D_m \) is turned off, and the zero-voltage-switched converter resembles a conventional square-wave power processor. The charging current flows through power switch \( Q \) and resonant inductor \( L_r \). Accordingly,

\[
i_{i}(t) = I_o \quad \text{and} \quad v_c(t) = 0 \tag{19}
\]

The power switch conducts \( I_o \) as long as it is kept on until \( t_4 \). At \( t_4 \), the power switch is turned off again, beginning another switching cycle. The duration of this mode is \( T_W \) expressed as

\[
T_W = T_1 - (T_1 + T_{II} + T_{III}).
\]

The input voltage of the filter can be considered the voltage \( v_i(t) \) across the diode \( D_m \) because of the relationship \( \omega_r \triangleq \frac{1}{\sqrt{L_r C_r}} \). Thus, the \( t_{m} c_o \)

The equation for \( v_i(t) \) is expressed by rearranging (4), (10) and (19)

\[
v(t) = V_m \left( 1 - \tau \right) \quad \text{for} \quad 0 < t < t_1 \tag{21}
\]
The output voltage of the filter is determined as the mean of \( v(t) \) and then determined using (22). Controlling the interval \( T_{IV} \) of the power switch, the average power supplied to the battery can be controlled.
The output voltage $V_o$ is determined as

$$V_{BA} = \frac{1}{T} \int_{0}^{s} v_x \, dt$$

where

$$v_x = V_{in} \left( 1 - \frac{t}{t_1} \right)$$

and

$$V_{BA} = V_{in} \left( 1 - f_s \right) \frac{T_3}{2}. \tag{23}$$

Let $\alpha = \omega_o (t_2 - t_3)$

$$l_o (T_s - T_{II} - T_{III}) = l_o \left[ (t_2 - t_1) - (t_3 - t_2) \right]$$

and

$$W_o = V_{BA} l_o T_s$$

The energy released by the filter inductor to the battery is

$$W_o = V_{in} l_o \int_{t_1}^{t_2} i_{lr} \, dt + V_{in} l_o \int_{t_2}^{t_3} i_{lr} \, dt$$

$$+ V_{in} l_o (T_s - t_3 - t_1) \tag{26}$$

Similarly, because of the condition $Z_o l_o > V_{in}$ must hold such that $l_o \omega_o L_r > V_{in}$

$$L_r > \frac{V_{in}}{l \omega_o} \tag{25}$$

Given $l_o$ and $T_s$, $T_i$, $T_{II}$ and $T_{III}$ and the output voltage $V_o$ can be determined. However, the voltage conversion ratio is normally best expressed in terms of load resistance $R$ and switching frequency $f_s$. $V_o = R l_o$, so the energy stored in the resonant inductor is

$$W_f = V_{in} \int_{t_1}^{t_2} i_{lr} \, dt + V_{in} \int_{t_2}^{t_3} i_{lr} \, dt$$

$$+ V_{in} l_o (T_s - t_3 - t_1) \tag{26}$$

The energy released by the filter inductor to the battery is

$$W_o = V_{in} l_o \int_{t_1}^{t_2} i_{lr} \, dt \tag{27}$$

$$\int_{t_1}^{t_2} i_{lr} \, dt = \frac{l_o - V_{in}}{\omega L} = -C_r V_{in} \tag{28}$$

$$\int_{t_2}^{t_3} i_{lr} \, dt = \frac{1}{2} 2 \cos^2 \left( \frac{L I}{2} \right) \tag{29}$$

Let $\alpha = \omega_o (t_2 - t_3)$

$$l_o (T_s - T_{II} - T_{III}) = l_o \left[ (t_2 - t_1) - (t_3 - t_2) \right]$$

$$= \frac{1}{T} \int_{0}^{s} v_x \, dt$$

$$= \int_{t_1}^{t_2} V_{in} \left( 1 - \frac{t}{t_1} \right) \, dt + V_{in} \int_{t_2}^{t_3} \, dt$$

$$= \frac{T_3}{T_s} t_1 \int_{t_1}^{t_3} \left( \frac{t_3 - t}{t_3 - t_1} \right) \, dt$$

$$V_{BA} = V_{in} \left( 1 - f_s \right) \frac{T_3}{2}. \tag{23}$$

$$W_o \text{ equals } W_f \text{ when the converter power dissipation is ignored.}$$

$$V_{BA} = \frac{1}{T} \int_{0}^{s} v_x \, dt$$

For a lossless system, in the steady state, these two energies
The output voltage varies with the switching frequency. Fig. 9 illustrates the equivalent circuit of Model IV and Fig. 10 shows key steady-state waveforms of the buck ZVS converter.

IV. DESIGN OF RESONANT ELEMENTS AND SWITCHING FREQUENCY

Fig. 5 illustrates the zero-voltage-switching resonant battery charger developed for photovoltaic arrays. The condition $Z_o I_o > V_{in}$ must hold to ensure that the operation is under zero-flux. Hence, the voltage ratio is expressed by (30) as

$$X = 1 - \frac{f_s}{2\pi f_r} \alpha + \frac{X(1 - \cos \alpha)}{r} + \frac{r}{2X}.$$  (30)

The relationship between input and output voltages is a function of the pulse width angle, the characteristic impedance of the resonant circuit and the output load current. This equation reveals that the output voltage can be controlled by varying the angle for any variation in input voltage and output load current.
The voltage ratio of the buck ZVS was numerically determined, as plotted in Fig. 11, with $f_s/f_r$ as the running parameter. Fig. 11 plots the dc voltage-conversion-ratio characteristics of the buck ZVS as functions of normalized output current $I_{oZ}/V_{in}$. The first step in designing the converter is to determine $f_s/f_r$, based on a set of dc characteristics curves for various $f_s/f_r$ in Fig. 11.

Table I

<table>
<thead>
<tr>
<th>Circuit Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage $V_{in}$</td>
<td>24V</td>
</tr>
<tr>
<td>Resonant inductor $L_r$</td>
<td>50µH</td>
</tr>
<tr>
<td>Resonant capacitor $C_r$</td>
<td>0.2µF</td>
</tr>
<tr>
<td>Characteristic impedance $Z_c$</td>
<td>16Ω</td>
</tr>
<tr>
<td>Switching frequency $f_s$</td>
<td>18KHz</td>
</tr>
<tr>
<td>Resonant frequency $f_r$</td>
<td>50KHz</td>
</tr>
<tr>
<td>Charging voltage $V_{ch}$</td>
<td>15V</td>
</tr>
<tr>
<td>$f_s/f_r$</td>
<td>0.36</td>
</tr>
<tr>
<td>Open circuit voltage of the battery</td>
<td>11V</td>
</tr>
<tr>
<td>Initial charging current $I_c$</td>
<td>2A</td>
</tr>
</tbody>
</table>

Fig. 12. Waveforms of the switching signal $V_{GS}$, the resonant voltage $v_x$, and the resonant current $i_L$.  

Fig. 13. Waveforms of the switching signal $V_{GS}$, and voltage $v_x$, of the freewheeling diode. 

Fig. 14. Waveforms of the charging current $I_c$, and the resonant current $i_r$. 

Fig. 15. Charging voltage curve.

V. EXPERIMENTAL RESULTS

A prototype of the buck converter with zero-voltage-switching resonant topology was established in a laboratory to confirm the functional operations. The developed charging circuit is applied to a 12 V, 4 Ah lead-acid battery. Table I lists the circuit parameters of the experiment results in the developed novel high-efficiency battery charger with a buck zero-voltage-switching resonant converter. The resonant capacitance $C_r = 0.2\mu F$, and resonant inductance $L_r = 50 \mu H$, are determined using (24) and (25).

Fig. 12 displays the waveform of the switch signal, the resonant voltage, and the resonant current $i_r$. Fig. 13 presents the waveform of the switching signal $V_{GS}$, and the voltage $v_x$ during freewheeling while Fig. 14 plots the waveform of charging.
battery charger with ZVS is significantly lower than that for the traditional hard-switching charger.

VI. CONCLUSIONS

This study presents the photovoltaic battery charger with ZVS technology for use in the charging test of a lead-acid battery charger, to demonstrate the effectiveness of the developed methodology. Under the same operating conditions, the measured temperature of power switches in the proposed battery charger with ZVS is maintained at 34 °C and is much lower than that of the traditional pulse-width-modulated (PWM) converter at 54 °C. The proposed battery charger with ZVS indeed reduces the temperature of the switch, reducing the switching losses. The circuit efficiency of the overall charging process exceeds 80% and greatly exceeds the 68% efficiency, of traditional converters.

REFERENCES
