A Flexible Wireless Power Transfer System with Switch Controlled Capacitor

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ABSTRACT Wireless power transfer (WPT) with a switch-controlled capacitor is proposed in this paper for compensating load impedance which is related to the coupling coefficient of coils and the load variations. Generally, the power supply in the WPT system is desired to output constant voltage under varied load impedance. In the proposed WPT system, the resonant capacitor in primary side is replaced by a switch-controlled capacitor (SCC) and only diode full-bridge rectifier in the secondary side is used. The SCC realizes adjustable capacitor which makes it flexible to control the capacitance. At the same time, the modulation of the SCC aims to guarantee an inductive input impedance of WPT system, achieving soft switching and reducing the high switching losses of the high-frequency inverter. In order to verify the feasibility of the proposed method, mathematical analysis and experimental validations have been thoroughly performed. The experiment results show that the WPT system maintains a stable output voltage by through the regulation of the SCC according to the load variations.

INDEX TERMS Resonant inverter, switch-controlled capacitor, WPT, zero voltage switching.

I. INTRODUCTION Wireless power transfer (WPT) technology with the magnetic field utilize coupling coils to transfer energy through a relatively large air gap and can get rid of the restraint of the cable [1]-[2]. With the merit of convenience and safety, WPT has employed in many charging applications, for the low-power application of several watts such as smartphones [3]-[4] or biomedical implants [5], for the medium power application of several kilowatts such as charging electric vehicle [6]-[7], and even for the megawatts application such as the trains [8].

So far, extensive work both in theoretical analysis as well as practical aspects of design for WPT systems has been done by researcher around the world. In most of the literatures, either the maximum efficiency principle or maximum power transfer principle have been discussed, but the design of high-frequency power converters for WPT system has rarely been discussed [9].

Generally, the WPT systems are desired to provide a constant output voltage against to the variations of load impedance. The load impedance is related to the coupling coefficient and load capacity and usually is a time varying parameter in practical system. Taking the typical battery load as an example, the equivalent resistance of battery varies in a wide range during the whole charging or discharging process. For realizing the desired stable output voltage against a variable load impedance for WPT systems, [5], [10], [11] achieve stable voltage gain against varied load and coupling factors by tracking split frequencies. However, the stability problem caused by the frequency bifurcation phenomenon does not been fully considered. [12], [13] use a dc/dc converter in primary side or secondary side, which is also an effective way to regulate the output voltage. However, the converters cost a lot and introduce additional losses. [14], [15] adopt the dynamic impedance matching circuits to maximize the power transfer. This method can also be applied to regulate the output voltage.

To improve the performance of the WPT system, resonance capacitor is needed to compensate the inductive reactance of the coil. There are four basic resonant topologies according to the connection circuit of the resonance capacitors, including series-series, series-parallel,
parallel-parallel and parallel-series topologies [16]. Moreover, the resonant topologies with more than four elements are also developed for the WPT systems. However, more resonant elements increase the cost and complicate the design. In addition, the inductor and capacitor normally have large manufacturing tolerance, and the inductance and capacitance may be deviated due to aging [17]. This will further reduce the performance of the WPT system.

In this paper, a switch-controlled capacitor (SCC), connected in the primary side, is introduced to compensate the inductive reactance of the coil. By controlling the equivalent capacitance of the SCC, the inverter could keep in resonance operation and the performance of the WPT system can be enhanced with the ZVS operation achieved by modulating the SCC to an inductive load for the inverter. Only the diode full-bridge rectifier is used in the secondary side where the capacitor is removed. It is beneficial for integration and can be applied to more application backgrounds.

This paper is organized as follows. In section II, the circuit configuration is proposed. In section III, the corresponding control strategy for the proposed system is present. The control method is validated by experiments in section IV. Finally, the conclusions are drawn in section V.

II. CIRCUIT CONFIGURATIONS

A. SCC

The concept of SCC was proposed in [18] to regulate the resonant converters with fixed operating frequency. The SCC is used to regulate the equivalent capacitance, and thus to control the resonant frequency of the inverter. The controllable resonant frequency in turn controls the output of converters. In the proposed WPT system of this paper, the SCC in the primary side is regulated not only for compensating the load impedance, but also for producing soft-switching condition.

Figure 1 shows the circuit structure and the operation waveforms of the SCC. As shown in Figure 1(a), the SCC consists of two source-to-source connected MOSFET switches, S_a and S_b, and a parallel linear capacitor C_o.

A sinusoidal current I_{ab} is applied to the SCC from terminal a to terminal b, as shown in Figure 1(b). It has

\[ I_{ab} = I_{ab} \sin(\omega_0 t) \]  

where I_{ab} is the amplitude of the input current and \omega_0 is the angular frequency of the input current. Gate signals for driving the switches S_a and S_b are synchronized with current I_{ab}. They have phase shift \alpha from i_{ab} and are complementary each other, as shown in Figure 1(b). In the interval of [\alpha, \alpha + \varphi], the voltage across the C_o is

\[ u_{ab} = \frac{1}{C_o} \int_{\alpha}^{\alpha + \varphi} I_{ab} \sin(\omega_0 t) dt = \frac{I_{ab}}{\omega C_o} \left[ \cos(\alpha) - \cos(\alpha + \varphi) \right] \]  

When \alpha=\alpha+\varphi, u_{ab}=0, substituting it into (2). It has

\[ \alpha = \pi - \frac{\varphi}{2} \]  

The voltage amplitude of the fundamental component can be calculated by Fourier series from (2). It has

\[ U_{ab(1)} = \frac{2}{\pi} \int_{0}^{\alpha/2} u_{ab} \cos(\omega_0 t) dt \]

\[ = \frac{I_{ab}}{\omega_0} \left[ - (\pi - \alpha + \frac{1}{2} \sin 2\alpha) \frac{2}{\pi C_o} \right] \]  

where U_{ab(1)} is the voltage amplitude of the fundamental component. From (4), the equivalent capacitance of the SCC can be expressed as

\[ C_{sc} = \frac{\pi C_o}{2\pi - 2\alpha + \sin 2\alpha} \]  

where C_{sc} is the equivalent capacitance of the SCC. Figure 2 shows the capacitance ratio between the equivalent capacitance C_{sc} and parallel capacitor C_o with respect to the changes of \alpha by use of (5). It is clearly shown in Figure 2 that the equivalent capacitance C_{sc} can be regulated effectively by changing \alpha. In the case of \alpha=\pi/2, where the capacitance ratio C_{sc}/C_o equals 1, the capacitor C_o is always connected with the circuit and the current flows continuously from terminal a to terminal b. The switches S_a and S_b are effectively OFF all the time, and there is no current flowing through them. In the case of \alpha=\pi, there is no current flowing from terminal a to terminal b. The capacitor C_o is always by-passed by switches which are effectively ON all the time. In this case, the SCC performs as a capacitor with infinite capacitance.
\[ \theta_p = \arctan\left(\frac{\sin(2\pi D)}{1 - \cos(2\pi D)}\right) \]
In this way, the input impedance angle $\phi$ should be selected to be larger than the phase angle $\theta_1$ in (11)

$$
\phi \geq \arctan \left( \frac{\sin(2\pi D)}{1 - \cos(2\pi D)} \right)
$$

The ZVS operation can be built up by meeting the inequality constraint of (16). It should be noted that (16) is only a necessary but not sufficient condition for ZVS condition because parasitic capacitances, such as output capacitance of switch devices and stray capacitance require enough energy from inductance for charging and discharging during dead time. In other word, the phase angle $\phi - \theta_1$ should be larger than the dead time setting of switches.

Figure 7 shows the operation modes of inverter and typical waveforms with following assumed.

1) The output capacitances of MOSFETs has the same values.

2) The resonant current is approximated to be sinusoidal. According to the current flow, we divide the operation of the inverter into six modes.

Mode 1 [$t_0$, $t_1$]: The resonant current flow is highlighted, and the rest of circuit is faded. At time $t_0$, $S_2$ is turned off. The triggering signal of $S_1$ is delayed to prevent $S_1$ and $S_2$ from being shorted. At this situation, the energy is delivered from load to source.

Mode 2 [$t_1$, $t_2$]: As shown in Mode 2 of Figure 7(b), at time $t_1$, the triggering signal of $S_1$ is raised to active $S_1$, the resonant current is negative as same to Mode 1 and flows through $D_1$, then, $S_1$ turns on at zero voltage.

Mode 3 [$t_2$, $t_3$]: after time $t_2$, the resonant current becomes positive, and through the $S_1$ and resonant circuit from the other direction. The energy is transferred from DC to the resonant circuit until the triggering signal of $S_1$ is removed.

Mode 4 [$t_3$, $t_4$]: At time $t_3$, $S_1$ is turned off. The resonant current is still positive and flowing through the diode $D_2$ of $S_2$ as shown in Mode 4 of Figure 7(b). The triggering signal of $S_2$ is delayed to trigger, entering the dead time interval to prevent $S_1$ and $S_2$ from being shorted.

Mode 5 [$t_4$, $t_5$]: at time $t_4$, the triggering signal of $S_2$ is raised to active $S_{12}$, the resonant current is negative as same to Interval 4 and flows though diode $D_2$, then, $S_2$ turns on at zero voltage.

Mode 6 [$t_5$, $t_6$]: after time $t_5$, the resonant current
becomes negative, and through the S2 and resonant circuit from the other direction. At time \( t_6 \), the switching period is completed, and the same operation continues.

### IV. CONTROL STRATEGY

#### A. REGULATION OF EQUIVALENT CAPACITANCE

According to the impedance matching rule, the equivalent capacitance \( C_{eq} \) with maximum voltage gain could be obtained by removing imaginary part of the impedance \( Z_m \) shown in (15). It is

\[
C_{eq} = \frac{(R_2 + R_{eq})^2 + (\omega_0 L_2)^2}{((R_2 + R_{eq})^2 + (\omega_0 L_2)^2)\omega_0^2 L_4 + \omega_0^4 M^2 L_2} \tag{17}
\]

The equivalent capacitance \( C_{eq} \) should be regulated by \( (6) \) to meet the requirement of \( (17) \), and then the system will be operated at the supposed resonant frequency. The voltage gain is defined based on the AC input voltage \( V_{in} \) and the AC output voltage \( V_{out} \), and it has

\[
G_i = \frac{j\omega_0 M R_{eq}}{(R_i + j(\omega_0 L_4 - \frac{1}{\omega_0 C_{eq}}))(R_2 + R_{eq} + j\omega_0 L_2) + (\omega_0 M)^2} \tag{18}
\]

Table I shows the parameters of the inverter system. With the parameters listed in Table I, the voltage gain as a function of equivalent capacitance \( C_{eq} \) is given in Figure 8. When \( f_0 = 100 \text{ kHz} \), the value of \( C_{eq} \) is calculated as 25.7 nF by \( (17) \). Figure 8 is shown that the maximum gain can be achieved at the point of 102.7 nF, 60.8 nF, 35.3 nF, 25.7 nF, 16.5 nF and 11.4 nF for equivalent capacitance \( C_{eq} \), where \( f_0 \) is 50 kHz, 65 kHz, 85 kHz, 100 kHz, 125 kHz and 150 kHz, respectively. However, the input impedance is very low at such point with maximum gain, as shown in Figure 9. Then large input current would be caused by low input impedance and may affect the safety operations of the inverter, resonant capacitor, and primary coil. Therefore, the equivalent capacitance \( C_{eq} \) needs to be regulated to offset the maximum voltage gain.

Figure 10 shows the phase angle curve of input impedance under the different operating frequency. As the frequency \( f_0 \) is set at 50 kHz and the value of \( C_{eq} \) is less than 102.7 nF, the inverter system will not realize ZVS operation because of the capacitive load characteristics. It should be noted that once certain operating frequency is determined, the regulated equivalent capacitance should be larger than the capacitance value with the maximum voltage gain at same frequency, which ensures the load with inductive characteristics.

### TABLE I

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Quantity</th>
<th>Value</th>
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<tbody>
<tr>
<td>( L_1 )</td>
<td>Inductance of primary coil</td>
<td>100 ( \mu )H</td>
</tr>
<tr>
<td>( L_2 )</td>
<td>Inductance of secondary coil</td>
<td>100 ( \mu )H</td>
</tr>
<tr>
<td>( R_i, R_s )</td>
<td>ESR</td>
<td>0.5 ( \Omega )</td>
</tr>
<tr>
<td>( M )</td>
<td>Mutual inductance</td>
<td>12 ( \mu )H</td>
</tr>
<tr>
<td>( R_{eq} )</td>
<td>Equivalent AC load</td>
<td>6 ( \Omega )</td>
</tr>
</tbody>
</table>

![FIGURE 8. Voltage gain versus equivalent capacitance \( C_{eq} \).](image)

![FIGURE 7. Operation modes of inverter. (a) Typical waveforms, (b) Current flow of resonant circuit at each interval.](image)
Figure 11 shows the equivalent capacitance $C_{eq}$ versus various equivalent loads $R_{eq}$ based on the given parameters of $f_0=100$ kHz and $|G_s|=1.0$. Obviously, the equivalent capacitance $C_{eq}$ is regulated via the phase shift angle $\alpha$ to track the operating frequency and voltage gain. Based on the given parameters of $f_0$ and $|G_s|$, there are two roots for $C_{eq}$ with a given equivalent load $R_{eq}$, as shown in Figure 11. In other words, there are two capacitance points for $C_{eq}$ to stabilize the load: one load with capacitive characteristics and another load with inductive characteristics. The larger root of $C_{eq}$ would be favorable because the equivalent load is inductive in this case, which may realize ZVS operation for the high-frequency inverter.

**B. CONTROL METHOD**

A control strategy is designed for the effective operation of the proposed WPT system. Figure 12 gives the flowchart of the control strategy. As shown in Figure 12, the operating frequency and voltage gain of the system should be first determined. The control strategy starts by measuring the output voltage and current, and then, calculating the equivalent load $R_{eq}$. Based on the equivalent load, the equivalent capacitance $C_{eq}$ is regulated. When the feedback signal $U_F$ is bigger than the reference signal $U_{ref}$, the duty cycle $D$ is increased to keep the output voltage stable.

In addition, by considering the system stability, the feedback control is implemented, as shown in Figure13. The feedback signals of voltage and current obtained by the RF modules are transferred wirelessly to provide for SCC PWMs and APWM generation. The detection of the equivalent load is processed for a given operating frequency and voltage gain. The adjusted duty cycle values are then used to control the SCC PWMs. The SCC PWMs are synchronized with the zero crossing points of the primary side current of the corresponding phase.

**FIGURE 9.** Input impedance versus equivalent capacitance $C_{eq}$

**FIGURE 10.** Phase angle curves of input impedance under different operating frequency.

**FIGURE 11.** Equivalent capacitance $C_{eq}$ versus various equivalent load $R_{eq}$

**FIGURE 12.** Flowchart of the control strategy for the half-bridge inverter.

$U_F$ is less than the reference signal $U_{ref}$, the duty cycle $D$ is increased to keep the output voltage stable.

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V. EXPERIMENTAL VERIFICATIONS

A. EXPERIMENTAL PLATFORM SETUP

In order to investigate the feasibility of the proposed system, a hardware prototype has been implemented and tested in the laboratory. The experimental setup of the proposed system is shown in Figure 13, and the parameters of the experimental platform are listed in Table II.

As shown in Figure 14, the power starting from a DC power supply flows through the half bridge inverter, the SCC, the coupled coils and the full bridge rectifier. Finally, it provides to the electrical load which serves as the adjustable load resistor. The inverter and the SCC are controlled by the PWM interfaces from digital signal processor (DSP) controller based on the feedback signals of the load. The feedback signals are transferred by the communication link. As shown in Figure 15. The coupled coils are made of Litz wire. The structure of coils directly faces each other and turn numbers of primary coil and the secondary coil are all 20. The original distance between them is 150 mm. However, when the coils are misaligned, the mutual inductance $M$ may be varied, and the voltage gain is changed. Then the SCC would be regulated to achieve a stable voltage gain.

B. EXPERIMENTAL RESULTS

Figure 16 shows the measured waveforms of the SCC with modulation under different operating frequency, where $u_{ab}$ is the voltage across the parallel capacitor $C_a$, $i_1$ is the current through the SCC, $V_{gs4}$ is the gate signal of $S_4$ and $V_{gs3}$ is the gate signal of $S_3$. Figure 16 (a) shows the scenario with operating frequency 85 kHz and the control angle 137.3°. Figure 16 (b) shows the scenario with operating frequency 100 kHz and the control angle 144.0°. In addition, because $u_{ab}$ is always zero at turn-on and turn-off points, implying that switch $S_a$ and $S_b$ are switched both ON and OFF at ZVS conditions.

![Figure 13. System block diagram with the proposed control strategy.](image1)

![Figure 14. Experimental platform of the proposed system.](image2)

![Figure 15. Structure of coupled coils.](image3)

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**TABLE II**

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Quantity</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_1$</td>
<td>Primary coil inductance</td>
<td>99.89 $\mu$H</td>
</tr>
<tr>
<td>$L_2$</td>
<td>Secondary coil inductance</td>
<td>100.02 $\mu$H</td>
</tr>
<tr>
<td>$N_1$</td>
<td>Number of primary coils</td>
<td>20 turns</td>
</tr>
<tr>
<td>$N_2$</td>
<td>Number of secondary coils</td>
<td>20 turns</td>
</tr>
<tr>
<td>$R_{eq}$</td>
<td>ESR</td>
<td>0.4 $\Omega$</td>
</tr>
<tr>
<td>$M$</td>
<td>Mutual inductance</td>
<td>12.04 $\mu$H</td>
</tr>
<tr>
<td>$C_s$</td>
<td>Series capacitor</td>
<td>66 nF</td>
</tr>
<tr>
<td>$C_a$</td>
<td>SCC capacitor</td>
<td>15 nF</td>
</tr>
<tr>
<td>$C_0$</td>
<td>Output capacitor</td>
<td>470 $\mu$F</td>
</tr>
</tbody>
</table>
Figure 17 shows the measured waveforms of the half inverter switches under different duty cycle, where \( u_{ds1} \) and \( u_{ds2} \) are the drain-source voltage of the switch \( S_1 \) and \( S_2 \), respectively, \( i_{ds1} \) and \( i_{ds2} \) are the current through the switch \( S_1 \) and \( S_2 \), respectively. The waveforms of drain-source voltage and current under the duty cycles with 0.2, 0.4, 0.6 and 0.8 are individually measured. It is clearly shown that the current \( i_{ds1} \) and \( i_{ds2} \) change from negative to positive when each switch \( S_1 \) or \( S_2 \) is turned on. In other words, the current flows through the parallel-diode of each switch before it is activated, thereby confirming that the inverter effectively performs the ZVS operation.

To maintain a stable output voltage, the equivalent capacitance \( C_{eq} \) is regulated to keep a constant voltage gain as the load changes. Figure 18 shows \( C_{eq} \) with various loads \( R_L \) based on the selection of \( f_s=100 \) kHz. Figure 18(a) shows the SCC operation when the load was 25 Ω. Here, the control angle is measured as 122.6°, the \( C_{eq} \) is calculated as 26.05nF. Hence, the inverter maintains an inductive load state and an adequate voltage gain with \( |G_i|=2.0 \). Figure 18(b) shows the SCC operation when the load was 50 Ω. Here, the control angle is measured as 124.3°, the \( C_{eq} \) is calculated as 27.30nF. Figure 18(c) shows the SCC operation when the load was 75 Ω. Here, the control angle is measured as 125.4°, the \( C_{eq} \) is calculated as 28.15nF. Figure 18(d) shows the SCC operation when the load was 100 Ω. Here, the control angle is measured as 125.8°, the \( C_{eq} \) is calculated as 28.42 nF. These results prove that the proposed WPT system is highly flexible, because it provides constant output voltage against load variations. That is very practical for industry applications, especially for battery charging.

C. COMPARISON

The comparisons of the proposed method and previous techniques are shown in Table III, where different control methods and purposes are used for the resonant tuning element with variable capacitance. In [21] and [22], the capacitor arrays are applied to select the suitable capacitance, which are bulky and cannot generate a continuous capacitance regulation. In [4], a PWM-controlled capacitor was proposed to realize self-tuning of the LCC converter. Since the capacitance value is regulated by single switch with varied PWM on/off duty ratio, the voltage stress will be high, and the range of variable capacitance is very small. In this paper, the load impedance is compensated by regulating the SCC. At the same time, the modulation of the SCC guarantees an inductive input impedance of WPT system, achieving soft switching and reducing the high switching losses of the high-frequency inverter.
VI. CONCLUSIONS

In this paper, a SCC is adopted in the primary side of the WPT system. The SCC can regulate equivalent capacitance easily which makes it flexible to control the resonance point of the high-frequency inverter. So, it can not only compensate the leakage of the coupled coils and large manufacturing tolerance, but also ensure inductive impedance for the WPT system, which achieves soft switching operation and reduces switching losses greatly for high-frequency inverter. The capacitance of the SCC can be regulated according to the operating frequency and the load variations. A prototype is built to verify the feasibility of the proposed WPT system, and through the experiment results, it is validated that the inverter effectively performs the ZVS operation. Meanwhile, various load conditions have been applied to verify the performance of the proposed system.

References