Development of Design Methodology for 60 Hz Wireless Power Transmission System

Hiroki Ishida* Member, Hiroto Furukawa** Member
Tomoaki Kyoden*** Non-member

(Manuscript received Feb. 3, 2016, revised May 6, 2016)

We previously reported a 60 Hz wireless power transmission (WPT) system, which is a system that uses the common utility frequency. In the study reported in paper, we solved several issues in order to install this system in a small electric vehicle. First, an accelerated finite difference time domain (FDTD) method using a graphics processing unit was developed to solve the issue of computation time. Next, theoretical equations for the transmission efficiency (\( \eta \)) and power (\( P_{\text{out}} \)) that include the stray load loss were derived from an equivalent circuit analysis. A new device was designed based on these theoretical equations, where by \( \eta = 70\% \) and \( P_{\text{out}} = 451 \text{ W} \) were achieved for a transmission distance of 150 mm. Finally, we attempted to wirelessly charge of a lead storage battery. The overall efficiency of the wireless charging system was maintained at 60\% during battery charging.

Keywords: wireless power transfer, magnetic resonance, utility frequency, wireless charging

1. Introduction

From the time that resonance wireless power transmission (WPT) was first developed by the WiTricity project (1), there has been a great deal of research conducted using this and other approaches (2–4). In addition, the direction of research has progressed towards the design of devices such as inverters and control techniques (5–8). Wireless charging of electric vehicles is a typical application of WPT (9–10).

A large value for the product of the coupling coefficient (\( k \)) and the quality factor (\( Q \)) yields a high transmission efficiency. Therefore, to obtain high efficiency, even when \( k \) is small (i.e., transmission distance is long), a large \( Q \) is required. A high-frequency power supply is thus used; however, it is necessary to consider the skin effect and proximity effect to reduce the winding resistance (11). The capability of the high-frequency system is very high and a total efficiency of 77\% at a distance of 300 mm in free space has been achieved (12) (efficiency between two coils is 95\% at a distance of 300 mm (12)). Wireless charging of electric vehicles could be achieved by embedding the transmitter coil into paved roads or car parking areas. Omar et al. reported successful 23 kHz WPT of over 2 kW with a dc-to-dc efficiency of over 75\% at a transmission distance of 100 mm, and also evaluated the power loss in concrete block (14). We have proposed the use of low frequency such as the common utility frequency to eliminate power loss in concrete (15). As another advantage, no high-frequency power supply unit would be required. There have been no attempts to use the common utility frequency after the development of the WiTricity system. We are aware of only one report from the University of California over 30 years ago, where an efficiency of 60\% was achieved over a distance of 100 mm using a 400 Hz power supply (16).

In our previous report, we achieved a power efficiency of 78\% and 165 W at a distance of 100 mm (63\% and 69 W at 150 mm), and also confirmed that energy loss into the concrete does not occur. Although optimization has not yet been performed, the 60 Hz-WPT system is expected to have higher potential. The design has not yet been optimized, because we have not completed a theoretical analysis of the common utility frequency WPT system. The efficiency-equation in the case of high frequency does not require core loss to be considered, because there is no advantage in using a magnetic core. However, a magnetic core is a necessary component for low frequency. In addition, although the stray load loss cannot be neglected, as with an induction motor having a gap between the rotor and stator, an equation that contains the stray load loss has not been presented until now.

Although transmission efficiency is strongly dependent on the shape of the magnet pole piece, a simulator has not been developed to determine the optimal shape. Accurate predictions of the transmission efficiency and power become possible if these issues are solved, and indicate that a 60 Hz-WPT system has high capability.

2. Experimental Procedure

The 60 Hz-WPT devices were produced with different magnet-pole shapes. The magnet-pole pieces were fabricated.
from 0.35 mm thick silicon steel plates containing 3.5% silicon. At a frequency of 60 Hz, no saturation of the pole pieces was observed up to a flux density of 0.7 T. The coils were wound from single-strand enamel-covered copper wire with a diameter of 2 mm. Experimental results from four devices (P1-P4) are presented in this paper. Although experimental results for P1-P3 have already been published previously (15), they are presented here for comparison with the calculation results. The specifications for these three devices have been previously reported and are thus omitted in this paper.

The transmission efficiency and power were predicted through analysis of the equivalent circuit for the 60 Hz-WPT system. The parameters of the equivalent circuit used in the calculation (i.e., the transformer constants) were determined experimentally using an actual WPT device. The transformer constants were measured using a frequency response analyzer (FRA5097, NF Co., Ltd.) and a programmable AC power source (EC1000SA, NF Co., Ltd.). To determine the magnetic permeability, an inductor with a closed-loop core was required, which was also fabricated in addition to the four WPT devices.

### 3. GPGPU-FDTD Simulation

The finite element method (FEM) for frequency-domain analysis and the finite difference time domain (FDTD) method for time-domain analysis are often used for electromagnetic field simulation (17)–(20). FEM is an indirect method for solving Maxwell equations that employs weighted residuals. In contrast, FDTD is a simple method for solving Maxwell equations by transforming them into difference equations. The FDTD method is used for numerical analysis of high-frequency WPT (21)–(23). However, Maxwell equations and Berenger’s perfectly matched layer (PML) can also be used for low-frequency analysis (24). Therefore, in the case of a low-frequency WPT analysis, there is no theoretical problem. We have previously reported the results of an electromagnetic field analysis using FDTD (15), where typical computing using a single-threaded program was performed. The calculation could not be performed using the actual size, because a large computation time is necessary for low-frequency analysis using FDTD. Therefore, there was a significant gap between what was ideal and practically possible. We therefore attempted to create a program using the parallel distribution processing method with general-purpose computing on graphics processing units (GPGPU) to significantly increase the processing speed.

Specifications and analysis conditions for the 60 Hz-WPT simulation are given in Table 1. We used a GeForce GTX 980 Ti (NVIDIA Co., Ltd.) as the GPU, and it was overclocked at up to 1.418 GHz from 1.0 GHz using ASUS GPU Tweak software. The number of CUDA cores for this device is 2861. Therefore, the processing speed reached 8.0 TFPS for single-precision floating point operations. Used Visual C++ 2013 and Cuda Toolkit 6.5 as our simulator development environment.

Yee cell sizes of the two dimensional (2D) analysis plane were \(\Delta x = 1.0\ mm\) and \(\Delta y = 1.0\ mm\). The time step was \(2.35 \times 10^{-13}\ s\), which meets the Courant-Friedrichs-Lewy condition (25). Berenger’s perfectly matched layer (PML) was used for the absorption of electromagnetic waves at the end of the analysis plane.

Calculations using the actual size and frequency could thus be performed. 2D magnetic field is expressed as a contour graph of magnetic flux density \(B\). Figures 1(a)–(c) show simulation results for the open state of the secondary coil for P1, P2, and P3 for a transmission distance \(\delta\) of 100 mm. The transmitter was placed on the lower side and the receiver on the upper side in the analysis plane. Specification of these devices have been shown in our previous paper (15). The initial relative permeability of silicon steel was taken to be 2000. The conductivity of the copper winding was \(59 \times 10^6\ S/m\). The input current of the primary coil was 10 A RMS. There are no needs to specify the self-inductance and load resistance because these are estimated by the simulator.

These contour graphs were obtained at 1/4 periods (4.17 ms) after the power supply was turned on. We have confirmed that the steady state was reached at 4.17 ms by comparison with simulation results using the quasi-static FDTD method (15). As an example, the results shown in Fig. 1(a) took 79 hours of computation time. The field is not high, even at the tip of the magnet-poles, which suggests that no abnormally high core loss occurs. \(k\) was estimated from the ratio of the magnetic flux \(\Phi_1\) generated in the primary coil to the interlinkage flux \(\Phi_2\), which is directed into the secondary coil. To compare the experimental and calculation results, Fig. 2 shows the variation in \(k\) with transmission distance for the three devices. The experimental data for P1 to P3 have already been published in a previous paper (15). Although the predicted \(k\) values are slightly different from the experimental results, the magnitude relationship of the three devices is consistent with the experimental results.

### 4. Equivalent Circuit Analysis

Condensers function to compensate leakage inductance. Three modes of connection were performed; the primary \(C_1\)
and secondary $C_2$ condensers were connected in parallel with the respective coils (PP mode), $C_1$ was connected in parallel while $C_2$ was connected in series (PS mode), and $C_1$ and $C_2$ were both connected in series (SS mode) (27). The equivalent circuit analysis process for the PP mode is given here, and a similar process was applied for the other modes. The equivalent circuit for the PP mode is shown in Fig. 3. To derive an equation for the theoretical transmission efficiency ($\eta$), an analysis method was constructed based on that of Tohi et al., in which the copper loss of the WPT system is considered (28).
Z = \left( \frac{x_L}{x_L + x_2} \right)^2 R_L + j \left( \frac{x_L x_1 + x_1 x_2 + x_2 x_L}{x_L + x_2} \right) \cdots (2) \\
Z \equiv R + jX

When C_1 is connected, the overall impedance Z' is expressed as

Z' = \frac{\omega C_1}{\omega^2 C_1 + j(XC_1 - X - x_C)} \cdots (3)

The condition of C_1, for which the imaginary part of Z' becomes zero, is given by

x_C = (\frac{R}{X})^2 + (\frac{X}{x_C})^2
\cdots (4)

The relationship between V_L and V_2 is expressed as

V_L = \frac{R_L(x_L + x_2)}{R_L x_L + j x_2(x_L + x_2)} V_2 \cdots (5)

The relationship between V_2 and V_{IN} is given as

V_2 = \frac{(x_L + x_2)^2}{Z} V_{IN} \cdots (6)

Thus, the relationship between V_L and V_{IN} is expressed as

V_L = \frac{R_L(x_L + x_2)}{R_L x_L + j x_2(x_L + x_2)} \frac{R_{IN} x_L x_2}{(x_L + x_2)^2} V_{IN} = \frac{V_{IN}}{Z} \frac{R_L x_L}{x_L + x_2} \cdots (7)

The relationship between I_1 and I_L is derived by dividing both sides of Eq. (7).

\frac{V_L}{R_L} = I_L = \frac{V_{IN}}{Z} \frac{x_L}{x_L + x_2} = I_1 \frac{x_L}{x_L + x_2}, \cdots (8)

where I_1 and I_L are in phase.

The relationship between I_L and I_2 is given as

I_L = \frac{x_2^2 - j R_L x_2 I_2}{R_L^2 + x_2^2} \cdots (9)

The absolute value of I_L is expressed as

[I_L] = \sqrt{\frac{x_2^2}{R_L^2 + x_2^2} + \left( \frac{R_L x_2}{R_L^2 + x_2^2} \right)^2} = \frac{x_2}{\sqrt{R_L^2 + x_2^2}} |I_L|, \cdots (10)

The phase difference between I_1 and I_2 is given by

I_2 = \frac{1}{\alpha} I_1 = \frac{x_2^2 - j R_L x_2}{R_L^2 + x_2^2} I_2, \cdots (11)

I_2 = \alpha \frac{x_2^2 - j R_L x_2}{R_L^2 + x_2^2} I_2 = \alpha \left( \frac{x_2^2}{R_L^2 + x_2^2} - j \frac{R_L x_2}{R_L^2 + x_2^2} \right) I_2, \cdots (12)
The following relationship is derived from Eq. (16):

$$\eta = \frac{R_L I_1^2}{R_L I_1^2 + r_s I_1^2 + r_L I_1^2 + (r_2 + r_s) I_2^2}$$

The relationship between $I_0$, $I_1$, and $I_2$ is expressed as:

$$I_0 = I_1 - I_2$$

$$|I_0|^2 = |I_1|^2 + |I_2|^2 - 2 |I_1||I_2| \cos \phi$$

$$L_1 \left[ \alpha^2 + 1 + \left( \frac{R_L}{x_c} \right)^2 \right] - 2 \alpha N \left[ \frac{R_L}{x_c} \right] \cos \phi.$$

For the equivalent circuit shown in Fig. 3, $\eta$, with consideration of the copper, core, and stray load losses is given as:

$$\eta = \frac{R_L I_1^2}{R_L I_1^2 + r_s I_1^2 + r_L I_1^2 + (r_2 + r_s) I_2^2}$$

The value of $R_L$ for which the copper loss is minimized is given as:

$$R_L = x_c \sqrt{\frac{r_1}{r_2} + 1} \cdot \frac{x_c}{r_2}.$$

Thus, the maximum transmission efficiency ($\eta_{\text{max}}$) with consideration of the copper, core, and stray load losses is given as:

$$\eta_{\text{max}} = \frac{1}{1 + \frac{2 (\sqrt{\alpha^2 + 1} - \alpha)}{\sqrt{\alpha^2 + 1}} + \frac{\sqrt{\alpha^2 + 1} - \alpha}{\sqrt{\alpha^2 + 1}} + \frac{\sqrt{\alpha^2 + 1} + \alpha}{\sqrt{\alpha^2 + 1}}}. \cdot \frac{1}{\alpha^2 + 1}.$$

Here, $k$ and the quality factors for the two coils (Q1 and Q2) are defined as:

$$Q_1 = \frac{\omega L_1}{r_1}, \quad Q_2 = \frac{\omega L_2}{r_2}, \quad k = \frac{M}{\sqrt{L_1 L_2}}, \quad M = \frac{x_c}{x_1 + x_2}.$$

The following relationship is derived from Eq. (16):

$$\alpha^2 = \frac{r_1}{r_2} = \frac{L_1^2 r_1}{M^2 r_2} = \frac{L_1 L_2 Q_1}{M^2 Q_2} = \frac{1}{k^2 Q_1}.$$  \hspace{1cm} (17)

$\eta_{\text{max}}$ in Eq. (15) can thus be rewritten using $k$ and $Q$:

$$\eta_{\text{max}} = \frac{1}{1 + \frac{2 (\sqrt{\alpha^2 + 1} - \alpha)}{\sqrt{\alpha^2 + 1}} + \frac{\sqrt{\alpha^2 + 1} - \alpha}{\sqrt{\alpha^2 + 1}} + \frac{\sqrt{\alpha^2 + 1} + \alpha}{\sqrt{\alpha^2 + 1}}}. \cdot \frac{1}{\alpha^2 + 1}.$$

The phase difference of Eq. (11) was also adopted the condition of Eq. (14) for minimum copper loss:

$$\cos \phi = \frac{x_c}{\sqrt{R_L^2 + x_c^2}} = \frac{1}{\sqrt{1 + \frac{x_c^2}{Q_1^2}} + 2}.$$  \hspace{1cm} (19)

Substituting Eq. (18) into Eq. (19) yields

$$\eta_{\text{max}} = \frac{1}{1 + \frac{2 \sqrt{Q_1 Q_2}}{Q_1} + \frac{1}{Q_1} + \frac{r_s (a^2 + 1)}{Q_1} + \frac{r_s (b^2 + 1)}{Q_1} \cdot \sqrt{2}}.$$

Equation (21) is true under any conditions; therefore, Eq. (20) can be approximated as shown by Eq. (22).

$$\eta_{\text{max}} \approx \frac{1}{1 + \frac{2 \sqrt{Q_1 Q_2}}{Q_1} + \frac{1}{Q_1} + \frac{r_s (a^2 + 1)}{Q_1} + \frac{r_s (b^2 + 1)}{Q_1} \cdot \sqrt{2}}.$$

When $Q_1$ and $Q_2$ are almost the same, $\alpha \approx \frac{1}{2}$ and $\frac{Q_1}{Q_2} \approx 1$ hold, so that

$$\eta_{\text{max}} \approx \frac{1}{1 + \frac{2 \sqrt{Q_1 Q_2}}{Q_1} + \frac{1}{Q_1} + \frac{r_s (a^2 + 1)}{Q_1} + \frac{r_s (b^2 + 1)}{Q_1} \cdot \sqrt{2}}.$$

To estimate the stray load loss resistance ($r_s$) is difficult. When $\delta$ is short, the stray load loss may be acceptably negligible. The magnetic field surrounding the magnetic pole expands into free space with increasing $\delta$; therefore, $r_s$ cannot be ignored in this case, and $r_s$ will be strongly dependent on $\delta$. Furthermore, $r_s$ will be dependent on the shape of the magnet-pole pieces, but this has not yet been investigated. We have assumed that $r_s$ is zero to investigate the effects of stray load loss, and then compared the experimental and calculated results. Figure 4(a) shows the calculation results with consideration of the copper and core losses. Experimental data have already been published in previous paper. Here, the core loss resistance ($r_s$) is a constant value and it can be easily identified by employing a measurement method similar to that for a power transformer. In the region where $\delta$ is short, $r_s$ can be ignored. The discrepancy between the two results increases with increasing $\delta$. Therefore, $r_s$ was calculated as a function of $\delta$ to compensate for the discrepancy with a best-fit polynomial curve. However, only one term of fourth order could be fitted, and the results are shown in Fig. 4(b). There is good agreement between the experimental and calculated values in all regions. The following best-fitting functions of $r_s$ were obtained for the three devices:

P1 (rectangular): $r_s (\delta) = 15 \times \delta^4 \cdot (10^{-9} \Omega)$, \hspace{1cm} (24-1)

P2 (double flare): $r_s (\delta) = 4.5 \times \delta^4 \cdot (10^{-9} \Omega)$, \hspace{1cm} (24-2)

P3 (single flare): $r_s (\delta) = 0.9 \times \delta^4 \cdot (10^{-9} \Omega)$. \hspace{1cm} (24-3)

Here, the unit for $\delta$ is millimeters. In the case of an induction motor, $r_s$ is a constant value because the gap between the stator and rotor is fixed. Identification of the $r_s$ value from theoretical calculation is difficult, even though it is a constant value. Here, the stray load loss was estimated as the remaining power loss after subtraction of the copper, core and mechanical losses from the total power loss. The $r_s$ value is equivalent to division of the stray load loss by the square of the load current. Thus, the method used here to identify the
Fig. 4. Dependence of $\eta$ on $\delta$, with consideration of (a) copper and core losses, and (b) copper, core, and stray load losses

Fig. 5. Correlation between $k$ and the $\delta^4$ coefficients

The transmission power ($P_{out}$) at maximum efficiency indicated in Eq. (25) is given as

$$P_{\text{out}} = R_L I_i^2 \frac{V_{IN}^2}{Z''} \eta_{\text{max}},$$  \hspace{1cm} (25)$$

where $V_{IN}$ is 100 or 200 V in Japan. $Z''$ is the overall impedance of the equivalent circuit in the state with condensers connected:

$$Z'' = \frac{x_{C1}^2 R}{R^2 + (X - x_{C1})^2},$$  \hspace{1cm} (26)$$

where

$$R \equiv \left( \frac{x_L}{x_L + x_2} \right)^2 R_L, \quad X \equiv \frac{x_L x_1 + x_1 x_2 + x_2 x_L}{x_L + x_2}.$$  \hspace{1cm} (27)$$

Here, the $R_L$ value was entered for the optimal conditions shown in Eq. (14).

Furthermore, $x_1$ and $x_2$ fulfilled the respective resonance conditions of Eqs. (4) and (1). The difference between the experimental and the calculated results is shown in Fig. 6. Experimental data have already been previously published\(^{(15)}\). There is good agreement in the long $\delta$ region. However, for the short distance region, there is poor agreement with the experimental results. The leakage inductances $x_1$ and $x_2$ become small in the short $\delta$ region, so that the relative contribution of the winding resistance cannot be neglected. Having ignored the effect of the winding resistance in Eq. (26), $Z''$ is estimated to be larger than the true value in the short $\delta$ region. To mitigate the mismatch, the equivalent circuit containing resistance components should be solved precisely; however, this has not yet been completed. The conclusion can be derived from Eq. (25), where the simple solution to obtain large $P_{\text{out}}$ is to apply a large input voltage, because $P_{\text{out}}$ is proportional to the square of the input voltage.

5. Design and Fabrication

Our purpose is to achieve wireless charging for a small electric vehicle (EV). The 60 Hz-WPT system is designed to allow wireless charging at home without the need for a high-frequency power supply. We have developed a small EV equipped with the 60 Hz-WPT, and photographs of the prototype are shown Fig. 7. The small EV has a 48 V electrical system by the in-series connection of four 12 V lead storage battery packs. Table 3 shows the requirements of the 60 Hz-WPT charging system. The device was designed in an attempt to satisfy these requirements and evaluation was conducted using equivalent circuit analysis (PP and SS mode).
60 Hz Wireless Power Transmission System (Hiroki Ishida et al.)

Table 3. Requirements of the 60 Hz-WPT system

<table>
<thead>
<tr>
<th>Transmission distance (δ)</th>
<th>100 mm</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage</td>
<td>100 V RMS @ 60Hz</td>
</tr>
<tr>
<td>Maximum output power</td>
<td>800 W</td>
</tr>
<tr>
<td>Maximum current of primary coil</td>
<td>10 A RMS</td>
</tr>
<tr>
<td>Weight</td>
<td>&lt; 10 kg</td>
</tr>
</tbody>
</table>

Fig. 7. Small EV equipped with the 60 Hz-WPT charging system. (a) Back view, and (b) internal components

Table 4. Specifications for the P4 prototype

<table>
<thead>
<tr>
<th>Parameters of coils for δ = 10 cm</th>
<th>Coupling coefficient (k)</th>
<th>Core loss resistance (r_c)</th>
<th>Stray load loss resistance (r_s)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Q factor at 60 Hz</td>
<td>61.1 (1)</td>
<td>0.239</td>
<td>0.05 Ω</td>
</tr>
<tr>
<td>Winding resistance</td>
<td>0.74 Ω (1)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Self-inductance</td>
<td>119.9 mH (1)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Coupling coefficient (k)</td>
<td>0.239</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Core loss resistance (r_c)</td>
<td>0.60 Ω</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Stray load loss resistance (r_s)</td>
<td>0.05 Ω</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Fig. 8. (a) Front and (b) side views of the P4 prototype

Table 4. Specifications for the P4 prototype

<table>
<thead>
<tr>
<th>Specifications</th>
<th>Weight</th>
<th>Magnetic pole area</th>
<th>Cross-sectional area</th>
<th>Number of turns (layers)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Improved single flare (P4)</td>
<td>10.0 kg</td>
<td>132 cm²</td>
<td>28 cm²</td>
<td>459 turns (6)</td>
</tr>
</tbody>
</table>

Figure 8 shows a photograph of the P4 prototype, and the specifications for P4 are given in Table 4. P4 was designed to be 2 kg lighter than P3, and an efficiency of 80% was theoretically predicted with δ = 100 mm. Here, r_c could not be identified by calculation; therefore, the values measured for P3 were used.

The size of the device was determined based on the design guidelines for a low-voltage transformer for converted to resistance at 75°C. According to the guidelines, the cross-sectional area required for a single-strand copper wire was over 3.0 mm². Therefore, a polyvinyl-formal enamel-coated copper wire with a diameter of 2.0 mm was selected. In a typical low-voltage transformer, the sectional area of the magnetic core is determined assuming a magnetic field of up to approximately B = 1.5 T. However, a larger cross-sectional area was required to obtain a large Q. To obtain Q > 60, a cross-sectional area of at least 28 cm² was necessary. A magnetic pole area of 132 mm² was necessary to obtain k = 0.22 at δ = 10 cm.

In the experiment, a sliding resistance (DW series, Yabashi Electric Co., Ltd.) was used as an experimental load. For example, the inductance included in the winding of 6.3 Ω was 0.5 mH, which is negligibly small compared to the self-inductance of the primary and secondary coils. The resistance change due to temperature increase was adjusted manually to keep constant. The experimental results, where the efficiency and power output were 79% and 190 W, confirm that the P4 prototype exhibits the designed performance. Figures 9(a) and (b) show the η and P_out characteristics for the P4 prototype, and the results for P3 are also presented for comparison. At δ = 100 mm, the difference between the two devices appears small; however, the difference increased with δ. At 150 mm, η was improved by 7%, and P_out was increased by 44 W. However, the output power of the 60 Hz-WPT system would be too small in consideration of applications such as small electric vehicles. For example, the high-frequency WPT system for the Nissan Leaf II has over 3 kW output at a transmission distance of 150 mm. Although the exact size and weight are not published, the latter system has overwhelming performance compared to the 60 Hz-WPT system. If applied to a small electric vehicle, the 60 Hz-WPT system would require 500 W or more.\(^{33}\)

We considered that high power can be obtained even at low input voltage by lowering the overall impedance. Thus, a simple solution is to select the SS mode. Figure 10(a) shows a comparison of the output power with input voltage for the three modes (PP, PS, and SS modes). For the SS mode, the output power reached 854 W at 100 V. We did not attempt to increase the input voltage over 100 V, because this would exceed the rated current of the condenser. Figure 10(b) shows a comparison of the transmission efficiency for the
60 Hz Wireless Power Transmission System (Hiroki Ishida et al.)

Fig. 9. Characteristics of the P4 prototype. (a) Dependence of $\eta$ on $\delta$, and (b) of $P_{\text{out}}$ on $\delta$. Results for the P3 prototype are also shown for comparison.

Three modes. For the SS mode, the maximum efficiency of 70% was maintained until 451 W. The weight of 10 kg would be too heavy for small electric vehicle. Therefore, further weight reduction is required by optimization of the design.

6. Wireless Charging System

Wireless charging of lead storage batteries was performed in anticipation of wireless charging for a small EV. Figure 11 shows the wireless charging circuits. The experiment was performed using the P4 device with a transmission distance of 15 cm. The capacitances of the condenser for resonance ($C_1$, $C_2$) were given by the following equation from the equivalent circuit analysis:

$$C_1 = \frac{1}{\omega_0(x_L + x_1)}$$
$$C_2 = \frac{1}{\omega_0(x_L + x_2)}$$

(28)

$60\mu F$ film condensers were selected according to Eq. (28). Two lead storage batteries (Type: 44B19R, 12 V, 34 Ah) connected in series were used as the load. The internal resistance in a lead storage battery changes during charging, and it also varies depending on the charge time. The amount of change in the battery used for the experiment was between 20 and 35 m$\Omega$.

A diode bridge (GBP2504) with a large internal resistance was selected for the rectifier. The internal resistance was 180 m$\Omega$, which is the average dynamic resistance in the operating voltage region of $V_F$ from 0.0 V to 0.93 V. The total load resistance was approximately 400 m$\Omega$ when the line resistance of 10 m$\Omega$ was considered. Therefore, the variable ratio of the load resistance caused by the internal resistance of the battery would be only 1.8%. Therefore, control to track the load variation was not necessary (as shown in Fig. 7(b)).

The output voltage of wireless charging system was 31 V under 50 V of input voltage, which is reasonable because the typical charging voltage per battery pack is 15.5 V. The common utility voltage in Japan is 100 V; therefore, the input voltage is just 100 V in the case of four battery packs connected in series. Our small electric vehicle has been developed with four battery packs connected in series.

Figure 12 shows a time course of the voltage and current during battery charging. The WPT system in the SS mode has immittance conversion characteristics. Common utility power acts roughly as a constant voltage supply, so that the WPT system in the SS mode acts almost as a constant current supply. Therefore, the charging current continues to flow, even if a charging voltage of 31 V is reached, which takes approximately 2 h.

Figure 13 shows the time course of the transmission efficiency during charging. Here, $\eta_{\text{wpt}}$ is $P_{\text{out}}/P_{\text{in}}$ ($\times$ 100%), which is the efficiency of the WPT system. $\eta_{\text{total}}$ is $P_{\text{out}}/P_{\text{in}}$ ($\times$ 100%), which is the efficiency including the power loss in the rectification circuit. Although the circuit was uncontrolled during charging, $\eta_{\text{total}}$ was kept at 60%, which is
The optimum load resistance for the SS mode was derived from equivalent circuit analysis, as follows:

\[ R_{L,max} = k \tau \sqrt{Q_1Q_2}, \]  

which is 6.3 Ω for the P4 device with a transmission distance of 15 cm. The transmission efficiency under optimum load resistance was determined using the same formula as Eq. (23). The experimental results also confirmed that the maximum efficiency (70%) is obtained when \( R_L = 6.3 \Omega \). We consider that the 4% decrease is caused by the difference between the actual load resistance (0.4 Ω) and the optimum load resistance. The following conclusions were made from calculation of Eq. (13): the copper loss in the secondary coil will increase when the load resistance deviates from the optimum value, while the percentages of other loss elements are not significantly changed. An increase in the current of the secondary coil was also confirmed in the experiment. The rate of increase in copper loss from the secondary coil was 4% when connected to the batteries, as shown in Fig. 11.

7. Conclusion

The accelerated FDTD simulation using GPGPU was confirmed to be useful for performance design. \( \eta \) and \( P_{out} \) can be estimated according to Eqs. (23) and (25), respectively. However, the cross-sectional area and number of coil turns were fixed when determining the magnet pole shape; therefore, a new simulator that integrates the equivalent circuit analysis and FDTD simulation is required to obtain maximum transmission efficiency.

The P4 prototype achieved an efficiency of 70% and an output of 451 W at a distance of 150 mm, which constitutes a 7% improvement in \( \eta \) and an increase of 382 W for \( P_{out} \) over our previous report. Even with a lead storage battery as a load, the decrease from the maximum efficiency is only 4%.

The device proposed here is very simple and easily fabricated with only silicon steel plates, single-strand copper wire, and phase-advanced condensers. In addition, this system could be directly connected to a wall socket. The manufacturing cost of the P4 prototype was approximately 600 USD (silicon steel plates: 100 USD, copper wire: 100 USD, and condensers: 400 USD). Therefore we consider that the 60 Hz-WPT is suitable for wireless charging of small electric vehicles.

Acknowledgment

This research is partially supported by the Japan Science and Technology Agency A-STEP, (Feasibility studies stage No.AS262Z00347L), and JSPS KAKENHI Grant Number 15K13934.

References


Hiroki Ishida (Member) received the Ph.D. degree in electrical engineering from Nagoya University of Technology, Niigata, Japan, in 2004. He joined the National Institute of Technology, Toyama College, in 2005, and has been an associate professor there since 2010. In 2015, he moved to Okayama University of Science as an associate professor in the Department of Applied Physics. His current research interests include wireless power transmissions. He is a member of the Institute of Electrical Engineers of Japan and the Japan Society of Applied Physics.

Hiroto Furukawa (Member) received the B.S. degree and M.S. degree in electrical engineering from Tokyo Denki University, Tokyo, Japan, in 1987 and 1989, respectively. He joined the National Institute of Technology, Toyama College, in 1991, and has been an associate professor there since 2003. He has been engaged in propagation characteristics of microwave in concrete blocks and its application for non-destructive diagnostics. He is a member of the Institute of Electrical Engineers of Japan and the Institute of Electronics, Information and Communication Engineers.

Tomoaki Kyoden (Non-member) received the Ph.D. degree in mechanical Engineering from Kanazawa University, Ishikawa, Japan, in 2014. He joined the National Institute of Technology, Toyama College, in 2010 and has been a lecturer there since 2016. His current research interests include focus on energy conversion high-efficiency technology and utilization of renewable energy. He is a member of the Japan Society of Applied Physics.