Load and Mutual Inductance Identification Method for Series-Parallel Compensated IPT Systems

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Abstract

Identifying the load and mutual inductance is essential for improving the power transfer capability and power transfer efficiency of Inductive Power Transfer (IPT) systems. In this paper, a steady-state load and mutual inductance identification method focusing on series-parallel compensated IPT systems is proposed. The identification model is established according to the steady-state characteristics of the system. Furthermore, two sets of identification results are obtained, and then they are analyzed in detail to eliminate the untrue one. In addition, the identification method can be achieved without extra circuits so that it does not increase the complexity of the system or the control difficulty. Finally, the feasibility of the proposed method has been verified by simulation and experimental results.

Key words: Inductive power transfer, Load identification, Mutual inductance identification, Resonant, Wireless power transfer

I. INTRODUCTION

With the continuous development and progress of wireless power transfer (WPT) technologies, inductive power transfer (IPT) technology has attracted a lot of attention recently [1], [2]. Due to the advantages of low maintenance cost, high reliability, and flexible features, IPT systems have been widely applied in many occasions such as transportation [3]-[5], electronic products [6], [7], biomedical electronics [8], [9], and so on.

In IPT systems, the energy receiving side can be moved freely from the energy transmitting side, which leads to variations of the mutual inductance. In addition, different types of energy receiving equipment present different load characteristics. Taking the wireless power supplies for electric vehicles as an example, different electric vehicles have different power requirements and different mutual inductances on account of different heights and chassis alignments which may change the air gap between the transmitting coil and the receiving coil. Due to the inevitable variations mentioned above in practical IPT systems, the reflecting impedance from the receiving side to the transmitting side varies. Furthermore, the variation of the reflecting impedance results in a resonant frequency drifting, which dramatically reduces the power transfer capability and system efficiency [10]. Additionally, in order to ensure efficient and reliable power transfers, the output power of the system should be adjusted since the load change reflects the change in the actual power requirements. Therefore, load and mutual inductance identification is essential for IPT systems and is a key issue to improve the transfer efficiency and the capability of IPT systems [11]-[15].

Previously, a few load identification methods have been researched [11]-[13], [15]. These methods achieve load identification when the mutual inductance is regarded as a constant parameter. Reference [14] achieves load and mutual inductance identification by switching an additional capacitor in the transmitting side. However, this method increases the complexity of the system and the control difficulty. Series-parallel compensated IPT systems have been discussed by a lot of scholars [16]-[18] and they have been applied in many fields such as biomedical implants [19] and battery charging [20] due to the current source characteristic of the parallel-compensated secondary side. However, existing load and mutual inductance identification methods mostly focus on series-series compensated IPT systems. Therefore, a steady-
Fig. 1. Main circuit of a series-parallel compensated IPT system.

A state load and mutual inductance identification method for series-parallel compensated IPT systems is proposed in this paper. Only the information of the input voltage, the output current of the inverter and the system operating frequency are required in this method. Furthermore, this identification method can be achieved without any extra circuits. As a result, the cost and complexity of system is decreased.

II. FUNDAMENTAL ANALYSIS

A. Basic Circuit Topology

The topology of a series-parallel compensated IPT system as a commonly-used compensation scheme is shown in Fig. 1.

A typical IPT system consists of two independent mutually coupled coils in the transmitting and receiving side of the system. In this system, a DC voltage source $E_{dc}$ supplies the power. A full-bridge inverter network is composed of four MOSFET switches, $S_1$, $S_2$, $S_3$ and $S_4$. The invert network generates a modulated ac voltage injected into the transmitting resonant network. A transmitting coil $L_p$ in series with the compensating capacitor $C_p$ constitutes a series resonant tank.

In the receiving side, the receiving coil $L_s$ receives power from a high-frequency magnetic field and constitutes the parallel resonant network with the compensating capacitor $C_s$. Finally, the receiving side supplies power to the equivalent load $R$ via the resonant network. $R_p$ and $R_s$ are the inherent resistances of $L_p$ and $L_s$, respectively. $M$ is the mutual inductance between the transmitting and receiving coils.

In order to achieve a maximum power transfer with a minimum VA rating of the supply, the system should operate at or near the inherent resonant frequency of the transmitting and receiving side. Thus, it is a logical choice to design the two inherent resonant frequencies at or near the same frequency $\omega_0$, which satisfies the relationship:

$$\omega_0^2 \approx \frac{1}{L_p C_p} \approx \frac{1}{L_s C_s}$$

(1)

B. The equivalent circuit model

In the transmitting side, a combination of the coil and the compensating capacitance can be regarded as a low-pass filter, and the high frequency harmonic currents in the resonant network are suppressed. Therefore, only the fundamental harmonic is considered for power transfer based on Fourier decomposition.

The output voltage of the full-bridge inverter is an approximately square wave with the amplitude of $E_{dc}$. Its fundamental harmonic $u_p$ can be expressed as:

$$u_p(t) = \frac{4E_{dc} \sin \omega t}{\pi}$$

(2)

The root-mean-square (RMS) value of $u_p$ is obtained by:

$$U_p = \frac{2\sqrt{2} E_{dc}}{\pi}$$

(3)

Therefore, the equivalent circuit model can be acquired, as illustrated in Fig. 2.

III. LOAD AND MUTUAL INDUCTANCE IDENTIFICATION

The output current $i_p$ of the inverter can be expressed as:

$$i_p(t) = \sqrt{2} I_s \sin (\omega t + \phi)$$

(4)

where, $I_p$ is the RMS of $i_p$, and $\phi$ is the phase difference between $u_p$ and $i_p$.

From Fig. 2, the KVL equation is derived as:

$$\begin{align*}
\begin{bmatrix}
i_p \\
i_s
\end{bmatrix} &= 
\begin{bmatrix}
Z_{11} & Z_{12} \\
Z_{21} & Z_{22}
\end{bmatrix}^{-1}
\begin{bmatrix}
U_p \\
0
\end{bmatrix}
\end{align*}$$

(5)

where

$$\begin{align*}
Z_{11} &= R_p + j \left( \omega L_p - \frac{1}{\omega C_p} \right) \\
Z_{12} &= j \omega M \\
Z_{21} &= R_s + j \omega L_s + R \parallel \frac{1}{j \omega C_s}
\end{align*}$$

(6)

Based on Eqn. (5), the expressions of $I_p$ is conducted:

$$I_p = \begin{bmatrix}
U_p & Z_{22} \\
0 & Z_{21}
\end{bmatrix}
\begin{bmatrix}
Z_{11} \\
Z_{22}
\end{bmatrix}$$

(7)

Then, the input impedance can be derived from Eqn. (7):

$$\frac{U_p}{I_p} = \alpha_i + \frac{\alpha_i^2 + \beta_i^2 + j \left( \beta_i - \beta_s \gamma_s \right)}{\alpha_s^2 + \beta_s^2}$$

(8)

Where
\[
\begin{align*}
\alpha_i &= R_p, \\
\alpha_z &= R_s + \frac{R}{1 + \omega^2 C^2 R^2}, \\
\beta_i &= \omega L_p - \frac{1}{\omega C_p}, \\
\beta_z &= \omega L_s - \frac{\omega C_s R^2}{1 + \omega^2 C_s^2 R^2}, \\
\gamma &= \omega M \tag{9}
\end{align*}
\]

Generally, in order to minimize the VA ratings of the power supply, it is desirable to make the system operate at the zero phase angle frequency. At this frequency, the input voltage \(u_p\) and input current \(i_p\) should be in phase, which means that \(\phi\) is equal to zero and the following relationships can be obtained based on Eqn. (8).

\[
\beta_i - \beta_z = 0 \tag{10}
\]

\[
\frac{U_p}{I_p} = \frac{U_p}{I_p} = \alpha_i + \frac{\alpha_i \beta_i^2}{\alpha_i^2 + \beta_i^2} \tag{11}
\]

Obviously, information on the equivalent load \(R\) is included in \(\alpha_i\) and \(\beta_z\), while information on the mutual inductance \(M\) is included in \(\gamma\). Consequently, two equations for the equivalent load \(R\) and the mutual inductance \(M\), described in Eqs. (10) and (11), are obtained. Then, the identification results of the equivalent load \(R\) can be obtained by combining and solving the above equations.

\[
R_i = \frac{-1 - \sqrt{1 - 4c(R_c + a - abc)(R_s - ab)}}{2c(R_c + a - abc)} \tag{12}
\]

\[
R_s = \frac{-1 + \sqrt{1 - 4c(R_c + a - abc)(R_s - ab)}}{2c(R_c + a - abc)} \tag{13}
\]

Moreover, the coefficients are given as follows:

\[
\begin{align*}
\alpha &= \frac{U_p}{I_p} - \alpha_i I_p, \\
\beta &= \omega L_s, \\
c &= \omega C_s \tag{14}
\end{align*}
\]

It is found that there are two solutions for \(R\). Obviously, one of them is an untrue identification result while the other is a reasonable solution. Therefore, in order to eliminate the untrue solution, a series of analysis is carried out.

To simplify the following analysing, a normalized angular frequency \(\omega_n = \omega / \omega_0\) is defined, and the parameters \(\beta_i\), \(\beta_z\), and \(\gamma\) in Eqn. (9) are rewritten by substituting Eqn. (1).

\[
\begin{align*}
\alpha_i &= R_p, \\
\alpha_z &= R_s + \frac{R}{1 + \omega^2 C^2 R^2}, \\
\beta_i &= \frac{\omega^2 - 1}{\mu \omega}, \\
\beta_z &= \frac{\omega^2 + \lambda \omega (\omega^2 - 1) R^2}{\lambda (1 + \lambda \omega^2 R^2)} \\
\gamma &= \omega^2 \omega_0 M \tag{15}
\end{align*}
\]

Based on Eqs. (15) and (16), the expressions of \(R_i\) and \(R_s\) can be rewritten as:

\[
\begin{align*}
R_i &= \frac{1 + \sqrt{\Delta}}{2\lambda\omega_0\left((\omega_0^2 - \lambda)R_s - a\right)}, \tag{17} \\
R_s &= \frac{1 - \sqrt{\Delta}}{2\lambda\omega_0\left((\omega_0^2 - \lambda)R_s - a\right)} \tag{18}
\end{align*}
\]

where

\[
\Delta = 1 - 4\omega_0\left((\omega_0^2 - \lambda)R_s - a\right)(\omega_0^2 - \lambda R_s) \tag{19}
\]

It should be noted that the rewritten expressions of \(R_i\) and \(R_s\) are based on Eqn. (16) as mentioned before. However, \(\omega_0 C_p\) and \(\omega_0 C_s\) are approximately equal to \(1/\omega_0 L_p\) and \(1/\omega_0 C_s\) in actual systems. Therefore, the identification results of the equivalent load \(R\) shown in Eqs. (12) and (13) are more accurate. The expressions of \(R_i\) and \(R_s\) shown in Eqs. (17) and (18) are only used for the subsequent analysis.

According to Eqs. (17) and (18), it is found that the value of the equivalent load \(R\) is correlated with \(\omega_n\). Therefore, in order to obtain the actual identification result of the equivalent load, the ranges of \(R_i\) and \(R_s\) are discussed in three cases, namely \(\omega_n < 1\), \(\omega_n = 1\), and \(\omega_n > 1\).

A. \(\omega_n < 1\)

In this case, it is found that the value of \(R_i\) is positive, while the value of \(R_s\) is negative, which are depicted in Fig. 3 and Fig. 4, respectively. In addition, \(K\) represents the ratio between \(U_p\) and \(I_p\). Obviously, \(R_i\) is the identification result of the equivalent load \(R\).

B. \(\omega_n = 1\)

When \(\omega_n = 1\), the following equation can be achieved by the transformation of Eqn. (10).
Fig. 4. Range of $R_2$ versus $K$ at different values of $\omega_n$ when $\omega_n < 1$.

\[ \frac{\beta_2}{\alpha_2^2 + \beta_2} \omega_n^2 M^2 = 0 \]  \hspace{1cm} (20)

where

\[ \begin{align*}
\alpha_2 &= R_s + \frac{R}{1 + \lambda^2 R^2} \\
\beta_2 &= \frac{1}{\lambda (1 + \lambda^2 R^2)}
\end{align*} \]  \hspace{1cm} (21)

Obviously, $\beta_2$ is greater than 0. Therefore, Eqn. (20) is valid only when $M = 0$. This means that the energy receiving part is completely removed.

C. $\omega_n > 1$

The coupling coefficient $k$ of the system is defined as:

\[ k = \frac{M}{\sqrt{L_p L_s}} \Rightarrow M = k \sqrt{L_p L_s} \]  \hspace{1cm} (22)

Eqn. (22) is substituted into Eqn. (10) to obtain the expression of $k$ below.

\[ k = \frac{1}{\omega_n} \sqrt{\frac{\mu \lambda \beta_1 (\alpha_2^2 + \beta_2^2)}{\beta_2}} \]  \hspace{1cm} (23)

Making

\[ \frac{d}{dR} k = 0 \]  \hspace{1cm} (24)

Then, the value of $R_0$ corresponding to the maximum coupling coefficient can be obtained as:

\[ R_0 = \frac{R}{\omega_n^2 - 1} \]  \hspace{1cm} (25)

The coupling coefficient $k$ is monotonically increasing when the load $R$ is in the range of $(0, R_0)$, and it decreases with a monotonically increasing $R$ when $R$ is larger than $R_0$. Therefore, the maximum value and the minimum value of $k$ can be expressed as follows:

\[ \begin{align*}
{k_{\text{max}}} &= k|_{R=R_0} = \sqrt{\frac{\omega_n^2 + \lambda^2 R_s^2 - 1}{\omega_n^2}} \\
{k_{\text{min}}} &= k|_{R=\infty} = \sqrt{\frac{\omega_n^4 + (\lambda^2 R_s^2 - 2) \omega_n^2 + 1}{\omega_n^4}}
\end{align*} \]  \hspace{1cm} (26)

Based on Eqn. (26), the curves of $k_{\text{max}}$ and $k_{\text{min}}$ for various values of $\omega_n$ can be plotted as shown in Fig. 5. Typically, the coupling coefficient $k$ is in the range of 0.1–0.3 [21]. Under this condition, the range of the normalized angular frequency $\omega_n$ can be calculated according to Fig. 5.

\[ \{ \omega_n > 1 \} \bigcap \left\{ \frac{\omega_n^2 + \lambda^2 R_s^2 - 1}{\omega_n^2} \leq 0.1 \right\} \bigcap \left\{ \frac{\omega_n^4 + (\lambda^2 R_s^2 - 2) \omega_n^2 + 1}{\omega_n^4} \leq 0.3 \right\} \]  \hspace{1cm} (27)

Consequently, the range of the load identification results can be obtained when $\omega_n$ satisfies the relationship given in Eqn. (27), which is illustrated in Fig. 6. It can be seen that: 1) both $R_1$ and $R_2$ are positive; 2) $R_1$ is larger than $R_{\text{uni}}$ and $R_2$ is less than $R_{\text{uni}}$. From Eqns. (17) and (18), a unique solution $R_{\text{uni}}$ exists when $\Delta$ is equal to zero. In addition, in this case, $R_1$ and $R_2$ are equal to $R_{\text{uni}}$.

A plot of $R_{\text{uni}}$ at various normalized angular frequencies is
shown in Fig. 7. It can be seen that $R_{\text{uni}}$ decreases with an increasing $\omega_n$. The minimum value of $R_{\text{uni}}$ is larger than 100Ω. In most applications, such as battery charging systems, the equivalent resistance is typically just a few tens of ohms [20], [22]. Therefore, $R_2$ is considered as an identification result of equivalent load.

In conclusion, the identification result of the equivalent load $R$ can be expressed as follows:

$$R = \begin{cases} R_1 & \omega_n < 1 \\ \text{None} & \omega_n = 1 \\ R_2 & \omega_n > 1 \end{cases}$$  \hspace{1cm} (28)

The expressions of $R_1$ and $R_2$ are given in Eqns. (12) and (13). Furthermore, the identification results of the mutual inductance $M$ can be obtained based on Eqn. (10).

$$M = \begin{cases} 1 & \omega_n < 1 \\ \frac{\beta_1(\alpha_1^2+\beta_2^2)}{\beta_1} & \omega_n \neq 1 \\ 0 & \omega_n = 1 \end{cases}$$  \hspace{1cm} (29)

According to Eqns. (28) and (29), the identification results are calculated based on the circuit parameters (namely $L_p$, $L_s$, $C_p$, $C_s$, $R_p$, $R_s$). This means that the results are influenced by variations of these parameters. In general, under normal circuit operating conditions, the circuit parameters do not change much. Therefore, these parameters are generally assumed to be constant [11]-[15]. Then, the identification results can be obtained by detecting the input voltage $E_{\text{dc}}$, the output current $I_p$ of the inverter and the system operating frequency $f$ when the system is in operation.

IV. SIMULATION AND EXPERIMENTAL STUDIES

A. Parameters and Structure of the Proposal

In order to verify the effectiveness of the proposed method, a simulation model based on MATLAB/Simulink is set up with reference to Fig. 1. The main parameters are the same in both the simulation and experimental systems, and they are listed in Table I.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transmitting coil $L_p$ (μH)</td>
<td>152.87</td>
</tr>
<tr>
<td>Receiving coil $L_s$ (μH)</td>
<td>153.60</td>
</tr>
<tr>
<td>Compensation capacitance $C_p$ (nF)</td>
<td>68.26</td>
</tr>
<tr>
<td>Compensation capacitance $C_s$ (nF)</td>
<td>67.73</td>
</tr>
<tr>
<td>Inherent resistance of transmitting coil $R_p$ (Ω)</td>
<td>0.49</td>
</tr>
<tr>
<td>Inherent resistance of receiving coil $R_s$ (Ω)</td>
<td>0.54</td>
</tr>
</tbody>
</table>

![Fig. 8. Structure of the proposed identification method.](image)

A structure diagram of the proposed identification method is shown in Fig. 8. The zero phase angle condition is achieved by the frequency adjustment based on the phase detection of the output current and voltage of the inverter. The values of $E_{\text{dc}}$, $I_p$ and $f$ are measured by the detection unit including the DC voltage measurement, the current RMS sampling and the operating frequency detection, respectively. Then, the measured data are transferred to the identification unit and the values of the load resistance and the mutual inductance are calculated by the algorithm depicted in Section III.

B. Simulation Results

The total simulation time is set at 0.2ms and the system has already reached the steady-state. The maximum simulation step is 0.1μs. The identification accuracy of the load resistance $R$ and the mutual inductance $M$ are presented in Fig. 9 and Fig. 10, when the mutual inductance varies from 15μH to 45μH and the load resistance variation ranges from 10Ω to 50Ω. It can be seen that the identification accuracy of $R$ and $M$ are more than 96% and 97%, when the mutual inductance $M$ is 15μH. Moreover, the identification accuracies of $R$ and $M$ increase to about 98% and 98.5%, when the mutual inductance increases to 20μH. The identification accuracy is slightly lower when the mutual inductance is too small.

Furthermore, the simulated values of $\omega_n$ are illustrated in Fig. 11. It can be seen that $\omega_n$ increases slightly with a growing $M$ and that the variation of $R$ has little effect on $\omega_n$. When $M$ and $R$ are set at 15μH and 50Ω, the minimum value of $\omega_n$ ($\omega_n=1.0051$) is achieved. In addition, through simulations, it is found that $\omega_n$ is less than 1 when the
Identification Accuracy of \( R \) (%)

Fig. 9. Simulated identification accuracy of the load resistance \( R \).

Identification Accuracy of \( M \) (%)

Fig. 10. Simulated identification accuracy of the mutual inductance \( M \).

coupling coefficient is very low \((k<0.01)\). For example, the simulated values of the load resistance and the mutual inductance are set at 20Ω and 1μH \((\omega_n<1)\). In addition, the identification results are 18.46Ω and 1.17μH with accuracies of 92.3% and 83%, respectively. However, as mentioned above, the coupling coefficient \( k \) typically ranges from 0.1 to 0.3 \((\text{i.e. the mutual inductance of this system ranges from approximately } 15\mu\text{H to } 46\mu\text{H})\). It can be considered that \( \omega_n>1 \) is the most ordinary operating condition of IPT systems. Therefore, the simulations and the experiments in this paper are focusing on the case of \( \omega_n>1 \).

Altogether, the simulation results verify that the proposed identification model is effective for identifying the load resistance and the mutual inductance.

C. Experimental Results

A prototype of the series-parallel compensated IPT system, as shown in Fig. 12, was built and practically tested. The full-bridge inverter circuit is made up of four MOSFETs \((\text{STP30NF20})\). A FPGA chip \((\text{Altera Cyclone II EP2C5T144C8})\) is selected as a controller of the system, which calculates the values of the load resistance and the mutual inductance.

Table II shows the identification results of \( R \) and \( M \) when the load resistance is fixed at 30Ω and the mutual inductance varies from 15μH to 45μH. Actually, the mutual inductances are 15.06μH, 20.48μH, 25.66μH, 30.51μH, 35.63μH and 40.08μH.

### Table II

<table>
<thead>
<tr>
<th>( R )</th>
<th>( M )</th>
<th>Identified ( R )</th>
<th>Identified ( M )</th>
</tr>
</thead>
<tbody>
<tr>
<td>30Ω</td>
<td>15.06μH</td>
<td>31.28Ω</td>
<td>15.61μH</td>
</tr>
<tr>
<td>30Ω</td>
<td>20.48μH</td>
<td>28.94Ω</td>
<td>20.98μH</td>
</tr>
<tr>
<td>30Ω</td>
<td>25.66μH</td>
<td>30.89Ω</td>
<td>26.22μH</td>
</tr>
<tr>
<td>30Ω</td>
<td>30.51μH</td>
<td>30.77Ω</td>
<td>29.82μH</td>
</tr>
<tr>
<td>30Ω</td>
<td>35.63μH</td>
<td>29.01Ω</td>
<td>36.47μH</td>
</tr>
<tr>
<td>30Ω</td>
<td>40.08μH</td>
<td>30.71Ω</td>
<td>40.98μH</td>
</tr>
</tbody>
</table>
and they are changed by adjusting the relative position between the transmitting coil and the receiving coil.

It is found that identification accuracy grows according to increases in the mutual inductance. The accuracy drops to the lowest value when the mutual inductance is set to 15.06 μH, which agrees well with the simulation results. The maximum identification error of \( R \) is 4.27%, and the maximum identification error of \( M \) is 3.65%.

Similarly, experiments with a fixed mutual inductance of 30.51 μH and load resistances changing from 10Ω to 50Ω have been performed. The load resistance is regulated by pressing the buttons shown in Fig. 12. The identification results of \( R \) and \( M \) are illustrated in Table III. The maximum identification error of \( R \) is 5.50%, and the maximum identification error of \( M \) is 4.42%.

In addition, more experiments were performed when the mutual inductance was in the range of 15μH–46μH. According to the obtained results, the identification accuracies of \( R \) and \( M \) can both be maintained over 90% when the load resistance ranges from 8Ω to 95Ω. The ranges of \( R \) and \( M \) can be expanded by sacrificing the identification accuracy. The feasibility of the proposed identification method has been verified with different loads and varying mutual inductances.

### V. CONCLUSION

In this paper, a load and mutual inductance identification method for series-parallel compensated IPT systems is proposed. Only information on the input voltage, the output current of inverter and the system operating frequency are required in this method. The simulation results and the set values were in close agreement. Experimental results showed that the identification accuracy is over 90% when the load resistance varies from 8Ω to 95Ω, and that the mutual inductance is in the range of 15μH–46μH. Both the simulation and experimental results have verified the feasibility of the proposed method. Although only the series-parallel compensation scheme is discussed in this paper, the proposed method has some reference value for other types of secondary parallel-compensated IPT systems. Moreover, the influence of some of the intrinsic factors, such as variation of the circuit parameters, on the identification results will be considered in future research.

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