Evaluation of losses in a secondary-side controlled wireless battery charging system

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Abstract: Secondary-side controlled inductive power transfer (IPT) systems are easy to control compared to primary and dual-side control methods, and more importantly, they do not require wireless communication between primary and secondary sides. In this paper, a wireless battery charging system based on secondary-side controlled IPT is analyzed and presented in detail. The proposed control method relies on the detection of the secondary coil current and the transferred power can be fully controlled by the secondary-side active rectifier with reduced number of power switching devices. The steady-state performance of the IPT system is evaluated using a fundamental harmonic analysis method for resistive and reactive control modes of the active rectifier by taking the converter losses into account. Theoretical and simulation results of the designed system are validated by experimental results obtained from the implemented laboratory prototype for charging of a 96-V lead-acid battery.

Key words: Inductive power transfer, secondary-side control, active rectifier, impedance matching

1. Introduction

The attractive features of wireless power transfer systems have found growing interest from researchers in many applications at various power levels. Among them, the inductive power transfer (IPT) method has been extensively studied for battery charging of electric vehicles, which require charging power at kW levels. In this method, power is transferred through an air-gap between two magnetically coupled coils, which are resonated by series or parallel capacitors on each side, allowing four possible combinations of compensation scheme. The series-series compensation topology provides better performance since it is insensitive to the load and coupling coefficient variations. The loosely coupled coils are operated by power-electronic converters at resonance frequencies in the kHz range to achieve high efficiency values. Although the system efficiency is better at higher frequencies, it is impractical to operate the converters at MHz level due to high stress and losses on switching devices [1–3].

The efficiency of an IPT system is dependent on loading conditions on its secondary side [4, 5]. In a battery charging system, the loading condition varies with time from battery floating to the full output power. Hence, the control of power transferred to the battery is the main concern as well as its efficient transfer over a wide range [6]. The transferred power can be controlled on only one side (primary/secondary-side control) or on both (dual-side control). The primary-side control adopts frequency, duty cycle, or phase shifting methods for power transfer and requires wireless communication with its secondary side [2, 6]. In dual-side control, battery
charge power can be controlled by adjusting the voltage levels of the converters on both sides to improve the power transfer efficiency. This is accomplished by operating the converters in coordination by means of a wireless communication between them [6–8]. On the other hand, secondary-side control, adopting duty cycle and phase shifting methods, is usually employed in applications where the transferred power is regulated by the secondary-side controller without requiring communication with the primary [9–15]. Hence, it requires less control complexity when compared to primary and dual-side control methods.

The battery charge current control on the secondary side relies on the adjustment of the effective impedance seen by the secondary coil. Typical power circuit configurations are buck, boost, and buck-boost DC-DC converter topologies cascaded to a diode bridge rectifier [5, 6, 9]. Alternatively, semibridgeless or full-bridge active rectifier topologies [7, 8, 10–15] replacing the diode bridge rectifier are also employed. Several control methods have been proposed in the literature for secondary-side control to adjust the secondary load impedance based on combined [10, 11] and independent control [13–15] of its resistive and reactive parts. These studies mainly focused on a specific control method; however, the efficiency of the system depends on the chosen power converter topology and control complexity.

In this paper, an IPT system based secondary-side control with minimum hardware and control requirements is aimed. For this purpose, the semibridgeless active rectifier topology [7, 10, 11] is used on the secondary side for a battery charging application. As an extension of our previous work [16], the losses in the system including the converter losses are modeled in detail using the fundamental harmonic approach. The coil efficiency is evaluated for the effective secondary impedance control method by rectifier conduction angle with and without tuning its reactive part. The effect of the conduction angle control on the subsystem efficiencies is analyzed. The theoretical results are verified experimentally on the IPT charger prototype designed for a 96-V lead-acid battery. Moreover, a simple control method relying only on detection of the secondary coil current is employed to reduce the control system complexity, unlike the counterparts proposed in the literature.

2. System overview

The circuit scheme of the series-series compensated IPT system is shown in Figure 1. It consists of two resonated coils with a full-bridge inverter on the primary side and a controlled rectifier on the secondary side. While the inverter provides a symmetrical square-wave output to the primary coil at an angular frequency of $\omega = 2\pi f$, the controlled rectifier at the secondary is used to adjust the battery charge current.

The steady-state circuit model of the system [3, 4, 14–17] is shown in Figure 2. In this model, a fundamental harmonic analysis method is used; therefore, the primary-side inverter is characterized by an ideal sinusoidal voltage source at the fundamental frequency. The coils are represented by a loosely coupled transformer model and the secondary-side rectifier is represented by equivalent AC input impedance of $Z_L = R_L + jX_L$. The solution of the circuit in Figure 2 yields the following primary and secondary-side voltage and current expressions:

$$V_1 = Z_{in}I_1 = \frac{Z_1Z_2 + \omega^2M^2}{Z_2}I_1,$$  \hspace{1cm} (1)

$$V_2 = Z_LI_2 = \frac{j\omega M(R_L + jX_L)}{Z_2}I_1,$$  \hspace{1cm} (2)
The efficiency of the resonated coils can be obtained from the ratio of output and input real powers as follows \[ \eta = \frac{\text{Re}\{Z_L|I_2|^2\}}{\text{Re}\{Z_{in}|I_1|^2\}} = \frac{\omega^2 M^2 R_L}{(R_2 + R_L)^2 R_1 + \omega^2 M^2 (R_2 + R_L) + R_1 X_2^2}, \] where \( X_2 = \omega L_2 - \frac{1}{\omega C_2} + X_L \) is the imaginary part of secondary impedance \( Z_2 \). Since \( Z_1 = R_2 \) and \( Z_2 = R_2 + R_L \) at resonance, the system efficiency under this condition becomes \[ \eta = \frac{\omega^2 M^2 R_L}{(R_2 + R_L)^2 R_1 + \omega^2 M^2 (R_2 + R_L)}. \]
Eqs. (5)–(8) show the dependency of the voltage gain, current gain, and efficiency on the load impedance. Eq. (7) implies that a proper reactive compensation is required on the secondary side in order to improve the efficiency of the system. Although the reactive part of the primary series reactance has no direct effect on the efficiency, its proper tuning affects the power transfer capability of the system according to Eqs. (3) and (4). Also, in order to achieve maximum efficiency, the load resistor should be closely matched to its optimal value, which is obtained by setting the derivative of Eq. (8) to zero as [14]:

$$R_{L,\text{opt}} = \sqrt{R_2^2 + \frac{R_2^2}{R_1^2} \omega^2 M^2}. \quad (9)$$

Since $\omega M \gg R_2$, the optimal load resistor value is close to the mutual reactance value, i.e. $R_{L,\text{opt}} \approx \omega M$, as mentioned in [4, 18] if both coils are identical.

The secondary-side controlled active rectifier adjusts the conduction angle ($\beta$) of the upper side diodes by controlling the gate pulses of the lower side switches. The operating modes of the circuit were depicted in [10, 11, 16]. The circuit can be controlled in purely resistive or reactive modes in order to adjust the effective impedance seen across the secondary terminals by properly providing switching instants as shown in Figure 3, where the secondary-side voltage and current waveforms are illustrated for each mode [16].

For purely resistive load conditions, the fundamental components of the secondary voltage and current are in phase. Therefore, from the ratio of their fundamental peak values the resistive load characteristic in terms of conduction angle is derived as follows [16]:

$$R_L = \frac{V_2}{I_2} = \frac{8}{\pi} \sin^2(\frac{\beta}{2}) R_{dc}. \quad (10)$$

For reactive load conditions, a phase shift is introduced between the fundamental components of the secondary current and voltage waveforms. Therefore, both the real and reactive parts of the load impedance are controlled together, and its expression is given as follows [10, 11, 16]:

$$Z_L = \frac{4}{\pi^2} R_{dc} (1 - \cos(\beta)) \sin(\frac{\beta}{2}) e^{j(\frac{\pi}{2} - \frac{\beta}{2})}, \quad (11)$$

where $0 \leq \beta \leq \pi$ for inductive load and $-\pi \leq \beta \leq 0$ for capacitive load.

![Figure 3](image)

**Figure 3.** Secondary voltage and current waveforms for (a) resistive, (b) inductive, and (c) capacitive loading conditions.

### 3. Evaluation of the system losses

The major losses affecting the system’s efficiency are the primary inverter losses, coil losses, and secondary rectifier losses. The calculation methods for these losses are given below in detail.
The losses in the resonated coils are represented by resistors \( R_1 \) and \( R_2 \) in Figure 2. These resistors primarily represent the coil resistors; however, the equivalent series resistance values of the resonance capacitor can be added to them. The coil losses are calculated from the following:

\[
P_{\text{coil}} = R_1 I_{1,\text{rms}}^2 + R_2 I_{2,\text{rms}}^2. \tag{12}
\]

The losses of the converters mainly consist of conduction and switching losses. The on-state conduction losses for an IGBT and a diode can be calculated using the following equations, respectively:

\[
P_{\text{cond,S}} = \frac{1}{T} \int_0^T V_{ce}(t) I_c(t) dt, \tag{13}
\]

\[
P_{\text{cond,D}} = \frac{1}{T} \int_0^T V_f(t) I_f(t) dt, \tag{14}
\]

where \( T = 1/f \) is the switching period of the inverter, \( V_{ce} \) is the IGBT collector-emitter voltage drop, \( I_c \) is the IGBT collector current, \( V_f \) is the diode forward voltage drop, and \( I_f \) is the diode current. At a specific temperature, the nonlinear voltage versus current characteristics of each device can be considered to be piecewise linear for simplification. Hence, the conduction losses of the IGBT and diode can be derived as [19, 20]:

\[
P_{\text{cond,S}} = V_{ce} I_{c,avg} + R_{ce} I_{c,\text{rms}}^2, \tag{15}
\]

\[
P_{\text{cond,D}} = V_f I_{f,avg} + R_f I_{f,\text{rms}}^2, \tag{16}
\]

where \( V_{ce0} \) and \( V_{f0} \) are the threshold voltages and \( R_{ce} \) and \( R_f \) are the dynamic resistances of the IGBT and diode, respectively.

The primary-side inverter provides a square-wave input voltage to the transmitter coil and operates in resonant mode. If the inverter supplies an inductive load at its terminals, all switches operate in zero-voltage switching mode, since their antiparallel diodes conduct before they begin to conduct. In that case, it can be assumed that the switching losses are negligible and only conduction losses are taken into account.

The current and voltage waveforms of the primary-side inverter switches are the same for each one and their conduction periods are illustrated in Figure 4. For sinusoidal inverter output current at a specific power factor, average and rms value of the primary-side current of the switches can be derived as follows:

\[
I_{D,\text{avg}} = \frac{I_{1,\text{peak}}}{2\pi} [1 - \cos(\alpha)], \tag{17}
\]

\[
I_{D,\text{rms}} = \sqrt{\frac{I_{1,\text{peak}}}{2\pi} \left[ \frac{\alpha}{2} - \frac{\sin(2\alpha)}{4} \right]}, \tag{18}
\]

\[
I_{S,\text{avg}} = \frac{I_{1,\text{peak}}}{2\pi} [1 + \cos(\alpha)], \tag{19}
\]

\[
I_{S,\text{rms}} = \sqrt{\frac{I_{1,\text{peak}}}{2\pi} \left[ \frac{\pi - \alpha}{2} + \frac{\sin(2\alpha)}{4} \right]}, \tag{20}
\]
where $\alpha$ is the angle between the fundamental inverter output voltage and current.

The losses on the secondary-side converter are due to conduction and switching of the IGBT and the body diodes depending on the conduction angle value. The conduction periods of each semiconductor device are illustrated in Figure 5. In the ideal commutation scheme, $S_7-S_8$ switches conduct for the $\pi - \beta$ angle, while $D_5-D_6$ diodes conduct for the $\beta$ angle throughout each half cycle and $D_7-D_8$ diodes conduct for each complete half-cycle.

In order to calculate the conduction losses from Eqs. (15) and (16), the average and rms values of each diode and IGBT can be derived as follows:

\[
I_{D5,\text{avg}} = I_{D6,\text{avg}} = \frac{I_{2,\text{peak}}}{\pi} \sin\left(\frac{\beta}{2}\right),
\]

\[
I_{D5,\text{rms}} = I_{D6,\text{rms}} = \frac{I_{2,\text{peak}}}{2} \sqrt{\frac{\beta + \sin(\beta)}{\pi}},
\]

\[
I_{D7,\text{avg}} = I_{D8,\text{avg}} = \frac{I_{2,\text{peak}}}{\pi},
\]

\[
I_{D7,\text{rms}} = I_{D8,\text{rms}} = \frac{I_{2,\text{peak}}}{2},
\]

\[
I_{S7,\text{avg}} = I_{S8,\text{avg}} = \frac{I_{2,\text{peak}}}{\pi} \left[1 - \sin\left(\frac{\beta}{2}\right]\right],
\]

\[
I_{S7,\text{rms}} = I_{S8,\text{rms}} = \sqrt{\frac{I_{2,\text{peak}}}{\pi} \left(\frac{\pi - \beta}{4} - \frac{\sin(\beta)}{4}\right)}.
\]

The secondary-side switches do not operate with soft switching in resistive control because the switching is done at the time when both current and voltage are not zero. Therefore, the switching strategy is based on hard switching. Consequently, switching loss calculation is made on normalizing turn-on and turn-off losses.
including the reverse recovery losses of the diodes given in the data sheet depending on the switching instants and DC bus voltage level as follows [20]:

\[ P_{SW,S7} = P_{SW,S8} = \frac{f(E_{on} + E_{off})I_{2,peak} \sin\left(\frac{\pi + \beta}{2}\right)V_{out}}{V_{nom}I_{nom}}, \quad (27) \]

where \( E_{on} \) and \( E_{off} \) are the turn-on and turn-off switching losses of the IGBT obtained at the nominal voltage \( (V_{nom}) \) and current \( (I_{nom}) \) values given in the data sheet.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( N_1, N_2 )</td>
<td>Spiral coil number of turns</td>
<td>24 turns</td>
</tr>
<tr>
<td>( h )</td>
<td>Distance between coils</td>
<td>75 mm</td>
</tr>
<tr>
<td>( D_{out} )</td>
<td>Outer diameter of each coil</td>
<td>400 mm</td>
</tr>
<tr>
<td>( k )</td>
<td>Coupling coefficient</td>
<td>0.45</td>
</tr>
<tr>
<td>( L_1, L_2 )</td>
<td>Self-inductance of IPT coils</td>
<td>140 ( \mu )H</td>
</tr>
<tr>
<td>( C_1, C_2 )</td>
<td>Resonant capacitor</td>
<td>470 nH</td>
</tr>
<tr>
<td>( R_1, R_2 )</td>
<td>Coil resistors</td>
<td>0.3 ( \Omega )</td>
</tr>
<tr>
<td>( C_{dc} )</td>
<td>Secondary dc bus capacitor</td>
<td>470 ( \mu )F</td>
</tr>
<tr>
<td>( f )</td>
<td>Operating frequency</td>
<td>20.6 kHz</td>
</tr>
<tr>
<td>( V_{in} )</td>
<td>DC input voltage</td>
<td>75 V</td>
</tr>
<tr>
<td>( V_{bat} )</td>
<td>Nominal battery voltage</td>
<td>96 V</td>
</tr>
</tbody>
</table>

The steady-state analyses of the system were performed at a constant battery voltage level. The designed coils in [21] are used in this study by setting the vertical distance between them to 75 mm and the system parameters are given in Table 1 [16]. First, the performance of resonated coils is investigated by neglecting the losses of inverter and controlled rectifier to simplify the analysis. The equivalent resistor \( R_L \) at the input terminal of the controlled rectifier can be adjusted by the conduction angle \( \beta \) according to Eq. (10) depending on loading conditions. Assuming that the converters are ideal, for the given \( V_{in}, V_{out} \), and \( \beta \) values, the voltage gain of the coils is constant according to Eq. (5) and can be calculated as follows:

\[ V_{1,peak} = \frac{4}{\pi}V_{in}, \quad (28) \]

\[ V_{2,peak} = \frac{4}{\pi}V_{out} \sin\left(\frac{\beta}{2}\right), \quad (29) \]

\[ M_V = \left| \frac{V_{2,peak}}{V_{1,peak}} \right| = \frac{V_{out}}{V_{in}} \sin\left(\frac{\beta}{2}\right). \quad (30) \]

The analyses are conducted on the equivalent circuit shown in Figure 2. Hence, for the given conditions, the only unknown in Eq. (5) is \( Z_L \), which is equal to \( R_L \) given in Eq. (10) for the resistive control mode. Unfortunately, determining the \( R_L \) value is not an easy task since it results in 4th order polynomial equations to be solved. Only the positive real-valued root of \( R_L \) gives the feasible solution. Once its value is determined, the circuit in Figure 2 can be solved easily. The same analysis can be applied to the reactive control mode of rectifier. This
time, the equivalent impedance \( Z_L \) at the input terminal of the controlled rectifier is changed according to Eq. (11). Since the impedance angle is determined only by the conduction angle, the real and reactive parts are controlled together. Hence,

\[
Z_L = R_L + jX_L = R_L(1 + j\tan(\frac{\pi - \beta}{2})).
\]  

(31)

Therefore, using Eq. (31), \( R_L \) and \( X_L \) can now be obtained from Eq. (5) by numerical analysis. The load impedance values are substituted into the circuit in Figure 2, and then it can be solved for different values of \( \beta \).

Figure 6 shows the variation of coil efficiency for resistive and reactive control modes of the rectifier assuming that the battery voltage level remains constant at its nominal value. The primary inverter is operated slightly above the resonance frequency and the converter losses are neglected. As can be seen, the coil efficiency remains high for a wide range of conduction angles in resistive control when compared to reactive control. In the case of reactive power control, the efficiency drops continuously towards light load conditions as \( \beta \) decreases. A comparison of efficiency values depending on the battery power level is shown in Figure 6 for both control modes. The characteristic of the reactive control is not preferable for battery charging, because the operating power of the battery charging system varies in a wide range from nearly zero to full power, especially in constant voltage mode. However, the pure resistive control will provide better efficiency performance and controllability for a wide range of battery charge current requirements. Therefore, it is preferred for the control of the experimental prototype.

In order to account for the converter losses, the model parameters of the semiconductor devices, which are given in Eqs. (15), (16), and (27), are determined from the data sheet information and summarized in Table 2. In this case, an analytical solution is not possible due to nonlinear converter characteristics. Hence, an iterative solution is adopted for given \( V_{in} \), \( V_{out} \), and \( \beta \) values and with an initial estimate of \( I_{out} \). The solution is based on the calculation of inverter input voltage by calculating the losses and electrical quantities at every stage of the circuit until the \( V_{in} \) value converges to its given value. Figure 7 shows the variation of subsystem and overall efficiency values for resistive control mode of the rectifier at various conduction angle values. The

![Image](image.png)

**Figure 6.** The effect of (a) rectifier conduction angle and (b) battery output power on the coil efficiency, neglecting converter losses for reactive (red) and resistive (blue) control.
Table 2. Converter model parameters.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{ceo}$</td>
<td>Forward IGBT voltage</td>
<td>1.05 V</td>
</tr>
<tr>
<td>$R_{ce}$</td>
<td>IGBT on state resistance</td>
<td>0.0417 Ω</td>
</tr>
<tr>
<td>$V_{fo}$</td>
<td>Forward diode voltage</td>
<td>1.9 V</td>
</tr>
<tr>
<td>$R_{f}$</td>
<td>Diode on state resistance</td>
<td>0.06 Ω</td>
</tr>
</tbody>
</table>

Figure 7. The effect of rectifier conduction angle on the subsystem efficiencies under resistive control mode.

The overall system efficiency is affected mostly by the rectifier efficiency at low values of conduction angle since it reduces considerably.

The charge current of the battery is adjusted by varying the conduction angle $\beta$ of the secondary-side active rectifier, which is designed to be controlled in purely resistive mode. Hence, the effective load impedance across the secondary terminals is equal to the resistance value as expressed in Eq. (10) while its reactive part is set to zero.

4. Implementation of current control

The control scheme of the proposed IPT system is shown in Figure 8, where the battery current is regulated by the conduction angle $\beta$, which is generated by a PI controller. The scheme in Figure 8 illustrates the charge current control of the battery in constant current mode; however, it can be easily changed to voltage control for constant voltage mode. The zero crossing instants of the secondary current $I_2$ are obtained by using a zero crossing detector and fed to the reset pin of a counter as shown in Figure 9 in order to obtain a sawtooth signal being synchronized to the half cycle of the secondary current [16]. The final value of the counter is stored to measure the duration of one half-period ($T_s/2$). For resistive mode operation, the PWM pulses must be centered in the middle of the period for both positive and negative cycles. For this purpose, the calculated conduction angle $\beta$ is converted to two compared values, COMPA and COMPB, which are $(1 - \beta)T_s/2$ and $(1 + \beta)T_s/2$, respectively, and then compared with the sawtooth signal in order to determine the switching instants by PWM pulse generation. The $\beta$ value is between 0 and 1. The generated gate pulse is then applied to both switches by a suitable gate driver, each of which conducts only for a half-period.
5. Experimental results

The primary-side coil is connected to a full-bridge inverter energized by a 1-kW adjustable DC power supply. The secondary-side active rectifier is designed to be controlled in resistive mode, as mentioned in Section 3. A 470-μF DC bus capacitor is used parallel to the battery in order to filter charge current. The output load of the secondary-side controlled rectifier is a 96-V battery package. The system is run at 75-V DC input voltage. The inverter on the primary side is controlled using the STM32F4 Discovery board, and the rectifier on the secondary side is controlled using the TMS320F28027 Launch Pad microcontroller board. The parameters in Table 1 are used in the prototype system shown in Figure 10.
The control scheme shown in Figure 8 and Figure 9 was implemented by a TMS320F28027 DSP. Discrete power IGBTs are used in the controlled rectifier and the generated PWM gate signal is applied to both IGBTs simultaneously. When the secondary-side controller detects zero cross $I_2$, an interrupt occurs. After that, a timer interrupt is set so that battery voltage and current are sampled four times in a half period of $I_2$. The DC bus current value is obtained by averaging the four samples taken in one half-period. Then the PI controller calculates the required conduction angle $\beta$. The oscilloscope waveform shown in Figure 11 is illustrating the timing and the operation of the circuit.

In Figure 12, the voltage and current waveforms of the primary and secondary-side are given for the simulation and experiments. The waveforms are recorded operating the rectifier open-loop by setting $\beta$ values to 60, 120, and 180 degrees, respectively. Experimental waveforms are in good agreement with simulation waveforms except for the voltage spikes at switching instants, which arise from the pcb layout. Figure 13 shows the individual measured efficiencies of the resonant inverter, coupled coils, and rectifier and overall efficiency with respect to the conduction angle in comparison with the simulated and calculated values. With the measured system model parameters, the analysis and simulation results can predict the real system performance well with
Figure 12. Simulation and experimental results of the voltage and current waveforms for primary (top) and secondary (bottom) sides: (a) $\beta = 60$, (b) $\beta = 120$, (c) $\beta = 180$. 
Figure 13. Experimental and theoretical analysis and simulation results for (a) inverter, (b) coil, (c) rectifier, (d) overall efficiency.

a 5% discrepancy in overall efficiency. The relatively low value of the overall efficiency mainly arises from the low operating voltage in the system. Because the on-state voltage drop of power switches in the power converters is not small enough with respect to the operating voltage, high conduction losses in the power converters occur. Additionally, the relatively high current at low voltage levels considerably increases the ohmic losses in all circuit stages including coupled coils. The measured output power versus conduction angle characteristics are also shown in Figure 14. It can be seen that the output power can successfully be controlled by the resistive conduction angle control method, and the results are consistent with the simulated and calculated values.

6. Conclusion
In this study, design of an IPT system with secondary-side control for battery charging of EVs has been introduced in detail. The converter losses that affect the system efficiency have been included in steady-state
analyses. The operating modes of the secondary-side rectifier and resistive/reactive control methods have been discussed. The implementation of the designed system was realized and the system performance was tested in the laboratory. The experimental results were observed to be in agreement with theoretical results. The overall system efficiency and power values are verified by the simulation and steady-state analyses.

The main advantage of the secondary-side controlled IPT systems is that they can operate successfully without need for communication between primary and secondary circuits. All control actions and power adjustments can be accomplished by the secondary side. A new control method that relies on the detection of the secondary coil current is proposed in this paper that reduces the control system complexity according to the methods proposed in the literature.
The reason for the relatively low system efficiency mainly arises from the low operating voltage that causes an increase of the conduction losses (i.e. ohmic losses and on-state voltage drop losses) significantly. In order to improve the overall efficiency, the system operating voltage should be increased to a level where the on-state voltage drop of the power switches becomes negligible, and also litz conductor should be used in coil wiring. Moreover, according to the results obtained, the resistive control mode should be employed since it is more efficient than reactive control mode for wide operating power ranges.

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