A Single-Stage Inductive-Power-Transfer Converter for Constant-Power and Maximum-Efficiency Battery Charging

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Abstract—This article proposes a single-stage inductive-power-transfer (IPT) converter operating as a wireless constant-power (CP) and maximum-efficiency battery charger. By maintaining a constant output power rather than providing a constant output current throughout the dominant stage of battery charging, the IPT converter can make the utmost of its power capability, thus having a faster charging rate. The proposed single-stage IPT converter adopts series-series compensation and includes a switch-controlled capacitor (SCC) and a semiactive rectifier (SAR) in the secondary side. Manipulating the SCC and the SAR to emulate the optimum impedance of the resonator and the load, we propose a novel operation approach combining the merits of load-independent transfer characteristic and load impedance matching, to achieve a simple solution to CP charging and maximum efficiency throughout the charging process. Since the control scheme is based on fixed operating frequency and secondary-side real-time regulation, wireless feedback communication is not required. Moreover, soft switching and low voltage stress can be easily achieved in this IPT converter.

Index Terms—Battery charging, constant power, inductive power transfer (IPT), maximum efficiency, soft switching.

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I. INTRODUCTION

INDUCTIVE power transfer (IPT) is a growing technology to wirelessly supply power in applications where physical connection is inconvenient or impossible, e.g., hostile environments being affected by dirt and moisture [1], [2]. Typically, with abilities to simplify charging operation and remove safety concerns associated with electrical connection, IPT converters are suitable for wireless battery charging in a variety of scenarios, such as mobile electronics, biomedical implants, small home appliances, and electric vehicles [3]–[7].

Thermal design for power electronics converters is mandatory [8]. In general, maximum extractable power of a power electronics converter is subject to practical thermal and physical constraints. Constant current (CC) charging is a common charging technique and a dominant charging process for widely used lithium-ion batteries [9]–[11]. As shown by the vertical CC line in Fig. 1, the charging current is kept constant, while the charging voltage is clamped to the terminal voltage of the battery and increases during charging. It can be observed that the charging power started with a minimum value and increased to a maximum value at the completion of CC charging. If CC charging is implemented, the IPT converter only delivers power at the maximum value for a very short duration near the completion of charging. Alternatively, to make the utmost of the power capability, the charger can control the output power to a predetermined maximum value and provide a constant-power (CP) charging for the battery [12]–[17], such that full-power delivery can be maintained throughout the whole charging process. As shown by the CP curve in Fig. 1, the charging current should be allowed to vary inversely with respect to the terminal voltage of
the battery to maintain the desired CP charging. Obviously, given an identical maximum charging power, CP charging provides a faster charging speed than that of CC charging [12]–[14]. Moreover, compared with CC charging, there is no problem of excessive thermal design with CP charging [15], [16].

For a conductive charger, it is relatively easy to facilitate CP charging in its battery management system, which operates the charger as a current source, constantly varying according to the charging power profile [12]–[17]. However, to our best knowledge, IPT converters with the ability of wireless CP charging are seldom explored in the literature, which motivates us to investigate the feasibility of the wireless CP charger. In general, an IPT converter should be designed to operate at some fixed operating frequencies with load-independent transfer characteristic for minimal control complexity and to operate within a restricted load range to achieve maximum efficiency [18]–[21]. However, only CC or constant voltage output is achievable at these fixed operating frequencies such that the output power of the IPT converter is determined by the load condition and cannot comply with the CP charging profile [22], [23]. An intuitive idea for CP output is using a two-stage IPT system, where a front-end converter can be used to modulate the input amplitude of the IPT converter or a load-side converter can be cascaded to the IPT converter for power regulation [24]–[27]. Due to the extra power stage, penalties of power loss, control complexity, and/or wireless feedback communication are inevitable. Moreover, keeping single-stage design in mind, the IPT converter should also have load matching ability to achieve high efficiency. Otherwise, the efficiency significantly degrades at some mismatched loading conditions [28]–[30]. Since the load range during battery charging is normally wide, it is difficult for a single-stage IPT converter to maintain the maximum efficiency, while permitting fixed operating frequency, soft switching, no extra cascading converter, and no wireless feedback communication [30]. Therefore, it is challenging for an IPT converter to achieve the required output for CP charging and maintain the maximum efficiency throughout the charging process.

Aimed at filling the gap of wireless CP charging, this article presents and explores a single-stage IPT converter which adopts series-series with a switched-controlled compensation capacitor (SCC) and a semiactive rectifier (SAR) in the secondary side. Combining the merits of load-independent transfer characteristic and load impedance matching, a novel operation approach is proposed for CP and maximum-efficiency charging. By controlling the SCC and the SAR, an optimum load and a constant secondary resonator current are maintained simultaneously, such that CP and maximum-efficiency charging can be simply implemented. The control scheme is based on fixed operating frequency and secondary-side real-time regulation, eliminating wireless feedback communication. Moreover, soft switching and low voltage stress can be easily achieved in this IPT converter.

II. PROPOSED WIRELESS CP BATTERY CHARGER

A. System Structure

Fig. 2 shows the proposed wireless CP battery charger based on a series-series compensated inductive power transfer (SSIPT) converter with an SCC and an SAR. In the schematic of the proposed system, the magnetic coupler has primary self-inductance $L_P$, secondary self-inductance $L_S$, and mutual inductance $M$. The coupling coefficient is defined as $k = \frac{M}{\sqrt{L_P L_S}}$. Coil losses in the primary and the secondary are represented by resistances $R_p$ and $R_{s, w}$, respectively. Both coils of the magnetic coupler are compensated by a capacitor in series connection. $C_p$ is the primary compensation capacitor with fixed capacitance value, while a fixed-value capacitor $C_1$ as well as an SCC in series connection is used for secondary compensation with variable capacitance. The SCC consists of a fixed-value $C_2$ and two MOSFET switches $Q_a$ and $Q_b$, with equivalent variable capacitance $C_{SCC}$. $D_a$ and $D_b$ are the antiparallel body diodes of $Q_a$ and $Q_b$, respectively. Compared with a single SCC, series connection of a fixed-value capacitor and an SCC can help to reduce the voltage stress of the SCC switches, which will be discussed in detail in Section IV-A. $v_{S, SCC}$ is the voltage across the SCC, and $i_{S, SCC}$ is the current flowing through the SCC. DC voltage source $V_i$ is modulated into ac voltage $v_p$ at an angular frequency $\omega$ to drive the primary coil by a full-bridge inverter with four MOSFET switches $Q_{1−4}$. AC output is rectified to dc output by the SAR with output filter capacitor $C_f$. The SAR consists of two diodes $D_5$ and $D_7$ in the upper legs, and two MOSFET switches $Q_6$ and $Q_8$ in the lower legs. $D_6$ and $D_8$ are the antiparallel body diodes of $Q_6$ and $Q_8$, respectively. Secondary ac voltage $v_S$ and ac current $i_S$ are the inputs of the SAR circuit.
$V_O$ and $I_O$ are dc charging voltage and current for the battery, respectively.

### B. Operation of the SAR

The switching sequences and the operating waveforms of the SAR are shown in Fig. 3. MOSFET switches $Q_6$ and $Q_8$ are turned on during the on-time of their antiparallel diodes to have zero-voltage switching (ZVS). Both $Q_6$ and $Q_8$ are turned on for half a cycle, and they are complements of each other. Then, $Q_6$ is turned off with a time delay of $\pi - \theta \in [0, \pi]$ to the zero-cross point where $i_S$ commutates from negative to positive, while $Q_8$ is turned off with a time delay of $\pi - \theta \in [0, \pi]$ to the zero-cross point where $i_S$ commutates from positive to negative. Thus, the conduction angle $\theta$ of the SAR has a maximum $\pi$ and minimum 0. It is noted that the change of $\theta$ will affect the phase angle between $v_S$ and $i_S$. As shown in Fig. 3, $v_{S,1}$ is the fundamental component of $v_S$ that lags $i_S$ with a phase angle given by $\gamma = \frac{\pi \theta}{2}$. Therefore, the equivalent load is an impedance instead of the usual pure resistance [30], [31].

Since the battery charging is a slow process compared with the operating period of the SS IPT converter, the battery is modeled as a resistor determined by charging voltage and charging current, i.e., $R_L = \frac{V_O}{I_O}$. It has been studied that the SAR together with the resistive load can be represented by an equivalent fundamental impedance [32], [33], given by

$$Z_{eq} = R_{eq} + jX_{eq}$$

where

$$R_{eq} = \frac{8}{\pi^2} R_L \sin^4 \left( \frac{\theta}{2} \right)$$

and

$$X_{eq} = -\frac{8}{\pi^2} R_L \sin^3 \left( \frac{\theta}{2} \right) \cos \left( \frac{\theta}{2} \right)$$

are equivalent resistance and capacitive reactance, respectively.

### C. Operation of the SCC

The switching sequences and the operating waveforms of the SCC are shown in Fig. 4. Both $Q_a$ and $Q_b$ are turned on for half a cycle, and they are complements of each other. Then, $Q_a$ is turned on with a time delay of $\varphi \in \left[ \frac{\pi}{2}, \pi \right]$ to the zero-cross point where $i_S$ commutates from negative to positive, while $Q_b$ is turned on with a time delay of $\varphi \in \left[ 0, \frac{\pi}{2} \right]$ to the zero-cross point where $i_S$ commutates from positive to negative.

$\phi$ is varied from 0.5 to 9.0. It can be observed that $\phi$ increases with the increase of $\phi$. Since $Q_a$ and $Q_b$ are turned on at zero voltage, soft switching can be achieved to minimize the switching losses. The available charging time (or discharging time) for $C_2$ in half a cycle is $\pi - \varphi$, which decreases with the increase of $\varphi$ and results in a small equivalent root-mean-square (rms) value of $v_{S,CC}$. Consequently, the equivalent capacitance $C_{SCC}$ of the SCC can be modulated by varying the control angle $\varphi$. It has been studied that $C_{SCC}$ can be calculated by considering the fundamental components of $i_S$ and $v_{S,CC}$ [34], [35]. The capacitive reactance donated by $C_{SCC}$ is highlighted as

$$X_{C_{SCC}} = \left( 2 - \frac{2\varphi - \sin 2\varphi}{\pi} \right) X_{C_2}$$

where $X_{C_{SS}} = -\frac{1}{\omega C_{SS}}$ and $X_{C_2} = -\frac{1}{\omega C_2}$. The complex expression (4) can be simplified into (5) by using quadratic curve fitting. Fig. 5 shows the exact and approximate curves of $X_{C_{SCC}}$ versus the control angle $\varphi$. It can be observed that $X_{C_{SCC}}$ can be modulated from a nominal reactance $X_{C_2}$ to zero as $\varphi$ is varied from $0.5\pi$ to $\pi$.
Unless specified otherwise, in the rest of the article, $X_{\text{subscript}}$ represents the reactance of the corresponding component indicated by its subscript.

D. Equivalent Circuit Model

An equivalent circuit model of the proposed system using the fundamental approximation is shown in Fig. 6. The simplification is sufficiently accurate for resonant circuits operating near the resonant frequency [18]–[21]. Here, the equivalent circuit model of the proposed system is similar to that of a conventional SSIPPT converter, except that the secondary compensation capacitance is variable and the load is not purely resistive. The load is represented by an equivalent impedance $Z_{\text{eq}}$ with resistance $R_{\text{eq}}$ in series with reactance $X_{\text{eq}}$. Variables $V_P$, $I_P$, $V_S$, and $I_S$ are phasors of the fundamental components of $v_P$, $i_P$, $v_S$, and $i_S$, respectively. Resistor $R_P$ includes losses from the primary coil and the inverter, while resistor $R_S$ includes losses from the secondary coil, the SCC, and the SAR. Equivalent series resistance of the compensation capacitors can be ignored [18]–[21]. Detailed calculation of $R_P$ and $R_S$ will be given in Section IV-B for loss analysis.

As shown in Fig. 6, $C_1$, $C_{\text{SCC}}$, and $X_{\text{eq}}$ donate capacitive reactances in the secondary, and they can be represented by an equivalent secondary compensation capacitance $C_{S,\text{eq}}$, with its reactance satisfying

$$X_{C_{S,\text{eq}}} = -\frac{1}{\omega C_{S,\text{eq}}} = X_{C_1} + X_{C_{\text{SCC}}} + X_{\text{eq}}. \quad (6)$$

Therefore, the equations for the circuit model in Fig. 6 are

$$\begin{align*}
(R_P + jX_{L_P} + jX_{C_P})I_P - jX_M I_S &= V_P \\
-(R_S + R_{\text{eq}} + jX_{L_S} + jX_{C_{\text{eq}}})I_S + jX_M I_P &= 0
\end{align*} \quad (7)$$

where $X_M = \omega M$, $X_{L_P} = \omega L_P$, $X_{C_P} = -\frac{1}{\omega C_P}$, and $X_{L_S} = \omega L_S$. The magnitudes of $V_P$, $V_S$, and $I_S$ are given by

$$\begin{align*}
|V_P| &= \frac{4}{\pi} V_I \\
|V_S| &= \frac{4}{\pi} \sin\left(\frac{\theta}{2}\right) V_O \\
|I_S| &= \frac{\pi}{2} \frac{I_O}{\sin^2\left(\frac{\theta}{2}\right)}.
\end{align*} \quad (9)$$

III. CONTROL SCHEME FOR CP AND MAXIMUM-EFFICIENCY CHARGING

A. Maximum Efficiency

Using the circuit model, shown in Fig. 6, the efficiency can be calculated by

$$\eta = \frac{|I_S|^2 R_{\text{eq}}}{|I_S|^2 R_{\text{eq}} + |I_S|^2 R_S + |I_P|^2 R_P}$$

$$= \frac{X_M^2 R_{\text{eq}}}{(R_{\text{eq}} + R_S)^2 + (X_{L_S} + X_{C_{S,\text{eq}}})^2 R_P + X_M^2 (R_{\text{eq}} + R_S)} \quad (12)$$

Given a chosen operating frequency $\omega$, the efficiency in (12) can be maximized as

$$\eta_{\text{max}} \approx \frac{1}{2\sqrt{2\pi R_{\text{eq}}}} + 1, \quad (13)$$

$$X_{L_S} + X_{C_{S,\text{eq}}} = 0 \text{ and } (14)$$

$$R_{\text{eq, opt}} = X_M \sqrt{\frac{R_S}{R_P}}. \quad (15)$$

with the assumptions $\frac{X_M}{\sqrt{R_{\text{eq}} R_S}} \gg 1$ and $\frac{R_S}{R_P} \gg 1$. Variables $X_{C_{S,\text{eq}}, \text{opt}}$ and $R_{\text{eq, opt}}$ are the optimum values of $X_{C_{S,\text{eq}}}$ and $R_{\text{eq}}$ leading to maximum efficiency, respectively [29], [30], [36], [37].

From (15), the battery resistance $R_{\text{eq}}$ varying in a wide range, i.e., $R_{\text{eq}} \in [R_{L,\text{min}}, R_{L,\text{max}}]$, should be transformed into a matched load resistance $R_{\text{eq, opt}}$ by the SAR. With (2) and (15), the conduction angle $\theta$ of the SAR is given by

$$\theta = 2 \arcsin \left(\frac{X_{\text{eq}}}{\sqrt{\frac{R_{\text{eq, opt}}}{R_P}} R_L}\right) \quad (16)$$

However, from (3), the change of $\theta$ also affects the load reactance $X_{\text{eq}}$, given by

$$X_{\text{eq}} = -R_{\text{eq, opt}} \cot \left(\frac{\theta}{2}\right). \quad (17)$$

In Fig. 7(a), the solid curve labeled with $\theta/\pi$ shows the change of $\theta$ with regard to $R_L$ for optimum load resistance. Indicated by the dashed line, optimum load resistance $R_{\text{eq, opt}}$ can be achieved by controlling $\theta$ of the SAR. However, the magnitude of $X_{\text{eq}}$ concurrently becomes larger with the decrease of $\theta$, as shown in the solid curve labeled with $|X_{\text{eq}}|/R_{\text{eq, opt}}$. The simulation parameters are given in Table I and will be used for the rest of this article unless specified otherwise.

To ensure (14), $C_{S,\text{eq}}$ should fully compensate $L_S$ at the operating frequency that the equivalent capacitive reactance $X_{C_{S,\text{eq}}}$ should be constant at $X_{C_{S,\text{eq}}} = -X_{L_S}$. Therefore, the variation of $X_{C_{S,\text{eq}}}$ caused by $X_{\text{eq}}$ should be offset by $X_{C_{\text{SCC}}}$, With (5), (6), and (17), the control angle $\varphi$ of the SCC can be derived as

$$\varphi \approx \pi - \frac{\pi}{2} \sqrt{\frac{|X_{C_{S,\text{eq}}} - |X_{C_1} - |X_{\text{eq}}|}{|X_{C_2}|}}. \quad (18)$$
Fig. 7. (a) Conduction angle $\theta$, normalized $R_{\text{eq}}$, normalized $X_{\text{eq}}$ versus load resistance $R_L$. (b) Phase delay angle $\varphi$, normalized $X_{C_{\text{SCC}}}$ versus conduction angle $\theta$.

**TABLE I**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Symbols</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Self inductance</td>
<td>$L_P$, $L_S$</td>
<td>88 $\mu$H, 102 $\mu$H</td>
</tr>
<tr>
<td>Coupling coefficient</td>
<td>$k$</td>
<td>0.26</td>
</tr>
<tr>
<td>Coil resistance</td>
<td>$R_{P\text{-w}}, R_{S\text{-w}}$</td>
<td>0.3 $\Omega$, 0.328 $\Omega$</td>
</tr>
<tr>
<td>Inverter switch</td>
<td>$R_{\text{in}}$</td>
<td>0.1 $\Omega$</td>
</tr>
<tr>
<td>SCC switch</td>
<td>$R_{\text{on2}}, V_{I2}$</td>
<td>0.1 $\Omega$, 0.7 $V$</td>
</tr>
<tr>
<td>SAR switch</td>
<td>$R_{\text{on3}}, V_{I3}$</td>
<td>0.1 $\Omega$, 0.7 $V$</td>
</tr>
<tr>
<td>Operating frequency</td>
<td>$f_0$</td>
<td>85 kHz</td>
</tr>
<tr>
<td>Compensation capacitance</td>
<td>$C_P, C_1, C_2$</td>
<td>40.8 nF, 44 nF, 166 nF</td>
</tr>
<tr>
<td>Optimum load resistance</td>
<td>$R_{\text{eq,opt}}$</td>
<td>fixed at 18 $\Omega$</td>
</tr>
</tbody>
</table>

As shown in Fig. 7(b), with the coordinated control of $\varphi$ with respect to $\theta$ given in red curve, $X_{C_{\text{SCC}}}$ is almost constant at $X_{C_{\text{SCC}}} = 0$, achieving maximum efficiency given in magenta curve.

### B. CP Charging

Equations (14) and (15) can be satisfied via controlling the conduction angle $\theta$ of the SAR and the control angle $\varphi$ of the SCC, as discussed in Section III-A. As shown in Fig. 8, with the controlled optimum load resistance $R_{\text{eq,opt}}$ and null reactance in the secondary, i.e., $X_{L_S} + X_{C_{\text{SCC}}} = 0$, the maximum efficiency, given by (13), can be maintained over a wide range of battery resistance $R_L$. Besides, it has been widely studied that an SSIPPT system can achieve load-independent output current, if operating with a null reactance in the primary [20], [22], [23]. Theoretically, if component losses are neglected, the magnitude of the load-independent output current $i_S$ is highlighted as

$$|i_S| \approx \frac{|V_P|}{X_M}.$$  \hspace{1cm} (19)

Therefore, with the merits of load-independent transfer characteristic and load impedance matching, a novel operation approach for CP output and maximum efficiency is proposed, as illustrated in Fig. 8. The magnitude of the output current is constant at $|I_S|$ due to the native load-independent current transfer characteristic [20], [22], [23], while the matching load for maximum efficiency is maintained constant at $R_{\text{eq,opt}}$ via control. Therefore, given an input voltage, the proposed system outputs a constant output power at maximum efficiency, given by

$$P_{O,\text{constant}} \approx |I_S|^2 \text{rms} R_{\text{eq,opt}}$$  \hspace{1cm} (20)

where subscript rms represents the calculation of rms value of the corresponding variable.

With (9)–(11), (19), and (20), the charging power, dc charging voltage, and dc charging current can be, respectively, designed with

$$P_{O,\text{constant}} \approx \frac{8}{\pi^2} \frac{V_I^2}{\omega M} \sqrt{R_S R_P}$$  \hspace{1cm} (21)

$$V_O = \frac{V_I}{\sin^2 \left( \frac{\theta_0 - \varphi}{2} \right)} \sqrt{R_S R_P}$$  \hspace{1cm} (22)

$$I_O = \frac{8}{\pi^2} \frac{V_I}{\omega M} \sin^2 \left( \frac{\theta_0 - \varphi}{2} \right)$$  \hspace{1cm} (23)

### C. Secondary Impedance Control Schemes

From (2), the equivalent load resistance $R_{\text{eq}}$ can be modulated to the optimum value $R_{\text{eq,opt}}$ by controlling the conduction angle $\theta$. Assuming that the equivalent load impedance $X_{\text{eq}}$ can be offset by the proper control of SCC impedance $X_{C_{\text{SCC}}}$, the output power $P_{O}$ is solely determined by the equivalent load resistance $R_{\text{eq}}$, given by $P_{O} \approx |I_S|^2 \text{rms} R_{\text{eq}}$. With (2), $P_{O}$ takes a monotonic relationship with the control variable $\theta$. As an illustration, monotonic curves of $P_{O}$ versus $\theta$ for various values of battery resistance $R_L$ are shown in Fig. 9. It can be observed that, when $P_{O}$ is constant at $P_{O,\text{constant}}$ in (20), the proposed system operates at its optimized efficiency. Therefore, a simple...
Fig. 9. Output power $P_O$ and efficiency $\eta$ versus conduction angle $\theta$ under different values of battery resistance $R_L$.

Fig. 10. Secondary impedance control diagram of the proposed CP charging system.

PI controller can be used to achieve constant output power and maintain maximum efficiency, with $P_{O,\text{constant}}$ in (20) being a control reference.

Fig. 10 shows the control diagram in practical implementation. Since the operating frequency in the primary is fixed, and only impedance control in the secondary is needed for CP output and maximum efficiency, wireless feedback communication between the primary and the secondary can be eliminated. The charging voltage $V_O$ and the charging current $I_O$ are measured by sensors. $P_O$ and $R_L$ can be calculated by a multiplier and a divider, respectively. A simple PI controller applies the correction to the difference between $P_O$ and $P_{O,\text{ref}}$, and forms a control signal $\theta$ for the SAR. Meanwhile, with $\theta$ and $R_L$, another control signal $\varphi$ is generated according to the analytical relationship given in (18), or alternatively, the measured relationship, such as Fig. 14. If there is no misalignment issue, both results are fine for the control. However, if there is occurrence of misalignment, (18) is more generally applicable. Zero-crossing detection of $i_S$ generates a synchronization signal for the pulsewidth modulation (PWM) generations. Angles $\varphi$ and $\theta$ are used to produce PWM driving signals for the SCC and SAR, respectively.

IV. DESIGN CONSIDERATIONS

A. Minimizing Voltage Stress of the SCC

As analyzed in Sections II-B and II-C, $|X_{\text{eq}}|$ ranges from $|X_{\text{eq}}|_{\text{min}}$ to $|X_{\text{eq}}|_{\text{max}}$ depending on the battery resistance $R_L$, and $|X_{C_{\text{S,eq}}}|$ ranges from zero to $|X_{C_2}|$ with the control angle $\varphi$ varying from $\pi$ to $0.5\pi$. Subscripts “max” and “min” represent the maximum and minimum values of corresponding variables, respectively. $X_{C_{\text{S,eq}}}$ can be controlled to offset the variation of $X_{\text{eq}}$, and thus, $X_{C_{\text{S,eq}}}$ can be constant at $X_{C_{\text{S,eq,opt}}}$ to fully compensate $X_{L_S}$. From (14), design of $C_1$ should first ensure the requirements of full compensation in the secondary, given as

\[ |X_{C_1}| + |X_{C_2}| + |X_{\text{eq}}|_{\text{min}} \geq |X_{C_{\text{S,eq,opt}}}| \quad \text{and} \quad (24) \]

\[ |X_{C_1}| + |X_{\text{eq}}|_{\text{max}} \leq |X_{C_{\text{S,eq,opt}}}| \quad \text{and} \quad (25) \]

The voltage stress of the SCC switches is determined by the maximum voltage across the SCC, as given by

\[ |V_{SCC,\text{max}}| = |X_{C_2}| |I_S| \quad \text{or} \quad |V_{SCC,\text{max}}| = |X_{C_2}||I_S| \quad \text{if} \quad \varphi = \pi \quad \text{and} \quad |X_{C_1}| = |X_{C_1}|_{\text{max}} \quad \text{is suggested.} \]

\[ |X_{C_1}|_{\text{max}} = |X_{C_{\text{S,eq,opt}}}| - |X_{\text{eq}}|_{\text{max}} \quad \text{and} \quad (27) \]

The black curve in Fig. 11 shows the relationship between the voltage stress $|V_{SCC,\text{max}}|$ and the design value of $|X_{C_1}|$. It can be observed that, the voltage stress is significantly reduced by designing a large $|X_{C_1}|$. Specifically, $|X_{C_1}| = |X_{C_1}|_{\text{max}}$ is suggested.

Moreover, when the control angle is maximum, i.e., $\varphi = \pi$, $C_{S2,\text{eq}}$ is shortened by the switches $Q_0$ and $Q_5$, and thus, maximum current stress of the SCC switches happens. Since $i_S$ is constant as (19), maximum current stress of the SCC switches is
given by
\[ |I_{SCC,\text{max}}| = |I_S| \]  
(28)

**B. Loss Analysis**

Theoretically, the proposed system operates with zero phase angle between the input voltage \(v_P\) and the input current \(i_P\). In practice, the input impedance can be designed to be slightly inductive to facilitate ZVS of the MOSFET switches \(Q_1-Q_4\). For switching loss reduction, a slight decrement of \(\omega_P\) can fulfill the requirement and will not affect the output and the efficiency too much [22], [23]. Therefore, \(R_P\), representing the primary-side losses, can be estimated by considering the primary-side coil resistance and the conduction losses of the inverter switches, given by
\[ R_P = R_{\text{on,1}} + 2R_{\text{on,1}} \]  
(29)
where \(R_{\text{on,1}}\) is the conduction resistance of the inverter switches \(Q_1-Q_4\). \(R_P\) can be considered as constant.

As analyzed in Section II-C, the SCC switches \(Q_a\) and \(Q_b\) are both soft switched. The conduction loss in the SCC switches can be estimated as
\[ P_{SCC} = I_{SCC,\text{rms}}^2 R_{\text{on,2}} + I_{SCC,\text{avg}} V_{i,2} \]  
(30)
where \(R_{\text{on,2}}\) and \(V_{i,2}\) are the ON-resistance and body-diode forward voltage of the MOSFET switches \(Q_a-Q_b\), respectively. \(I_{SCC,\text{rms}}\) and \(I_{SCC,\text{avg}}\) are the rms value and average value of the current flowing through the SCC switches \(Q_a-Q_b\), given, respectively, as given by
\[ I_{SCC,\text{rms}} = \sqrt{\frac{1}{\pi} \int_{\pi-\phi}^{\phi} (|I_S| \sin x)^2 \, dx} \]  
(31)
\[ I_{SCC,\text{avg}} = \frac{1}{\pi} \int_{\pi-\phi}^{\phi} |I_S| \sin x \, dx. \]  
(32)

Similarly, neglecting the small switching-off losses of \(Q_6\) and \(Q_8\), the conduction loss in the SAR can be estimated as
\[ P_{SAR} = I_{S,\text{rms}}^2 R_{\text{on,3}} + i_{S,\text{avg}} V_{i,3} \]  
(33)
where \(R_{\text{on,3}}\) is the ON-resistance of the MOSFET switches \(Q_6\) and \(Q_8\), and \(V_{i,3}\) is the forward voltage of the diodes \(D_5-D_8\). \(i_{S,\text{rms}} = \frac{|I_S|}{\sqrt{2}}\) and \(I_{S,\text{avg}} = \frac{2|I_S|}{\pi}\) are the rms value and average value of \(i_S\) injecting into the SAR, respectively.

Incorporating the losses in the SCC and the SAR, equivalent series resistance \(R_S\) representing the losses in the secondary-side can be calculated as
\[ R_S = R_{S,\text{sw}} + P_{SCC} + P_{SAR} \]  
(34)
\[ \frac{R_S}{R_{\text{eq}}^2} \]  
\[ \frac{R_S}{R_{\text{eq}}} \]

Loss resistance ratio \(\sqrt{\frac{R_S}{R_{\text{eq}}}}\) is simulated, as shown by the blue curve in Fig. 12. Since the loss of the SCC increases with the increase of the control angle \(\varphi\), \(\sqrt{\frac{R_S}{R_{\text{eq}}}}\) varies from 1.1 to 1.3 with respect to the battery resistance \(R_L\). Theoretically, the optimum load resistance \(R_{\text{eq,opt}}\) should vary with the variation of \(\sqrt{\frac{R_S}{R_{\text{eq}}}}\). In practical, a slight deviation from the optimum load resistance will not affect the efficiency too much, and thus, \(R_{\text{eq,opt}}\) can be fixed for simplicity. The simulated efficiency shown by the green curve in Fig. 12 slightly decreases mainly due to the increase of \(R_S\). Nevertheless, the efficiency is approximately maintained maximum over the whole load range.

**V. EXPERIMENTAL VERIFICATION**

**A. Specifications and Prototype**

To verify the CP output and maximum-efficiency performance throughout the charging process, an experimental prototype is built, as shown in Fig. 13. The system parameters are given in Table II. According to the charging specifications, the equivalent battery resistance approximately ranges from 18 to 50 \(\Omega\). An electronic load is used to emulate the battery. The input dc power and output dc power are measured by a Yokogawa WT1800 Precision Power Scope.

**B. Measured Operating Points, Output Power, and Efficiency**

The operating frequency of the inverter is fixed at 85 kHz. Following the proposed operation approach in Section III, the conduction angle \(\theta\) of the SAR and the control angle \(\varphi\) of the SCC are adjusted to achieve CP output and maintain maximum efficiency. The measured operating points (marked with “○”) are shown in Fig. 14, with \(\varphi\) and \(\theta\) varying from 0.53 to 0.83\(\pi\) and from 0.95 to 0.57\(\pi\), respectively.
TABLE II  
SYSTEM PARAMETERS

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Symbols</th>
<th>Measured Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Battery Specifications</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Rated charging power $P_O$</td>
<td>$V_I$</td>
<td>48 V</td>
</tr>
<tr>
<td>Battery terminal voltage $V_O$</td>
<td>$Q_1$–$Q_4$, $Q_5$, $Q_6$, $Q_7$, $V_F$, $V_P$, $R_m$, $R_{in}$</td>
<td></td>
</tr>
<tr>
<td>Switches</td>
<td>$D_S$, $D_T$</td>
<td>MBR20200 with $V_F$, $V_P$, $R_m$, $R_{in}$</td>
</tr>
<tr>
<td>Diodes</td>
<td>$L_{P}$, $L_{S}$</td>
<td>85.09 μH, 101.13 μH</td>
</tr>
<tr>
<td>Coupling coefficient $k$</td>
<td>0.262</td>
<td></td>
</tr>
<tr>
<td>Coil resistance</td>
<td>$R_{P}$, $R_{S}$</td>
<td>0.38 Ω, 0.41 Ω</td>
</tr>
<tr>
<td>Primary compensation $C_P$</td>
<td>41 nF</td>
<td></td>
</tr>
<tr>
<td>Secondary compensation $C_1$, $C_2$</td>
<td>155 nF, 55 nF</td>
<td></td>
</tr>
<tr>
<td>Operating frequency $f$</td>
<td>$\frac{f}{D}$</td>
<td>85 kHz</td>
</tr>
</tbody>
</table>

The corresponding charging current (marked with “□”), as shown in Fig. 15(a). The corresponding output power (marked with “○”), as shown in Fig. 15(a), which is consistent with the analysis in Section IV-B. To sum up, the proposed operation approach can ensure CP charging and maximum efficiency throughout the charging process.

Waveforms of the inverter, the SCC, and the SAR are measured at the start, middle, and end of CP charging, as shown in Fig. 16. ZVS is achievable in the inverter, SCC, and SAR. The maximum voltage stress of the SCC switches is about 55 V, as shown in Fig. 16(a), which coincides with the analysis in Section IV-A.

C. Transient Response Against Load Change

The closed-loop secondary impedance control scheme demonstrated in Section III-C is implemented in a microcontroller for CP charging and maximum efficiency throughout the charging process. Transient waveforms for step load changing are shown in Fig. 17. The output voltage $V_O$ and output current $I_O$ are measured and shown as CH7 in blue and CH8 in green, respectively. The control variables are observed from digital-to-analog outputs, where CH9 in red and CH11 in magenta represent the conduction angle $\theta$ of the SAR and the control angle $\varphi$ of the SCC, respectively. $P_O$ is calculated by multiplying $V_O$ and $I_O$ and shown as the waveform in yellow. It can be observed that $P_O$ is tightly regulated by direct control of $\theta$, while $\varphi$ is coordinately controlled with the variation of $\theta$ and $R_L$. No wireless communication is needed for the control of the proposed system.

D. Discussion on Misalignment Issue

For stationary IPT charging applications, the coupling coefficient is commonly fixed once the positioning process is finished, and it will rarely fluctuate during the charging process [27], [30], [32]. Although misalignment may occur due to low-precision positioning, variation of the coupling coefficient $k$ is normally restricted to a small range with a proper design of the loosely coupled transformer [38]. The small variation of $k$ slightly affects the magnitude of the load-independent output current $I_S$, given in (28), but the output power $P_O$ given by $P_O \approx \frac{I_S^2}{R_{eq}}$, can be directly regulated to be constant by controlling the equivalent load resistance $R_{eq}$ via the conduction angle $\theta$ of the SAR, as shown in Figs. 9 and 10. Moreover, with (16)–(18), the control angle $\varphi$ of the SCC is regardless of the variation of $k$ and responsible for an optimum null reactance.
Fig. 16. Measured operating waveforms of the inverter, the SCC, and the SAR at the (a) start $R_L = 18 \Omega$, (b) middle $R_L = 30 \Omega$, and (c) end $R_L = 50 \Omega$ of CP charging.

Fig. 17. Transient waveforms for $R_L$ step changing from 20 to 40 $\Omega$ and back to 20 $\Omega$.

in the secondary side, as required by (14). Therefore, even if there exists a misalignment issue, our proposed control can still ensure CP output. It should be noted that, for abnormal seriously misaligned conditions where $k$ is extremely small, excessively large input power is required to provide the nominal CP output, and thus, overcurrent protection in the primary side should be implemented to prevent damages to the circuit.

A validation of the proposed method is conducted under aligned condition ($k = 0.262$) and misaligned condition ($k = 0.23$). An identical output power of 147 W can be ensured even if there is a misalignment, as shown in Fig. 18(a). The slight efficiency degradation is mainly brought by the reduction of $k$ in misaligned condition. The corresponding relationships between $\theta$ and $\phi$ are given in Fig. 18(b).

VI. CONCLUSION

A single-stage IPT converter, which can operate as a wireless CP battery charger and maintain maximum efficiency throughout the charging process, is proposed in this article. A novel operation approach combines the merits of load-independent transfer characteristic and load impedance matching, by controlling the control angle of the SCC and the conduction angle of the SAR. The operating frequency of the IPT converter is fixed, and only a simple control in the secondary side
is employed to achieve CP output and to ensure load matching for the maximum efficiency. No wireless feedback communication is needed for the control, and all power switches realize ZVS.

REFERENCES

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