Charging Time Control of Wireless Power Transfer Systems Without Using Mutual Coupling Information and Wireless Communication System

W.X. Zhong, Member, IEEE and S.Y.R. Hui, Fellow, IEEE

Abstract—A charging time control method for wireless power transfer systems with a secondary-side hysteresis output power control is presented. It is a primary-side control method that adopts the combined use of three concepts, namely (i) an intermediate capacitor in the receiver circuit as a power flow indicator, (ii) hysteresis switching actions of a shunt decoupling power switch in the receiver circuit to regulate the DC voltage of such intermediate capacitor and (iii) the turn-on and turn-off times of the decoupling switch detected on the primary side for closed-loop control. This method has the advantage of eliminating the needs for (i) precise information of the mutual inductance between the transmitter the receiver coils and (ii) wireless communication system for feedback purpose. Practical results obtained from a hardware prototype are included. They confirm the proposed operating principles and indicate that the method can automatically lead to optimal energy efficiency operation.

Index Terms—Wireless power transfer (WPT), primary-side control

I. INTRODUCTION

WIRELESS power transfer (WPT) systems have attracted lots of attentions in research communities and industry in the last two decades. Recent research efforts reported in the literature focus on several major applications, including wireless charging of (1) mobile robots [1], (ii) consumer electronics [2] and (iii) medical implants [3-5]. Applications of WPT face a common issue of the need for some form of feedback control. In general, control executed in the transmitter circuit is called the primary-side control. Primary-control methods for WPT systems can be classified into the following groups:

(i) Use of primary side control that requires wireless communication systems [6]-[8] to feed information obtained from the receiver circuit back to the transmitter circuit for closed-loop control (Fig. 1).

(ii) Use of primary (transmitter) control that requires information of the mutual coupling (k) or mutual inductance (M) [9]-[13] between the Tx and Rx coils.

The second group of primary-side control methods that require the information or estimation of the mutual inductance (M) or mutual coupling (k) between the Tx coil and the Rx coil can be found in the references [9]-[13]. In general, they either use pre-determined value of M or estimate the value of M so that the output voltage can be calculated from information available on the transmitter (primary) side.

In this paper, a method that eliminates the needs for both of the mutual coupling information and wireless communication system for feedback control is proposed. It is based on the combined use of several concepts, including (i) an intermediate capacitor in the receiver (secondary) circuit as a power flow

Fig. 1. Primary control of a wireless power transfer system using a wireless communication system for feedback purpose [8].
indicator, (ii) the hysteresis switching actions of a shunt decoupling power switch in the receiver circuit to regulate the DC voltage of such intermediate capacitor and (iii) the turn-on and turn-off times of the decoupling switch detected on the primary side for closed-loop control. This method is suitable for stationary wireless charging of portable electronics and has been experimentally verified and practical measurements are included in this paper to highlight its operating principles.

II. EXISTING CONCEPTS RELEVANT TO THE PROPOSED METHOD

Fig. 2. An equivalent circuit for an inductive power transfer system with a shunt decoupling switch S in the secondary circuit (redrawn from [14]).

Another type of decoupling switch has been described in [15], in which a series switch is incorporated in the charging protection circuit (Fig. 3a). This “series” switch Ms is used to protect the battery from over-voltage and over-current condition (Fig. 3b). Under normal situation, this series switch Ms is turned on, linking the battery load to the rectified DC voltage Vs in the secondary circuit. When the battery voltage Vb exceeds a certain upper threshold voltage, Ms will be turned off through the control of the “voltage protection circuit” in order to avoid over-voltage condition. Ms will be turned on when Vb falls to a lower threshold voltage level. Ms will also be turned off when the charging current exceeds a certain maximum level through the control of the “current protection circuit” (Fig. 3b). When the series decoupling switch (Fig. 3b) is off, the load is isolated from the secondary circuit electrically. The reflected impedance to the primary side becomes smaller and will cause a reduction in the primary voltage.

III. THE PROPOSED CHARGING TIME CONTROL

A. Basic Principles

In many WPT applications, there is a need for a simple primary control method that (i) is independent of the precise knowledge of the mutual inductance between the Tx and Rx coils, (ii) does not require any wireless or radio-frequency (RF) communication systems between the Tx and Rx circuits. An example is the charging of the mobile phones on a wireless charging pad. When the user places the mobile phone on the charging surface, the mutual inductance between the Tx and Rx coils may not be the same in each time.

Although (i) the switching actions of the decoupling switch in the secondary side of wireless power transfer system will affect the reflected impedance to the primary side and (ii) such switching actions will cause sudden changes in the magnitude of the primary current, there has not be any systematic theory and method on how to utilize these two pieces of information for closed loop control on the primary side without requiring the mutual inductance information and RF communication system to feed back the load information.

The proposed primary-side control method can be realized in a WPT system using (1) a primary-side input power control and (2) a secondary-side hysteretic output power control, without any knowledge of mutual inductance and RF communication system for feeding back the load conditions to the primary side. It involves the combined use of three existing elements as follows:

(i) The use of the voltage of an intermediate capacitor in the secondary circuit as an indicator of power/energy transfer,
(ii) The use of a hysteretic control of the decoupling switch to “charge” and “discharge” this intermediate capacitor so as to regulate the capacitor voltage within the upper and lower tolerance band of a nominal voltage level;
(iii) The use of the charging time detected in the primary circuit as the feedback variable for controlling the input
power of the primary circuit for closed-loop control of the WPT system.

Now consider a general schematic of a WPT system in Fig. 4. Assume that the transmitter coil resonator and the receiver coil resonator are identical. If the load $R_L$ on the secondary side is reflected to the primary side, the reflected load resistance $R_{reflected}$ becomes [16]:

$$R_{reflected} = \frac{\omega^2 M^2}{R_{p2} + R'_L},$$

(1)

where $R'_L$ is the equivalent impedance seen into the rectifier, $R_{p2}$ is the resistance of the secondary winding, $\omega$ is the angular frequency and $M$ is the mutual inductance between the primary ($Tx$) and secondary ($Rx$) windings. The output power delivered to the secondary side ($P_2$) is:

$$P_2 = i_p^2 \frac{\omega^2 M^2}{R_{p2} + R'_L},$$

(2)

where $i_p$ is the current in the primary ($Tx$) winding.

If the power losses of the converters and $R_{p2}$ are negligible compared to the output power, $P_2 = P_{out}$, where $P_{out}$ is the output power and can be expressed as

$$P_{out} = i_p^2 \frac{\omega^2 M^2}{R'_L}.$$  

(3)

On the other hand, the output power can also be expressed as

$$P_{out} = \frac{V_{out}^2}{R_L},$$

(4)

where $V_{out}$ is the output DC voltage of the secondary circuit.

By combining (3) and (4),

$$V_{out}^2 = \frac{R_L}{R'_L} \omega^2 M^2 i_p^2.$$  

(5)

Taking square root on both sides of (5),

$$V_{out} = \frac{R_L}{\sqrt{R'_L}} \omega M i_p.$$  

(6)

Equation (6) indicates that, for a given angular frequency $\omega$ and mutual inductance $M$, the output voltage $V_{out}$ of the secondary circuit is proportional to the primary current $i_p$.

$$V_{out} = \omega \lambda i_p.$$  

(7)

where $\lambda = \frac{R_L}{\sqrt{R'_L}} M$.

For “stationary” wireless charging (which means that there is no dynamic movement between the Tx and Rx coils during the charging period), the mutual inductance $M$ can be considered as a constant. Therefore, the coefficient $\lambda$ in (7) is a constant for each set of relative locations of the Tx and Rx coils. Equation (7) now gives us an understanding that, for a given angular frequency $\omega$, $V_{out}$ is proportional to $i_p$. However, it is important to note that this coefficient $\lambda$ is also an unknown, although it is a constant. So it is necessary to develop a new control method that does not need to calculate $\lambda$. Such control method will then be independent of $\lambda$ and $M$.

The operating principle of the proposed method can be applied to a WPT system with a decoupling switch and a capacitor in the secondary circuit. It is demonstrated in an example shown in Fig. 5. It should be noted that the load in Fig. 5 is represented as a resistive load $R_L$ to simplify the explanation. In practice, such load is not restricted to a resistive load. For example, it may consist of a DC-DC power converter charging a battery under the control of a battery charging controller. This means that the intermediate capacitor $C_o$ may not be connected directly to a pure resistive load. The resistive load in Fig. 5 therefore can represent a sub-circuit feeding the actual physical electric load.

Fig. 5 shows a WPT system with a “shunt” decoupling switch that is used together with a hysteresis controller for regulating the voltage across the intermediate capacitor $C_o$. Since the switch S is placed on the DC side of the rectifier, a directional switch can be used. On the transmitter side, a primary-side controller is used to control the input power. The details of such primary-side controller will be given later in this document. In summary, the proposed method involves (1) a primary-side controller for controlling the input power and (2) a hysteresis controller for controlling the decoupling switch in order to regulate the voltage across (3) an intermediate capacitor in the secondary circuit.

**B. Secondary-Side Hysteresis Control**

Now, it is necessary to explain how the primary-side control and the secondary-side hysteresis control work together. We first explain the operation of the hysteresis controller for regulating the voltage across the capacitor $C_o$ using the secondary circuit of Fig. 5 as an example. The voltage of the capacitor can be used as a power transfer indicator. Using the
“secondary circuit” of Fig.5, the dynamic equations of the intermediate capacitor are:

\[ i_c = i_i - i_2, \]  
\[ i_c = C_o \frac{dv_o}{dt}, \]  

Equations (8) and (9) show that the intermediate capacitor current can be used as an indicator for power transfer. For the capacitor voltage regulated to a nominal (reference) value \( V_o^* \) within a tight tolerance by the hysteresis control, if the current supplied to the capacitor \( (i_i) \) meets the load current \( (i_2) \), the capacitor current \( (i_c) \) becomes zero according to (8). If there is any imbalance between \( i_1 \) and \( i_2 \), it will be reflected in the capacitor voltage \( (V_o) \) according to (9). Therefore, by using a secondary-side hysteresis control to ensure that the \( V_o \) within a narrow tolerance band of \( V_o^* \), the load demand will be automatically met by power supply.

When the shunt decoupling switch S is turned on, \( i_1 = 0 \) and \( i_c = -i_2 \) according to (8). Thus the capacitor \( C_o \) is being discharged and the capacitor voltage \( (V_o) \) is ramping down as shown in the timing diagram in Fig. 6.

[Fig. 6. Waveforms of the capacitor voltage, the gate signal for the decoupling switch and the primary current in the time domain.]

When the shunt decoupling switch S is turned off, \( i_2 > 0 \) and \( i_c = i_1 - i_2 \). Since the primary-side control is designed to provide sufficient power for the load in the secondary, \( i_1 > i_2 \) under normal operation. When S is turned off, \( i_1 \) is positive and the capacitor voltage \( (V_o) \) is ramping up as shown in Fig. 6.

The secondary-side hysteresis control can be understood from charging and discharging process by the decoupling switch in Fig. 6. When the decoupling switch S is turned on, the load is decoupled electrically from the secondary \( (R_s) \) circuit. The reflected impedance of the secondary circuit on the primary-side will appear as a large impedance. Consequently, the current in the primary winding \( (i_p) \) will suddenly drop as indicated in Fig. 6 when S is turned on. When S is turned off again, the reflected impedance will become relatively small again, therefore \( i_p \) will increase back to the normal magnitude. The magnitude of \( i_p \) in the primary circuit provides the information of the switching states of the decoupling switch S in the secondary circuit. Therefore, it forms the link between the primary-side input power control and the secondary-side hysteresis control.

C. Primary-Side Input Power Control

Since the magnitude of the primary winding current contains the information of the switching state of the decoupling switch S, the primary-side input power control can be coordinated with the secondary-side hysteresis control through the monitoring of the magnitude of the primary winding current \( (i_p) \). The primary-side input power control involves the detection of at least one electronic variable in the primary circuit, such as the current or the voltage of the primary \( (T_X) \) winding. In general, any change of the reflected impedance can be reflected in the change of magnitude of the primary current. From the magnitude of the primary winding current \( (i_p) \), the capacitor charging time \( t_c \) can be estimated by sensing the decoupling switch turn-on moment \( t_{c,1} \) and the decoupling switch turn-off moment \( t_{c,2} \). The difference between \( t_{c,2} \) and \( t_{c,1} \) equals to the charging time \( t_c \). Then \( t_c \) will be compared with a charging time reference \( t_{c,ref} \) to produce a control signal \( \Delta t \), which in turn can be fed to a compensator or controller (such as lead-lag or proportional-integral PI controller) to vary the control variable of the transistor circuit.

\[ \Delta t = t_{c,ref} - (t_{c,2} - t_{c,1}) \]  

\( \Delta t > 0 \) means that the charging time is “shorter” than the designated value, therefore the input power should be “decreased” in order to prolong the charging time.

\( \Delta t < 0 \) means that the charging time is “longer” than the designated value, therefore the input power should be “increased” in order to shorten the charging time. The control variables of the primary-side power converter should be controlled to increase the input power on the primary side.

For power converters using phase-shift control, the variable is the phase shift angle. The phase shift angle \( (\alpha) \) refers to the phase angle between the rising edges of the turn-on signals of the diagonal switch pair of S1 and S3 as shown in Fig. 7.

[Fig. 7. Control block and switching diagram for the primary full-bridge inverter applying phase-shift control method.]
D. Determination of the Reference Charging Time $t_{c-ref}$

Due to the nonlinearity of the switches, it is difficult to derive a precise analytical expression for the optimum charging time $t_{c-ref}$ in (10). However, the optimal value of $t_{c-ref}$ can be easily obtained in a searching process. The searching process for the optimal $t_{c-ref}$ is explained as follows. The controller starts the searching from zero phase-shift (which means applying the maximum input voltage to the primary coil at a minimum charging time). The input power is regularly monitored. As the controller increases the phase-shift of the inverter in order to reduce the input voltage, the charging time will increase accordingly. The searching process will end when a minimum input power for a given load is found. Finding the minimum input power for a given load power is equivalent to locating the condition for the maximum energy efficiency. Thus, an optimum charging time is found. This searching process can be regularly carried out for detecting any slow changing of load and addition or removal of load.

IV. SIMULATION AND PRACTICAL VERIFICATION

The invention has been verified initially with circuit simulation and then with practical experiments for a WPT system with two identical coils as shown in Fig. 8 and the structure of the system is shown in Fig. 9. A DC-AC power converter with phase-shift control is used for verification. The parameters of the system are specified in TABLE I. The operating frequency of the power inverter in the transmitter circuit is set at 97.56 kHz which is approximately equal to the resonant frequencies of two resonators. The litz wire has 24 strands of no. 40 AWG (0.08 mm diameter) and a ferrite plate with a thickness of 1 mm is used for shielding. The nominal output DC voltage in the secondary circuit is set at 15 V.

<table>
<thead>
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<th>Parameter</th>
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<th>Parameter</th>
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<tr>
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<td>0.3 Ω</td>
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<tr>
<td>$C_2$</td>
<td>105.3 nF</td>
<td>$C_3$</td>
<td>106.2 nF</td>
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Fig. 8. The coil of a practical WPT system.

Fig. 9. A WPT system with a coupling coefficient of about 0.44.

A. Simulation Study

A pure AC voltage source ($v_{in}$ of Fig. 5) is used to feed the primary coil resonator (comprising $C_1$ and $L_1$) in this simulation study so as to evaluate the relationship between the switching action of the decoupling switch, the magnitude of the primary winding current and the charging/discharging times of the intermediate capacitor in the secondary circuit. The root-mean-square (RMS) value of the input voltage ($v_{in}$) is assumed initially set at 5 V. The hysteresis control in the receiver circuit regulates the voltage of the intermediate capacitor at a nominal value of 15 V with a hysteresis tolerance band of ± 0.35 V.

Fig. 10 shows the simulated current and voltage waveforms of the primary resonator and the output voltage waveform when the input source voltage is 5 V. $v_{GS}$ represents the gate-source voltage of the N-type MOSFET. When $v_{GS}$ is high, the decoupling MOSFET is turned on to decouple the load. The reflected impedance on the primary side suddenly increases and so the magnitude of the current in the primary resonator becomes low. The output capacitor discharges from about 15.35 V to 14.65 V in this interval. The discharging time is approximately 90 µs. When the MOSFET is off, it is the charging interval and the time duration is about 140 µs. From Fig. 10, it can be seen that the charging time of the intermediate capacitor can be estimated by monitoring the changes of the primary winding current or voltage.

B. Experimental Verification

Experiments have been carried out based on the system of Fig. 7. The DC source voltage of the phase-shift power inverter is set at 7 V and the switching period equals to 10.25 µs. Fig. 11 shows the phase shifted voltage waveform fed into the primary resonator. Here conduction ratio is used to represent the conduction time of the inverter legs over the switch period. For example, when the phase shift is zero, the conduction ratio is 1 which produces a typical 50% duty cycle square wave. With
increasing phase shift, the conduction ratio decreases and the input voltage to the primary resonator drops accordingly.

Fig. 11. Measured voltage applied to the primary coil.

Fig. 12. Measured waveforms of the output voltage (top trace: 2 V/div), primary coil current (middle trace: 2 A/div) and secondary coil current (bottom trace: 2 A/div).

Fig. 13. Measured output voltage (top trace: 5 V/div) and averaged input DC current signal (scaled down signal).

Fig. 14. Measured energy efficiency as a function of the conduction ratio [(180°-α)/180°] when the load resistance is 50 Ω.

The variation of the energy efficiency of the system as a function of the conduction ratio for the same load of 50 Ω can be measured by manually adjusting the phase-shift angle. Such plot is shown in Fig. 14. The search for maximum efficiency is equivalent to reaching the top of this efficiency curve. When the input voltage is lower than the optimum value, the efficiency will decrease. Accordingly, the input current will increase which has been shown in Fig. 13. When the controller senses this current increase, it will then stop the searching and

reflected load impedance becomes a large value that the magnitude of the primary current becomes small (as shown in the middle trace of Fig. 12). When the decoupling switch is turned off, the reflected load impedance becomes relatively small and the magnitude of the primary coil current becomes relatively large. This set of practical measurements confirm the proposed concept that the magnitude of the primary current on the transmitter side provides the information of the switching actions of the hysteresis control in the receiver circuit.

The searching process in reaching the optimal charging time is recorded in Fig. 13 when the load resistance is 50 Ω. For the system to reach the highest energy efficiency operating point, the averaged input current from the constant DC voltage source (Vs) of the power inverter in Fig. 7 should be minimized. The time for the system to reach this point can be observed by checking how long it takes for the averaged input DC current from Vs to reach a steady state value. Because the current from the DC voltage source consists of discrete current pulses (typically in rectangular shape) due to the switching actions of the inverter, the sensed input current can be filtered with a small capacitor so as to derive the averaged input current. Such averaged input current is captured and displayed in Fig. 13. The top trace of Fig. 13 is the output voltage and the bottom one is the filtered input current signal captured by the controller. Observation of the averaged input DC current waveform indicates that, as the phase shift increases, the input current decreases until it reaches the minimum value. Further increase in the phase shift would lead to an increase in the input current. So the operating point moves back to its minimum input current level so that the optimal energy efficiency point can be reached. In this setup, the system takes about 520 ms to reach this optimal operating point.
return the operation to the optimum point. Then the charging time at the optimum point will be recorded and used as the reference charging time value. Fig. 15 and Fig. 16 show the same set of output voltage and input DC current waveforms before and after the optimal point is reached. In order to highlight the ripples of the waveforms, the actual ground levels of these waveforms are set below the display area. The enlarged views of the waveforms are shown in the two bottom traces. It is noted that the charging time after the optimal point is reached is longer.

The transient performance of the system with the load changing from 75 Ω to 50 Ω (from light load to heavy load) is shown in Fig. 17. The first top trace in Fig. 17 is the measured output voltage (regulated at 15 V) and the second top trace is the load-change signal that triggers the change of load from 75 Ω to 50 Ω. A high load-change signal means that the load is 75 Ω and a low signal means it is 50 Ω. Note that the ground level of the output voltage waveform is set below the display area in Fig. 17. The averaged output voltage is actually regulated at 15 V. The two bottom traces are the enlarged waveforms of the output voltage and the load-change signal. It can be observed from the enlarged waveforms that the output voltage regulation remains effective under transient load change conditions. Right after the load change, the charging time becomes longer because the initial inverter voltage was operating for a load of 75 Ω and it takes a longer time for the same inverter output voltage to charge the output capacitor for providing more power for the load of 50 Ω. But the control loop quickly reaches the new steady operation point as shown in the enlarged steady-stage waveforms after the transient load change as shown in Fig. 18 in which the conduction ratio is defined as (180°−α)/180°.

An additional test has been conducted to evaluate the performance of the proposed method. In this case, the input voltage is 15V and the output voltage is 5V. A resistive load of 5Ω is used for a load power of 5W. This setting emulates a typical 5V charging system for portable consumer electronics. Fig. 19 shows the variation of the energy efficiency with the conduction ratio. It is observed that an efficiency of 71% can be achieved. This energy efficiency is comparable with that of commercial Qi wireless charging systems at the rated power of 5W.
V. CONCLUSIONS

A charging time control method without the needs for mutual coupling information and RF communication system between the primary and secondary coils has been proposed for WPT systems. It consists of a primary-side input power control and a secondary-side output power hysteresis control. By using a hysteresis control to regulate the intermediate capacitor voltage in the secondary circuit and detecting the charging time of such an intermediate capacitor in the primary circuit, the charging time is used as control variable to vary the input power dynamically to meet the load demand. The proposed method overcomes many issues arising from coil misalignment between the transmitter and receiver coils. Because it does not need the precise relative positions of the transmitter and receiver coils, it is suitable for a wide range of WPT systems in which degree of free positioning of the receiver coils is allowed. Examples of such applications include wireless charging of electric vehicles, consumer electronics products and medical implants.

ACKNOWLEDGMENT

This work is supported by the Hong Kong Research Grant Council under the GRF project: HKU712913E. The authors are grateful to the University of Hong Kong for its support in the patent application of the proposed method [17].

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