Implementation and Analysis of an Improved Series-Loaded Resonant DC–DC Converter Operating Above Resonance for Battery Chargers

Ying-Chun Chuang, Yu-Lung Ke, Senior Member, IEEE, Hung-Shiang Chuang, and Hung-Kun Chen

Abstract—The well-established advantages of resonant converters, including simplicity of circuit configuration, ease of the control scheme, low switching losses, and low electromagnetic interference, among others, have led to their attracting more interest. This work develops a highly efficient battery charger with an improved series-loaded resonant converter for renewable energy applications to improve the performance of traditional switching-mode charger circuits. The switching frequency of the improved series-loaded resonant battery charger was at continuous conduction mode. Circuit operation modes are determined from the conduction profiles. Operating equations and operating theory are also developed. This study utilizes the fundamental wave approximation and a battery equivalent circuit to simplify the circuit analyses. The mean charging efficiency of the proposed topology is as high as 87.5%.

Index Terms—Battery charger, series-loaded resonant converter.

I. INTRODUCTION

RECHARGEABLE batteries are extensively applied in various applications such as cellular phones, laptop computers, uninterruptible power supplies, electrical vehicles, renewable energy storage systems, and others [1]–[4]. Such equipment continuously consumes electrical energy, and they require a charging circuit in a rechargeable battery [5]–[9]. For several years, most battery chargers available on the market were of linear-mode converters, in which an active power element regulates the output voltage. A linear-mode converter with an active power element is generally used as a variable resistance to dissipate unwanted or excess voltage [10]–[13]. Such an arrangement results in the dissipation of large amounts of power in the active power element, potentially reducing the charging efficiency to as low as 50%. The basic requirements of battery charger circuits are smallness and high efficiency. Their low efficiency has therefore prevented linear-mode converters from being applied to battery chargers, and since the early 1970s, the uptake of switch-mode power converters has been increasing. Unlike linear-mode converters, switch-mode power converters use active power switches to operate in either the saturation region or the cutoff region. Since either region will lead to a low switching voltage or a low switching current, power can be converted with higher efficiency using a switch-mode power converter as a battery charger circuit [14]–[18]. Accordingly, switch-mode battery chargers with efficiencies of greater than 70% can be easily designed at low cost and with relatively small size and light weight.

In all switch-mode power converters, the controllable switches are operated in a switch mode in which they turn on and off the entire charging current during each switching period. Hence, the controllable switches are subjected to high switching stresses and high switching power losses that increase linearly with the switching frequency of the battery charger. Another significant shortcoming of the switch-mode operation is the electromagnetic interference (EMI) that is generated by the large di/dt and dv/dt that are associated with a switch-mode operation. Unlike switch-mode converters, the combination of proper converter topologies and switching strategies can solve the problems of switching stresses, switching power losses, and EMI, by turning on and off each of the converter switches when either the switch voltage or the switch current is zero. A new class of dc–dc power converters was then introduced in the late 1980s. This group of topologies is known as soft-switching resonant converters. The potential advantage of the soft-switching resonant converters over linear-mode and switch-mode converters is the reduced switching power losses and, consequently, higher power density, with maintained high efficiency. Additionally, the higher switching frequency causes such converters to exhibit shorter transient responses [19]–[21].

The literature describes numerous soft-switching techniques to improve the charging behavior of resonant dc–dc converters, among which, the class-D half-bridge series-loaded resonant converter designed for battery chargers is the simplest and has various advantages. One of the main advantages of class-D half-bridge series-resonant converters is the low voltage across the active power switches, which equals half of the input supply voltage [22]–[26]. This makes them suitable for high input
Fig. 1. Block diagram of dc–dc resonant converter for battery chargers.

Fig. 2. Improved series-loaded resonant converter with a voltage-doubler rectifier for battery chargers.

Fig. 3. Idealized voltage and current waveforms.
rectifier for a switching frequency $f_s$ that exceeds the resonant frequency $f_r$. Operation above resonance is preferred because the power switches turn on at zero current and zero voltage; thus, the freewheeling diodes do not need to have very fast reverse-recovery characteristics. Therefore, only two diodes are required—instead of a bridge rectifier—as shown in Fig. 2. During the positive half-cycle of the resonant tank current, the power is supplied to the battery through diode $D_{r1}$. During the negative half-cycle of the resonant tank current, the power is supplied to the battery through diode $D_{r2}$, increasing the efficiency above that of a traditional bridge rectifier.

The class-D half-bridge series-loaded resonant converter with a voltage-doubler rectifier for battery chargers is analyzed based on the following assumptions.

1) The switching elements of the converter are ideal, such that the decline in forward voltage in the ON-state resistance of the switch is negligible.

2) The equivalent series resistance of the capacitance and stray capacitances is negligible.

3) The characteristics of passive components are assumed to be linear, time invariant, and frequency independent.

4) The filter capacitor $C_r$ at the output terminal is usually very large, and therefore, the output voltage across the capacitor can be treated as a dc voltage in each switching cycle.

5) The load quality factor of the improved class-D half-bridge series-loaded resonant converter is sufficiently high that the resonant current $i_{lr}$ is sinusoidal.

The steady-state operation of the improved series-load resonant charging circuit in one switching period includes four modes.

**A. Mode I: (Between $\omega_o t_0$ and $\omega_o t_1$)**

The periodic switching of the resonant energy tank voltage between $+V_s/2$ and $-V_s/2$ generates a square-wave voltage across the input terminal. Since the output voltage is assumed to be a constant $V_o$, the input voltage to the voltage-doubler rectifier $v_b$ is $V_o/2$ when $i_{lr}$ is positive and is $-V_o/2$ when $i_{lr}$ is negative. Hence, Fig. 4 shows the equivalent circuit of the class-D half-bridge series-loaded resonant converter with a voltage-doubler rectifier for the battery charger circuit in Fig. 2.

Before $\omega_o t_0$, the freewheeling diode $D_1$ is turned on and conducts a current that equals the resonant tank current $i_{lr}$; the active power switch $S_1$ is excited.

At the instant $\omega_o t_0$, the resonant tank current $i_{lr}$ reverses and naturally commutates from diode $D_1$ to the power switch $S_1$. In this mode, the power switches turn on naturally at zero voltage and at zero current. Accordingly, the active power switch is negative after turn-on and positive before turn-off. The initial condition of the capacitor $C_r$ is $V_{co}$. Then, the instantaneous resonant inductor current and the voltage across $C_r$ can be evaluated, where the angular resonance frequency $\omega_o = 2\pi f_o = 1/\sqrt{LrC_r}$ and the characteristic impedance $Z_o = \sqrt{Lr/C_r}$, respectively,

$$i_{lr}(t) = \frac{1}{Z_o} \left[ \frac{V_s}{2} - V_{co} - \frac{V_o}{2} \right] \sin \omega_o t$$

$$V_{cr}(t) = \frac{V_s}{2} - \frac{V_o}{2} - \left[ \frac{V_s}{2} - V_{co} - \frac{V_o}{2} \right] \cos \omega_o t.$$  

The current in the switches is turned on at zero voltage and zero current to eliminate turn-on losses, but the switches are turned off at nonzero current; therefore, turn-off losses may exit. Fortunately, small capacitors can be placed across the switches to act as snubbers to eliminate turn-off losses.

**B. Mode II: (Between $\omega_o t_1$ and $\omega_o t_2$)**

At $\omega_o t_1$, before the half-cycle of the current $i_{lr}$ oscillation ends, the switch $S_1$ is forced to turn off, forcing the positive current to flow through the bottom freewheeling diode $D_2$. Fig. 5 shows the equivalent circuit.

The negative dc voltage applied across the resonant tank causes the current that flows through the diode to go quickly to zero at $\omega_o t_2$. During this interval, the inductor current $i_{lr}$ is expressed as follows, where $I_{L1}$ is the initial current in the inductor $i_{lr}$:

$$i_{lr}(t) = \frac{1}{Z_o} \left[ \frac{V_s}{2} - V_{co} - \frac{V_o}{2} \right] \sin \omega_o t - I_{L1} \cos \omega_o (t - t_0)$$

$$I_{L1} = i_{lr}(t_1) = \frac{1}{Z_o} \left[ \frac{V_s}{2} - V_{co} - \frac{V_o}{2} \right] \sin \beta.$$  

The voltage $v_{cr}$ across the resonant capacitor $C_r$ is given by (4), where $V_{c1}$ is the initial voltage across the capacitor $C_r$.

$$V_{cr}(t) = V_{c1} + \left[ \frac{V_s}{2} - \frac{V_o}{2} - V_{c1} \right] \left[ 1 - \cos \omega_o (t - t_0) \right]$$

$$+ \frac{I_{L1}}{\omega_o} \sin \omega_o (t - t_0)$$

$$V_{c1} = V_{cr}(t_1) = \frac{V_s}{2} - \frac{V_o}{2} - \left[ \frac{V_s}{2} - V_{co} - \frac{V_o}{2} \right] \cos \beta.$$  

**C. Mode III: (Between $\omega_o t_2$ and $\omega_o t_3$)**

Before $\omega_o t_2$, the trigger signal $v_{gs2}$ excites the active power switch $S_2$. When the inductor current $i_{lr}$ changes direction, the freewheeling diode $D_2$ is turned off, and the active power switch $S_2$ is turned on. Fig. 6 shows the equivalent circuit.
Mode III begins at $\omega_o t_2$, when the diode $D_2$ is open circuited as shown in Fig. 6, producing a resonant stage between inductor $L_r$ and capacitor $C_r$. The zero-voltage turn-on of the active power switch $S_2$ is achieved because the current has already flowed through the freewheeling diode $D_2$ before the lower switch $S_2$ is turned on. The inductor $L_r$ and capacitor $C_r$ resonate. Then, the inductor current $i_{Lr}$ and the capacitor voltage $v_{cr}$ of the resonant circuit are as given by (5) and (6), where $V_{c2}$ is the initial voltage across the resonant capacitor $C_r$.

$$i_{Lr}(t) = \frac{1}{Z_o} \left[ -\frac{V_o}{2} - V_{c2} + \frac{V_o}{2} \right] \sin \omega_o (t - t_2)$$  \hspace{1cm} (5)

$$v_{cr}(t) = -\frac{V_o}{2} + \frac{V_o}{2} - \left[ -\frac{V_o}{2} - V_{c2} + \frac{V_o}{2} \right] \cos \omega_o (t - t_2)$$

$$V_{c2} = v_{cr}(t_2)$$

$$= v_{c1} + \left[ -\frac{V_o}{2} - \frac{V_o}{2} - V_{c1} \right] (1 - \cos \omega_o \alpha)$$

$$+ \frac{I_{L1}}{\omega_o} \sin \omega_o \alpha.$$  \hspace{1cm} (6)

At the instant $\omega_o t_3$, the active power switch $S_2$ is turned off, and Mode III ends.

**D. Mode IV: (Between $\omega_o t_3$ and $\omega_o t_4$)**

A turn-off trigger signal is applied to the gate of the active power switch $S_2$. Then, the inductor current naturally commutates from the active power switch $S_2$ to the freewheeling diode $D_1$. Fig. 7 shows the equivalent circuit.

Applying Kirchhoff’s law to Fig. 7 yields the inductor current $i_{Lr}$ as given by (7), where the initial current $I_{L3}$ in the inductor $L_r$ is given by (5) at $t = t_3$.

$$i_{Lr}(t) = \frac{1}{Z_o} \left[ \frac{V_o}{2} - V_{c3} + \frac{V_o}{2} \right] \sin \omega_o (t - t_3)$$

$$+ \frac{I_{L3}}{\omega_o} \cos \omega_o (t - t_3).$$  \hspace{1cm} (7)

Then, the initial inductor current can be written as follows:

$$I_{L3} = I_{Lr}(t_3) = \frac{1}{Z_o} \left[ -\frac{V_o}{2} - V_{c2} + \frac{V_o}{2} \right] \sin \omega_o (t_3 - t_2).$$

**III. OPERATING CHARACTERISTICS**

Fig. 8 shows the simplified equivalent circuit of improved class-D half-bridge series-loaded resonant converter with a voltage-doubler rectifier for battery chargers. The switching frequency of the active power switches is assumed to exceed the resonant frequency such that the resonant current is continuous. Given a large capacitive filter at the output terminal, the output voltage may be assumed to be constant.

The charger circuit in Fig. 2 can be simplified to the schematic circuit shown in Fig. 8 to facilitate the analysis of the operation of the class-D half-bridge series-loaded resonant converter with a voltage-doubler rectifier. Since the output voltage is assumed to be a constant $V_o$, then the input voltage to the voltage-doubler rectifier $v_b$ is $V_o/2$ when $i_{Lr}$ is positive and is $-V_o/2$ when $i_{Lr}$ is negative.

The improved class-D half-bridge series-loaded resonant converter with a voltage-doubler rectifier for battery chargers is analyzed based on the fundamental frequency of the Fourier series of the voltages and currents. Then, the output voltage $v_b$ of the voltage-doubler rectifier is given by a Fourier series which is given by

$$v_b(t) = \sum_{n=1,3,5,\ldots}^{\infty} \frac{2V_o}{n\pi} \sin (n\omega t).$$  \hspace{1cm} (9)
Fig. 9. Equivalent ac circuit of improved class-D half-bridge series-loaded resonant converter for battery chargers.

Equation (10) gives the fundamental component of voltage $v_b$

$$v_{b1} = \frac{2V_o}{\pi} \sin (\omega t).$$

(10)

The currents at the output of the voltage-doubler rectifier $i_{DR1}$ and $i_{DR2}$ are the full-wave rectified forms of the inductor current $i_{Lr}$. Hence, the average of the rectified inductor current $\langle i_{Lr} \rangle$ equals the output charging current $I_o$. If the inductor current $i_{Lr}$ is approximated as a sine wave of amplitude $I_{LM1}$, then the average value of charging current $I_o$ is given by

$$I_o = \frac{2I_{LM1}}{\pi}.$$ 

(11)

The output resistance in this equivalent circuit is determined from the ratio of voltage to current at the voltage-doubler rectifier. The following thus defines resistance:

$$R_e = \frac{V_{b1}}{I_{LM1}} = \frac{4}{\pi^2} \frac{V_o}{I_o}.$$ 

(12)

The relationship between the input and output is approximated by an ac circuit analysis using the fundamental frequencies of the voltage and current equations. Fig. 9 shows the equivalent ac circuit.

To achieve resonant operation, the resonant circuit must be underdamped. That is,

$$R_e \leq 2 \sqrt{\frac{L_r}{C_r}}.$$ 

(13)

The input part of the improved class-D half-bridge series-loaded resonant converter for battery chargers has a dc input voltage source $V_s$ and a set of bidirectional power switches. The active power switches are controlled to generate a square-wave voltage $v_o$. Since a resonant circuit forces a sinusoidal current, only the power of the fundamental component is transferred from the input source to the resonant circuit. Therefore, only the fundamental component of this converter needs be considered. The following defines a voltage transfer function of this improved converter:

$$\frac{V_o}{V_s} = \frac{1}{\sqrt{1 + \left( \frac{X_L - X_C}{R_e} \right)^2}}.$$ 

(14)

The reactance $X_L$ and $X_C$ depend on the switching frequency. Accordingly, the output voltage can be regulated by changing the switching frequency of the converter. The normalized output voltage $V_o/V_s$ is plotted as a function of $f_s/f_o$ at various loaded quality factors $Q$ (Fig. 10). The output voltage is maximal at the resonant frequency $f_o$. This figure indicates that the improved half-bridge series-loaded resonant converter has an output of twice the peak value of the traditional half-bridge series-loaded resonant converter [22].

The energy that flows into the battery during the interval $\omega_o t_0 \leq \omega_o t \leq \omega_o t_1$ is given by

$$W_{o1} = \int_{t_0}^{t_1} \frac{V_{bo}}{2} i_{Lr}(t) dt$$

$$= \frac{V_o}{2 \omega_o} \left[ \frac{V_s}{2} - V_{co} - \frac{V_o}{2} \right] (1 - \cos \beta).$$

(15)

The following term is defined:

$$A = \frac{1}{\omega_o} \left[ \frac{V_s}{2} - V_{co} - \frac{V_o}{2} \right] (1 - \cos \beta).$$

(16)

The energy that flows into the battery during the interval $\omega_o t_1 \leq \omega_o t \leq \omega_o t_2$ is given by

$$W_{o2} = \int_{t_0}^{t_2} \frac{V_{bo}}{2} i_{Lr}(t') dt'$$

$$= \frac{V_o}{2} \int_{0}^{t_2} \left\{ \frac{1}{Z_o} \left[ - \frac{V_s}{2} - \frac{V_o}{2} - V_{c1} \right] \sin \omega_o t' + I_{L1} \cos \omega_o t' \right\} dt'$$

$$= \frac{V_o}{2 \omega_o} \left\{ \frac{1}{Z_o} \left[ - \frac{V_s}{2} - \frac{V_o}{2} - V_{c1} \right] (1 - \cos \alpha) + I_{L1} \sin \alpha \right\}.$$ 

(17)

Substituting (3) and (4) into the aforementioned equation yields

$$W_{o2} = \frac{V_o}{2 \omega_o} \left\{ \left[ - \frac{V_s}{Z_o} + \frac{1}{Z_o} \left( \frac{V_s}{2} - V_{co} - \frac{V_o}{2} \right) (1 - \cos \beta) \right] - \frac{1}{Z_o} \left( \frac{V_s}{2} - V_{co} - \frac{V_o}{2} \right) (1 - \cos \alpha) \right. \right.$$

$$+ \left. \left. \frac{1}{Z_o} \left[ - \frac{V_s}{2} - V_{co} - \frac{V_o}{2} \right] \sin \beta \sin \alpha \right\}.$$ 

(18)
The following term is defined:

\[ B \equiv \left\{ \left[ \frac{V_S}{Z_o} + A - \frac{1}{Z_o} \left( \frac{V_S}{2} - V_{co} - V_o \right) \right] (1 - \cos \alpha) \right. \\
\left. + \frac{1}{Z_o} \left( \frac{V_S}{2} - V_{co} - V_o \right) \sin \beta \sin \alpha \right\}. \tag{19} \]

Hence, the total energy that flows into the battery during the interval \( \omega_o t_0 \leq \omega_o t \leq \omega_o t_2 \) is determined by

\[ W_o = W_{o1} + W_{o2} = \frac{V_o}{2\omega_o} (A + B). \tag{20} \]

The energy from the input dc source during the interval \( \omega_o t_0 \leq \omega_o t \leq \omega_o t_1 \) is given by

\[ W_{S1} = \frac{V_S}{2} \int_0^{t_1} i_{Lex}(t) \, dt = \frac{V_S}{2\omega_o} A. \tag{21} \]

The energy from the input dc source during the interval \( \omega_o t_1 \leq \omega_o t \leq \omega_o t_2 \) is given by

\[ W_{S2} = -\frac{V_S}{2} \int_0^{t_2-t_1} i_{Lex}(t') \, dt' = -\frac{V_S}{2\omega_o} B. \tag{22} \]

Accordingly, the energy that is generated by the input dc source is given by

\[ W_S = W_{S1} + W_{S2} = \frac{V_S}{2\omega_o} (A - B). \tag{23} \]

For a lossless system, these two energies are equal in the steady state. Therefore, (24) gives the output voltage

\[ V_o = V_S \frac{A - B}{A + B}. \tag{24} \]

The most important advantage of the improved series-loaded resonant converter is that the maximum output voltage \( V_o \) in Fig. 2 can approach the input dc source voltage \( V_S \), unlike in a traditional series-loaded resonant converter where \( V_o \) can only approach \( 0.5V_S \) [22]. Restated, the improved series-loaded resonant converter must switch only half as much current for the same \( V_S \) and output power. This advantage makes the improved series-loaded resonant converter the preferred configuration for rapidly charging battery applications.

IV. EXPERIMENTAL RESULTS

The input of the proposed improved series-loaded resonant converter was connected to a system that comprised a dc source with an output voltage of 20 V. A prototype of the battery charger with improved series-loaded resonant topology was established in a laboratory to verify the functional operations. The developed charger circuit is applied to a 12-V 48-Ah lead-acid battery. The conditions of the experiment were as follows: switching frequency \( f_s = 24 \) kHz, resonant frequency \( f_r = 22 \) kHz, charging current \( I_o = 8 \) A, charging voltage \( V_{BA} = 15 \) V, and the open-circuit voltage of battery \( V_{oc} = 11 \) V. Under these operating conditions, the two parameters of the improved series-loaded resonant converter are as follows:

\[ C_r = 3.2 \mu F \]
\[ L_r = 16 \mu H. \]

The waveforms were measured using a digital multimeter. Fig. 11 shows the waveforms of the trigger signals \( V_{GS1} \) and \( V_{GS2} \). Fig. 12 shows the voltage and current waveforms of the active power switch \( S_1 \). Fig. 13 shows the waveforms of the resonant voltage \( v_{cr} \) and the resonant current \( i_{cr} \). Fig. 14 shows the input and output voltage waveforms of the resonant tank terminals. Fig. 15 shows the voltage and current waveforms of the voltage-doubler rectifier diode \( D_{R1} \). Fig. 16 shows the voltage variation curve of the battery. The terminal voltage of the battery rises from 10.5 to 15.5 V in 500 min. Figs. 17 and 18 show the charging current and the charging efficiency,
presented the use of a battery charger with an improved series-loaded resonant topology in the charging test of a lead-acid battery charger to demonstrate the effectiveness of the developed approach. The circuit efficiency of the overall charging process exceeds 83%. Accordingly, the charging efficiency can be improved using an improved series-loaded resonant converter with voltage-doubler rectifier topology. Favorable performance is obtained at lower cost and with fewer components.

V. CONCLUSION

This work has designed an improved series-loaded resonant converter with a voltage-doubler rectifier for battery chargers. The circuit structure is simpler and cheaper than other control mechanisms which require many components. This paper has respectively. The charging current $I_o$ declines as the voltage $V_o$ of the battery increases. The charging current takes 330 min to fall below 6.8 A. The minimal and maximal efficiencies of the battery charging circuit are about 83% and 95%, respectively, and the mean charging efficiency of the charger is 87.5%.

REFERENCES


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