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Transmitter Pulsation Control for Dynamic Wireless Power Transfer Systems

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Abstract—Wireless power transfer (WPT) is a convenient, flexible, and safe alternative to its wired counterpart. Due to these benefits, WPT has seen rapid growth in recent years. Typical WPT systems are usually used for static applications, where the load operates within a predefined area. On the other hand, dynamic WPT (D-WPT) is usually used to power loads which dynamically change their positions, and one of the ways to achieve this is by increasing the overall charging area via multiple transmitters (Tx). Multiple Tx's require communication and control, which increases the overall complexity of the system. This paper proposes a control algorithm which aims to maintain the overall efficiency of the D-WPT system by turning on and off the Tx when the mobile receiver (Rx) approaches and departs, respectively. Transient analysis of the D-WPT system is used to derive its control variables. In addition, energy and power loss of this proposed control algorithm are also investigated. The feasibility of the proposed control algorithm is demonstrated with 1-Tx and 3-Tx experimental setups. The experimental results validate the theoretical analysis model and the proposed control algorithm, and the standby (when not charging) input power forms 7.6% of the charging input power.

Index Terms—Dynamic charging, efficiency, transmitter (Tx) control, wireless power transfer (WPT).

I. INTRODUCTION

WIRELESS power transfer (WPT) has numerous advantages over its conventional wired counterpart, which include convenience, reliability, safety, power isolation, and ease of operation in hostile conditions. These benefits lead to the implementation of WPT in areas such as medical implants [1]–[3], transportation [4]–[6], material handling systems [7], electric vehicles [8]–[10], and mobile devices [11]–[13]. However, the WPT system has two key limitations: charging range and orientation. The dynamic WPT (D-WPT) system, therefore, aims to solve these issues by extending the coverage for the mobile receiver (Rx) to be wirelessly charged within its range of motion. D-WPT can

Fig. 1. Typical segmented D-WPT system.

be used in the above-mentioned WPT applications and has particularly gained traction in dynamic electric vehicles charging [14]–[16]. There are two main types of Tx magnetic coupler structures. The first structure type is track based, where the magnetic track spans the charging area, and the Rx is charged when it moves along the long magnetic track [4]. The simple design and minimal component count allow multiple pickups along the track. While the track is able to create a substantial charging space, a relatively larger Tx track means that the Rx merely covers a small portion of the Tx, leading to poor coupling levels. This, in turn, causes low efficiencies and electromagnetic interference (EMI) issues. The design proposed in [5] mitigated these effects by segmenting the Tx track into segments. Not only are the coupling levels improved, but also the electromagnetic fields of inactive segments are reduced significantly. These segments are powered by a single inverter that relies on sensors to turn ON or OFF the segments, and this causes the complexity of the D-WPT system to increase. Another structure type is segment based, which allows the individualized magnetic couplers to span the charging area and each magnetic coupler to be powered by an independent power source [14], [17], [18]. Fig. 1 shows the circuit of a typical segmented D-WPT system where there are n number of Tx and one mobile Rx. By individually controlling each magnetic coupler and its

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power source, a larger portion of the electromagnetic field created by the respective Tx can be captured by the Rx. This would increase the coupling level between a particular Tx and Rx, boosting its efficiency. However, the disadvantages include higher component and power electronics count. To add on, the overall efficiency and received power of the segmental systems are also extremely susceptible to misalignment, as the efficiency of the D-WPT system in [19] is reported to have varied from 85% to 5%. In order to smooth out the power flow during the dynamic charging process, [16] proposed to implement electrochemical capacitors to level out the currents on the transmitting and receiving sides while traversing from one Tx coil to another. Similar to the segmental track systems, sensors are needed to control the individual magnetic couplers and power sources for higher efficiencies, which again increase the system complexity. In order to reduce the required power converters of the segmental approach, Zhou and Mi [15] employed multiple LCC-compensated Txs that are connected to a single power converter. This is possible as the input power is automatically distributed among the LCC-compensated Txs according to their respective coupling levels. Furthermore, coil current regulation is proposed to regulate the otherwise constant coil current. While the EMI issue is mitigated by the dramatically reduced standby coil current, the standby (when not charging) input power forms 44.2% of the charging input power. This implies that the system is highly inefficient during standby. Also, the standby and charging input powers for both coil current-regulated and -nonregulated environments are increased. This would mean that the auxiliary circuit used for coil current regulation incurred additional power loss.

In this paper, a novel D-WPT Tx control scheme is proposed for a typical D-WPT system, as shown in Fig. 1, and has used a 3-Tx setup as reference. This control scheme would enable control of the D-WPT system without external position sensors, auxiliary circuits, or communications. The transient analysis of the input Tx current would form the basis of the control algorithm, where the main control variables are the current limits and pulse frequency. The performance metrics of the control algorithm, which are defined to be its energy and power losses, are also investigated.

This paper is organized as follows. Section II introduces the input current transient analysis of the series-compensated WPT system using fundamental circuit analysis. Section III presents the proposed control algorithm and introduces the control variables: current limits and pulse frequency. Using these variables, the flowchart of the control algorithm is presented. The energy and power losses incurred by this control algorithm are also evaluated. In Section IV, experimental verification of the proposed control algorithm is presented using the 1-Tx WPT system. A comparison of the 3-Tx WPT system with the proposed control algorithm is also provided. Last, conclusions are drawn in Section V.

II. TRANSIENT ANALYSIS

In this section, the transient analysis of the series-series (S-S) compensation circuit D-WPT system is derived. The analysis will be used as the basis for the proposed control algorithm in the next section.

Fig. 2. Series-series compensation circuit.

Fig. 3. Coil-to-coil efficiency versus coupling coefficient.

A. Circuit Analysis

The S-S compensation circuit given in Fig. 2 is used in this work, where capacitors (C_1 and C_R) are connected in series with the respective coils at both the transmitting and receiving sides. The S-S compensation circuit is used because the compensation capacitance required for both Tx and Rx is independent of the coupling. Assuming that the coils' skin and proximity effects are included in the respective coil resistances (R_1 and R_R) and the coils are mutually coupled (M), the system matrix in the frequency domain becomes

$$\begin{bmatrix} sL_1 + \frac{1}{sC_1} + R_1 & -sM \\ -sM & sL_R + \frac{1}{sC_R} + R_R + R_O \end{bmatrix} \begin{bmatrix} i_1 \\ i_R \end{bmatrix} = \begin{bmatrix} \frac{V_{in}\omega^2}{s^2 + \omega^2} \\ 0 \end{bmatrix} \quad (1)$$

where $s = j\omega$.

According to [20], the coil-to-coil efficiency can be derived from (1) to be

$$\eta = \frac{1}{1 + \frac{R_R}{R_o} + \frac{R_1}{R_o} \left(\frac{R_R + R_o}{\omega_0 M} \right)^2} \quad (2)$$

where ω_0 is the resonance frequency of the WPT system.

By optimizing both WPT's operating frequency and load resistance, and using measured coil resistances of $0.3 \, \Omega$ given in Table II, it is shown in Fig. 3 that the coil-to-coil efficiency decreases as the coupling decreases. This can be attributed to the rise of input power, which, in this case, is the input current, as the setup is powered by a constant voltage source. Therefore, the input current can be used as a control parameter for estimating the coupling coefficient.

TABLE I
SIMULATION PARAMETERS

Parameter	Value
Tx, Rx inductances, L_1, L_R (μH)	31.6
Tx resonance frequency (kHz)	992.5
Rx resonance frequency (kHz)	989
Coil Resistances, R_1 and R_R (Ω)	0.3
Load Resistance R_O (Ω)	20
Cut-off coil-to-coil efficiency (%)	60
Minimum mutual inductance, $M_{\min}$ (μH)	0.622

Fig. 4. Components of the input current in transient.

In order to investigate the input current, the steady-state and transient components of the input current i_1 are derived from (1). As the aim of the analysis is to study the transient response and not the coil efficiency and to ease derivation and computation, R_1 and R_R are neglected to simplify the analysis. Hence

$$i_1 = i_{1-SS} + i_{1-T} \quad (3)$$

$$i_{1-SS} = \frac{(\omega_0^2 C_1 C_R R_O / k^2) \omega_0}{s^2 + \omega_0^2} V_{in} \quad (4)$$

$$i_{1-T} = \frac{\alpha V_{in} [(C_R R_O (k^2 - 1)) s^2 + (k^2 - \omega_0^2 C_R^2 R_O^2) s - \omega_0^2 C_R R_O]}{k^2 [(1 - k^2) s^4 + \omega_0^2 C_1 R_O s^3 + 2\omega_0^2 s^2 + \omega_0^4 C_1 R_O s + \omega_0^4]} \quad (5)$$

where $k = M / \sqrt{L_1 L_R}$, $\alpha = \omega_0^3 C_1$.

From (4), the time domain of the steady-state component can be represented as

$$i_{1-SS}(t) = \frac{\omega_0^2 C_1 C_R R_O V_{in}}{k^2} \sin \omega_0 t. \quad (6)$$

However, as (5) does not have radical factors, the numerical expression of the transient component in time domain is

$$i_{1-T}(t) = N_1 e^{-\alpha_1 t} \cos(\omega_1 t + \theta_1) + N_R e^{-\alpha_R t} \cos(\omega_R t + \theta_R) \quad (7)$$

where the terms $N_{1,R}$, $\theta_{1,R}$, $\alpha_{1,R}$, and $\omega_{1,R}$ can be solved numerically. Using the circuit parameters in Table I, the input current response can be seen in Fig. 4, together with its steady-state and transient components. The frequency of the input current follows the steady-state component. Due to the active behavior of the D-WPT system, the coupling coefficient is expected to vary. In addition, depending on the type of

Fig. 5. Maximum input current versus coupling coefficient and load resistance.

application, the load resistance can also vary. With these variations, the maximum input current can be derived from (7), and is plotted in Fig. 5 against coupling and load resistance using the circuit parameters listed in Table I.

From Fig. 5, it is clear that the peak input current increases as the coupling coefficient decreases regardless of the load resistance. This important relation forms the basis of the control scheme, which will be explained in the next section

III. PROPOSED TX PULSATION CONTROL SCHEME

The analysis of the proposed control scheme consists mainly of three parts:

- 1) input current limit (which is derived from the minimum coupling coefficient $k_{\min}$, which in turn is derived from the cutoff coil-to-coil efficiency);
- 2) pulse frequency that defines the speed and responsiveness of the control scheme;
- 3) power loss of such control scheme.

To start off, we will cover the control algorithm flowchart. Following this, the theoretical analysis of the proposed control scheme and its important parameters will be covered and supplemented with simulation results.

A. Control Algorithm

A flowchart detailing the process of the proposed control algorithm flowchart is given in Fig. 6. At the start, we have to define the cutoff coil-to-coil efficiency for the respective current limits according to (9) and (10). Basically the operation of the proposed control scheme is as follows: the Tx is being powered ON briefly and only when the Tx and Rx are not within proximity (i_{1-ON} will be exceeded), then Tx will cease to operate and will restart again, depending on the pulse frequency. If both Tx and Rx are within proximity (i_{1-ON} will not be exceeded), then Tx will continue operation. When the Rx remains within the proximity of Tx, the Tx will continue operation but when the Rx moves further away from the Tx (i_{1-OFF} will be exceeded), the Tx will be turned OFF. The time differentiation between the two current limits t_1 is crucial for the proposed control scheme to maintain correctness. The pulse frequency (or its period T_{pulse}) is an important variable of this control algorithm as it determines the time that the WPT system starts up after being switched OFF.

B. Current Limits

In this section, we will derive the current limits (i_{1-ON} and i_{1-OFF}) and time t_1 to differentiate between the two current

Fig. 6. Proposed control algorithm flowchart.

Fig. 7. i_{1-ON} current normalized to i_{1-OFF} current.

limits. In order to determine the current limits, the cutoff coil-to-coil efficiency has to be defined. Using the predefined cutoff coil-to-coil efficiency η_{min} , the minimal mutual inductance M_{min} can be derived from (2) to be

$$M_{min} = \sqrt{\frac{\frac{\eta_{min} R_1}{R_O} \left(\frac{R_R + R_O}{\omega_0} \right)^2}{1 - \eta_{min} - \eta_{min} \frac{R_1}{R_O}}}. \quad (8)$$

The key parameters in this control scheme are the switching ON and OFF input current limits, i_{1-ON} and i_{1-OFF} , which are calculated from (7) based on the minimal coupling k_{min}

$$i_{1-ON} = i_{1-OFF} \sin(\omega_0 t) + N_1 e^{-\alpha_1 t_1} + N_2 e^{-\alpha_2 t_1} \quad (9)$$

$$i_{1-OFF} = \frac{\omega_0^2 C_1 C_R (R_R + R_O)}{k_{min}^2 + \omega_0^2 C_1 C_R R_1 R_R + \omega_0^2 C_1 C_R R_1 R_O} \quad (10)$$

where $k_{min} = M_{min} / \sqrt{L_1 L_R}$.

The difference between the two current limits is that i_{1-ON} may contain overshoot during its transient response compared to i_{1-OFF} , which only consists of the steady-state component. This is shown in Fig. 7, where the input current limit i_{1-ON} is normalized to i_{1-OFF} against the coupling coefficient. At lower couplings, the two current limits are similar but when a higher

Fig. 8. Time (t_1) of the input current's maximum amplitude versus coupling coefficient.

cutoff coupling is chosen, the disparity between both current limits increases.

Next, to differentiate between the two different current limits, we define t_1 as the time for the input current i_1 to reach its transient peak, and Fig. 8 shows the numerical solution of t_1 against the coupling coefficient. The selection of t_1 will affect the integrity of the proposed control scheme. If minimum values of t_1 are used, the two current limits will not be properly differentiated, and the control scheme might be erroneous. Therefore, to differentiate between the two current limits in (9) and (10), the highest t_1 according to k_{min} must be selected and doubled, so that the rising and falling times during the transient period are accounted for, thereby ensuring algorithm reliability. Next, if the Tx is switched OFF, the control scheme will restart operation according to the pulse frequency.

C. Pulse Frequency

The pulse frequency is defined as follows: assuming that the Tx is initially switched OFF, the time taken for the control scheme to restart its operation as described in previous section is determined by its pulse frequency. Therefore, the higher the pulse frequency, the lower will be the latency between each operation. However, the energy stored in the Tx during the transient period before exceeding i_{1-ON} has to be dissipated. If the time period between each pulse is too small for the WPT system to dissipate its energy before the next cycle, the resultant action of the control scheme can be erroneous.

The lowest energy dissipation rate occurs when the WPT system is uncoupled ($k = 0$), since the dissipation of energy will be isolated within the transmitting side only. Using the parameters from Table I and the uncoupled condition, we show the input current dissipation profile for the aforementioned worst case scenario in Fig. 9. As there is no finite solution to the dissipation current, the system is deemed to have dissipated its input current when the input current falls below 1% (as marked by the dotted line) of its current limit i_{1-ON} . In this case, according to Fig. 9, we take 1 ms as the upper limit of the pulse period (or 1 kHz to be the highest pulse frequency).

After determining the upper limit of the pulse frequency, both power and energy losses of this proposed control scheme can be calculated, as described in the next section.

D. Power and Energy Loss

According to the maximum pulse frequency derived in the previous section, the pulse frequency will be selected to

Fig. 9. (a) Input current dissipation profile under uncoupled condition. (b) Zoomed-in view of input peak current dissipation profile and dotted line shows the 1% mark.

Fig. 10. Energy loss of the proposed control algorithm versus coupling coefficient for various load resistances.

be 1 kHz in order to obtain the maximum power loss. The energy loss of the proposed control scheme is defined to occur during the time span when the Tx is briefly powered ON but is subsequently switched OFF as the Rx is not within the desired proximity of the Tx. The power loss arising from the normal operation of the WPT circuit is not factored in as it is independent of the proposed control scheme. The power loss is derived from (3) and the aforementioned time span, which is dependent on the input current transient response and i_{1-ON} from (9). The power loss of the proposed control scheme is defined as

$$P = f_{\text{pulse}} \int_0^{t_1} I_1 \cdot V dt. \quad (11)$$

With the power loss defined in (11) and time span, the energy loss of the proposed control scheme is formed in Fig. 10 to illustrate the impact of the coupling coefficient and equivalent load resistance on the proposed control scheme energy loss. Comparing the energy loss of a particular load resistance against the coupling coefficient, we can see that for coupling coefficients lower than $k_{\min}$, more energy loss is incurred when the coupling coefficient is nearer to the respective $k_{\min}$ (as the time taken for i_1 to exceed the current limit i_{1-ON} is significantly more) than when the WPT system is less coupled. It is also important to note that when comparing the energy losses between the various load resistances, the same current limit i_{1-ON} will result in a different $k_{\min}$. In addition, as suggested in [21], the WPT system's efficiency is dependent on the load resistance, which explains the increasing energy loss that comes along with the increasing load resistance.

Fig. 11. 1-Tx WPT system diagram.

In addition, (11) clearly shows the proportionality between pulse frequency and power loss. Having such a correlation, the pulse frequency can be chosen based on the nature of the application.

For low-speed applications such as mobile phone charging, a pulse frequency of 1 Hz would translate to a maximum charging latency of 1 s. For high-speed applications such as in-motion charging of electric vehicles, choosing a pulse frequency of 1 kHz would incur a displacement of 14 mm during its period (assuming typical urban speeds of 50 km/h), which is relatively small compared to typical vehicular Rx coils. In the following sections, 1 kHz will be used as the pulse frequency in the analyses.

IV. EXPERIMENTAL VERIFICATION

In this section, we will demonstrate the proposed control scheme via experimental verification. The experimental verification is divided into two sections, namely, 1-Tx setup and 3-Tx setup. The former is used to verify the transient analysis, control algorithm, and losses, while the latter is used to demonstrate the coil-to-coil efficiency improvements.

A. 1-Tx Setup

In order to verify the input current transient response and the proposed control algorithm, an experimental system has been implemented based on the 1-Tx WPT setup shown in Fig. 11. The two helical coils, Tx and Rx, create the displacements y_{rx} and z_{rx} according to their relative positions. The parameters of the 1-Tx WPT experimental setup are listed in Table II, and the simulation results are obtained from MATLAB/Simulink while the experiment is carried out using the Keysight E5061B ENA Series 2-port network analyzer (measured S-parameters are used to generate an experimental transient response using the Agilent ADS tool [22]), which is shown in Fig. 12.

First, for the experimental setup, z_{rx} is fixed at 20 cm while y_{rx} varies from 0 to 40 cm in intervals of 5 cm. The coil-to-coil efficiency against y_{rx} is shown in Fig. 13, presenting close comparison of experimental and simulation results. Next, the experimental and simulation input current transient responses (at $z_{rx} = 10$ cm and $y_{rx} = 0$ cm) are verified to match well in Fig. 14. The proposed algorithm is implemented in Agilent ADS and its cutoff coil-to-coil efficiency is set at 60%, which corresponds to a minimum mutual inductance of $0.496 \mu\text{H}$ according to (8). Using both (9) and the minimum

TABLE II
EXPERIMENTAL PARAMETERS

Parameter	Values
Tx inductance, L_1 (μH)	31.6
Rx inductance, L_R (μH)	31.6
Tx resonance frequency (kHz)	992.5
Rx resonance frequency (kHz)	989
Coil Resistances, R_1 and R_R (Ω)	0.3
Coils' radius (cm)	8.9
Load Resistance R_O (Ω)	20
Cut-off coil-to-coil efficiency (%)	60
Minimum mutual inductance, M_{min} (μH)	0.622
Gap distance, z_{rx} (cm)	10 & 20
Range of y_{rx} (cm)	0 to 40 (in intervals of 5)
Distance between Tx, y_{tx} (cm) [for 3-Tx]	30

Fig. 12. Experimental 1-Tx WPT system setup.

Fig. 13. Experimental and simulation coil-to-coil efficiency variation with respect to y_{rx} when $z_{rx} = 20$ cm.

mutual inductance, the input current switching ON limit i_{1-ON} can be calculated to be 0.73A. The proposed algorithm is implemented in Agilent ADS as shown in Fig. 15, where the excitation voltage is removed when the transient current exceeds the limit and then current dissipation occurs when $z_{rx} = 20$ cm and $y_{rx} = 15$ cm. This is in comparison with the input current transient response when $z_{rx} = 20$ cm and $y_{rx} = 10$ cm, where the input current does not exceed the limit and continues to operate normally. Last, for dissipation, the same limit i_{1-ON} of 0.73A is implemented for the least coupled position ($z_{rx} = 20$ cm and $y_{rx} = 40$ cm). The system

Fig. 14. Experimental and simulation input current transient responses when $z_{rx} = 10$ cm and $y_{rx} = 0$ cm.

Fig. 15. Experimental implementation of i_{1-ON} ($= 0.73\text{A}$) for $z_{rx} = 20$ cm and $y_{rx} = 10$ cm, and $z_{rx} = 20$ cm and $y_{rx} = 15$ cm.

Fig. 16. (a) Input current dissipation when $i_{1-ON} = 0.73$ A, and $z_{rx} = 20$ cm and $y_{rx} = 40$ cm. (b) Zoomed-in view (dotted line is 1% of current limit i_{1-ON}).

Fig. 17. Pulsing the WPT system at 1 kHz when $z_{rx} = 20$ cm and $y_{rx} = 15$ cm.

is dissipated within the period (1 ms) of the pulse frequency, as depicted by the transition in Fig. 16.

Fig. 17 shows multiple pulse excitations, where the WPT system ($z_{rx} = 20$ cm and $y_{rx} = 15$ cm) has a pulse frequency

Fig. 18. Experimental and analytical energy losses against y_{rx} when $z_{rx} = 20$ cm.

Fig. 19. Experimental and analytical normalized input power of the WPT system against y_{rx} when $z_{rx} = 20$ cm.

of 1 kHz. As the input current exceeds the predefined i_{1-ON} , the input voltage is removed before being applied again at the onset of the next pulse period. When the input current exceeds i_{1-ON} , the WPT system's input energy is considered as loss energy, whereas when i_{1-ON} is not exceeded, the system continues operation and there is no energy loss arising from this control algorithm. Fig. 18 shows the comparison of experimental and analytical energy loss against y_{rx} when $z_{rx} = 20$ cm. Multiplying the energy loss by the highest pulse frequency, the highest input power during the standby mode (when $z_{rx} = 20$ cm and y_{rx} is between 15 and 40 cm) is about 14.5 mW. On the other hand, according to Fig. 19, when the Rx is aligned to the Tx ($z_{rx} = 20$ cm and $y_{rx} = 0$ cm), the WPT system's input power is 0.191 W. Therefore, the proposed control algorithm has a lower standby-to-charging-mode input power percentage of 7.6% compared to the 44.2% reported in [15].

B. 3-Tx Setup

The proposed control scheme is applied to the 3-Tx WPT system, and its experimental setup and circuit diagram are shown in Figs. 20 and 21, respectively. Compared to the 1-Tx WPT system, two additional Txs are added to the 3-Tx WPT system and a 4-port Rohde and Schwarz ZVB 8 Vector Network Analyzer is used in this experiment. The Txs are connected individually to the vector network analyzer's ports. Also, z_{rx} is changed from 20 to 10 cm in order to maintain operation continuity within the span of y_{rx} . Without the proposed control scheme, the 3-Tx systems are kept switched ON throughout. Coil-to-coil efficiency dips significantly when the Rx is aligned to Tx₃, which corresponds to $y_{rx} = 30$ cm. This is due to the low coupling coefficient between both

Fig. 20. 3-Tx WPT experimental setup.

Fig. 21. 3-Tx WPT system diagram.

Fig. 22. Experimental coil-to-coil efficiency results for 3-Tx WPT system ($y_{tx} = 30$ cm, $z_{rx} = 10$ cm, and $y_{rx} = 0$ to 30 cm) using proposed control algorithm.

transmitters (Tx₁, Tx₂) and Rx, which increased Tx₁ and Tx₂ input currents, and also overall input power. Employing the proposed control scheme, the 3-Tx WPT coil-to-coil efficiency is compared in Fig. 22, where η_{min} is kept at 60%. First, when $y_{rx} = 0$ to 10 cm, only Tx₂ is switched ON. In the middle ($y_{rx} = 15$ cm), both Tx₂ and Tx₃ are switched ON. At the end ($y_{rx} = 20$ to 30 cm), only Tx₃ is switched ON. The same switching order can be said for the other side of the 3-Tx setup when y_{rx} is negative. As shown in Fig. 22, by controlling the respective Tx using the proposed control scheme, the coil-to-coil efficiency of the D-WPT system can be improved. With this control algorithm in place, coil-to-coil efficiency improvements of up to 52% (when $y_{rx} = 30$ cm) can be attained and the coil-to-coil efficiency of the WPT system can be expected to operate above the cutoff coil-to-coil efficiency while enabling dynamic control of the WPT system without any secondary controller or communication.

The proposed control algorithm is employed in an S-S-compensated WPT system, and it can be employed on other systems of different compensation circuits and operating frequencies using the analysis and control algorithm in this paper accordingly. Also, the power level can be easily increased by increasing the voltage levels in proper manner.

V. CONCLUSION

This paper proposed a control algorithm for D-WPT system on the primary side to control the transmitter(s) in order to improve the overall coil-to-coil efficiency. The main control variable, which is the Tx's input current, has been theoretically analyzed and modeled. The other important aspects of the algorithm such as current limits, pulse frequency, and power loss have been analyzed. Using these control parameters, a 1-Tx WPT experimental system was used to demonstrate the feasibility of the proposed control algorithm. The control algorithm was also implemented on a 3-Tx experimental setup, and it shows significant overall coil-to-coil efficiency improvements of up to 52%.

REFERENCES

- [1] K. M. Silay, C. Dehollaini, and M. Declercq, "Inductive power link for a wireless cortical implant with two-body packaging," *IEEE Sensors J.*, vol. 11, no. 11, pp. 2825–2833, 2011.
- [2] A. K. Ram Rakhyani, S. Mirabbasi, and M. Chiao, "Design and optimization of resonance-based efficient wireless power delivery systems for biomedical implants," *IEEE Trans. Biomed. Circuits Syst.*, vol. 5, no. 1, pp. 48–63, Feb. 2011.
- [3] R.-F. Xue, K.-W. Cheng, and M. Je, "High-efficiency wireless power transfer for biomedical implants by optimal resonant load transformation," *IEEE Trans. Circuits Syst. I, Reg. Papers*, vol. 60, no. 4, pp. 867–874, Apr. 2013.
- [4] J. Huh, S. W. Lee, W. Y. Lee, G. H. Cho, and C. T. Rim, "Narrow-width inductive power transfer system for online electrical vehicles," *IEEE Trans. Power Electron.*, vol. 26, no. 12, pp. 3666–3679, Dec. 2011.
- [5] J. Shin *et al.*, "Design and implementation of shaped magnetic-resonance-based wireless power transfer system for roadway-powered moving electric vehicles," *IEEE Trans. Ind. Electron.*, vol. 61, no. 3, pp. 1179–1192, Apr. 2014.
- [6] Z. Li, K. Song, G. Wei, J. Jiang, and C. Zhu, "A 3-kW wireless power transfer system for sightseeing car supercapacitor charge," *IEEE Trans. Power Electron.*, vol. 32, no. 5, pp. 3301–3316, May 2016.
- [7] J. T. Boys, G. A. Covic, and A. W. Green, "Stability and control of inductively coupled power transfer systems," *IEE Proc.-Electr. Power Appl.*, vol. 147, no. 1, pp. 37–43, Jan. 2000.
- [8] J. M. Miller, O. C. Onar, and M. Chinthavali, "Primary-side power flow control of wireless power transfer for electric vehicle charging," *IEEE J. Emerg. Sel. Top. Power Electron.*, vol. 3, no. 1, pp. 147–162, Jan. 2014.
- [9] S. Li and C. C. Mi, "Wireless power transfer for electric vehicle applications," *IEEE J. Emerg. Sel. Topics Power Electron.*, vol. 3, no. 1, pp. 4–17, Mar. 2015.
- [10] H. H. Wu, A. Gilchrist, K. D. Sealy, and D. Bronson, "A high efficiency 5 kW inductive charger for EVs using dual side control," *IEEE Trans. Ind. Informat.*, vol. 8, no. 3, pp. 585–595, Aug. 2012.
- [11] B. Choi, J. Nho, H. Cha, T. Ahn, and B. Choi, "Design and implementation of low-profile contactless battery charger using planar printed circuit board windings as energy transfer device," *IEEE Trans. Ind. Electron.*, vol. 51, no. 1, pp. 140–147, Feb. 2004.
- [12] Y. Jang and M. M. Jovanovic, "A contactless electrical energy transmission system for portable-telephone battery chargers," *IEEE Trans. Ind. Electron.*, vol. 50, no. 3, pp. 520–527, Jun. 2003.
- [13] C.-G. Kim, D.-H. Seo, J.-S. You, J.-H. Park, and B. H. Cho, "Design of a contactless battery charger for cellular phone," *IEEE Trans. Ind. Electron.*, vol. 48, no. 6, pp. 1238–1247, Dec. 2001.
- [14] J. M. Miller, P. T. Jones, J. M. Li, and O. C. Onar, "ORNL experience and challenges facing dynamic wireless power charging of EV's," *IEEE Circuits Syst. Mag.*, vol. 15, no. 2, pp. 40–53, Feb. 2015.
- [15] S. Zhou and C. Mi, "Multi-paralleled LCC reactive power compensation networks and their tuning method for electric vehicle dynamic wireless charging," *IEEE Trans. Ind. Electron.*, vol. 63, no. 10, pp. 6546–6556, Oct. 2016.
- [16] J. M. Miller *et al.*, "Demonstrating dynamic wireless charging of an electric vehicle: The benefit of electrochemical capacitor smoothing," *IEEE Power Electron. Mag.*, vol. 1, no. 1, pp. 12–24, Mar. 2014.
- [17] J. P. K. Sampath, D. M. Vilathgamuwa, and A. Alphones, "Efficiency enhancement for dynamic wireless power transfer system with segmented transmitter array," *IEEE Trans. Transport. Electrification*, vol. 2, no. 1, pp. 76–85, Jan. 2015.
- [18] F. Lu, H. Zhang, H. Hofmann, and C. Mi, "A dynamic charging system with reduced output power pulsation for electric vehicles," *IEEE Trans. Ind. Electron.*, vol. 63, no. 10, pp. 6580–6590, Oct. 2016.
- [19] O. C. Onar, J. M. Miller, S. L. Campbell, C. Coomer, C. P. White, and L. E. Seiber, "A novel wireless power transfer for in-motion EV/PHEV charging," in *Proc. Conf. Proc. IEEE Appl. Power Electron. Conf. Expo. (APEC)*, Apr. 2013, pp. 3073–3080.
- [20] B. X. Nguyen *et al.*, "An efficiency optimization scheme for bidirectional inductive power transfer systems," *IEEE Trans. Power Electron.*, vol. 30, no. 11, pp. 6310–6319, Nov. 2015.
- [21] B. X. Nguyen, G. Foo, A. Ong, D. M. Vilathgamuwa, and U. K. Madawala, "Efficiency optimization for bidirectional IPT system," in *Proc. IEEE Transp. Electrification Conf. Expo*, Sep. 2014, pp. 1–5.
- [22] J. Kim, D.-H. Kim, and Y.-J. Park, "Analysis of capacitive impedance matching networks for simultaneous wireless power transfer to multiple devices," *IEEE Trans. Ind. Electron.*, vol. 62, no. 5, pp. 2807–2813, May 2015.

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