Modeling and Decoupled Control of Inductive Power Transfer to Implement Constant Current/Voltage Charging and ZVS Operating for Electric Vehicles

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ABSTRACT To prevent degradation of battery life and reduce switching loss in inductive power transfer (IPT) systems for electric vehicles (EVs), it is vital to realize constant current/voltage (CC/CV) charging and zero-voltage-switching (ZVS) operation. However, due to the existence of couplings between the charging and ZVS operating control-loops, the controllers are neither optimal nor robust. In this paper, a model-based decoupled control method is proposed to improve the performance of the charging and ZVS operating controllers. Since the operating point of the CC and CV modes are different, two dynamic models for the corresponding charging modes are built and linearized. Besides, the couplings between the charging and ZVS operating control-loops are analyzed, and in turn, decoupled by designing compensators. To optimize the system performance, we chose PI controllers, and their parameters are obtained by calculating the settling time and overshoot of the close-loop dominant poles according to the control performance requirements. In the experiment, a 500W series–series IPT system prototype is built and tested to verify the improvement of the control performance. Compared to the coupled control method and the traditional closed-loop control feedback method, the recorded response curves and transient waveforms demonstrate that the proposed decoupled control method brings better performance to the system.

INDEX TERMS inductive charging system; modeling; zero voltage switching; decoupled control;

I. INTRODUCTION Inductive power transfer (IPT), or wireless power transfer (WPT) is a highly-regarded technology that makes the energy transfer process safer, more reliable, and more user-friendly [1] [2]. For the application of the wireless charging systems (WCS) for electric vehicles (EVs), since the power can be directly picked up from the parking spaces or roads, the WCS are considered as one of the practicable methods to increase the competitiveness of EVs for thermal vehicles [4]-[7].

In order to reduce the switching loss and to prevent the degradation of the battery’s lifespan, the WCS is usually expected not only to provide constant current/voltage (CC/CV) charging [6] [7] but also to realize zero-voltage-switching (ZVS) operation [7] [8] [10]. However, there is a common challenge for the WCS, that is, the CC/CV charging and ZVS operating are easily fluctuated by the changes in the battery resistance during the charging process [5]-[7], which leads to neither optimal nor robust charge process for EVs. Therefore, numerous controllers or strategies in previous works are designed to mitigate the perturbation.

The proportional–integral (PI) control method is adopted in [7] and [10]-[14], while phase-locked-loop (PLL) [8] [10], proportional–differential (PD) controller [13], algorithms [14] [16] and sliding-mode controller [18][19] are developed to regulate the charging current, voltage and ZVS angle. In these studies, the duty cycle [7] [8] and the operating frequency [7]-[10] of the inverter are usually selected as the controlled quantities to form a charging and a ZVS operation control-loops, respectively. Although a ZVS angle [7] [10] can be accurately calculated by considering the whole time
delays synthetically, the parameters of the PI and PLL controllers are not optimized, resulting in a volatile control performance of the output current or voltage. Considering that the primary current and the secondary output voltage are easily adjusted by the power source and DC-DC converters [11]-[18], the PI or PD controllers [11]-[14], customized algorithms [14] [16] and sliding-mode controller [18] are used to control the outputs of the pulse width modulation (PWM) regulators or active rectifiers. To tune the parameters of the controllers, an empirical Ziegler–Nichols (ZN) method is adopted in [11] [12]. Furthermore, the above researchers mainly focus on the steady state of the system rather than the dynamic performance like overshoot and settling time. Thus, the dynamic performance of the control-loops is neither analyzed nor considered in the design of the controllers [14]-[18].

It is well known that the dynamic performance of the model is essential for designing controllers because the steady-state model is incapable of providing an accurate insight into the transient behaviours of the system and evaluating the maximum stresses during startup and shutdown processes [20]. Thus, in many industrial applications, the settling time and overshoot of the controllers should be optimized according to the performance requirements. Moreover, because the WCS with the charging and ZVS operation controllers are multivariable systems, the couplings between the model inputs should be clarified. Therefore, there is a need for dynamic modeling, which can be used to analyze the dynamic performance and couplings during the design process of the controllers.

In addition, it is necessary to take into account the fact that the operating point of the system changes as the charging mode switching while designing controllers for both CC and CV modes. To deal with this problem, two closed control-loops with PI controllers for CC/CV modes based on the first harmonic equivalent models are proposed in [21] [22] to regulate the charging process and to maintain the ZVS operation. The output gain and phase of each charging mode are analyzed in the frequency domain. Unfortunately, although the control performance can be improved, these models are static and, as such, they are inconvenient to analyze the dynamics and couplings of the system during the design of controllers. Therefore, different modeling methods are proposed in [20] [23]-[32] to find an appropriate dynamic model. For example, a model based on the Laplace phasor transform theory [23] [24] is adopted to analyze the dynamics of the circuits. The LCL circuit [20] [25], LCC circuit [26], converters [27] and S-S compensation circuits [28]-[31] of the IPT and WPT system are modeled using generalized state space averaging (GSSA) method. Typically, based on [20] [29], authors in [31] propose two transfer function models for control-to-output corresponding to the CC and CV modes. Compared to [21] [22], these models provide good insights on the envelope of charging current and voltage and give a numerical analysis on the dynamic performance such as the magnitude-phase characteristics of the CC and CV control-loops. However, it is difficult to describe the state of the ZVS operation by the proposed state variables, resulting in an additional current stress optimization controller on the primary side to maintain the ZVS operating. In [32], a Coupled-Mode theory [34] based model, covering the ZVS angle, charging current and voltage, is proposed for analyzing and controlling the output voltage. Since the state variables of the dynamic differential equations of the circuit can be replaced by the phase angle and amplitude of the coupled modes, it is more suitable to design a controller to regulate the charging and ZVS operation of this model. Also, based on this modeling method, the couplings between the inputs of the multivariable WCS models can also be decoupled by evaluating the static gain of each control loop.

The main purpose of this paper is to improve the performance of the WCS throughout the whole charging period by proposing a decoupled control method. To overcome the perturbance of the system operating point, two dynamic models for CC and CV modes are constructed and linearized, to analyze the dynamic performance of the system. Moreover, the couplings between the inputs of the charging and ZVS operating controllers are quantified using static gains. By introducing decoupled compensators and PI controllers, the decoupled control method for IPT system is proposed and implemented to regulate the charging process and ZVS operation. Compared to the coupled control methods and the traditional closed-loop control feedback methods, the recorded response curves and transient waveforms demonstrate that the proposed decoupled control method brings better performance. Differentiating from the existing works, the contributions of this paper are as follows:

(1) Considering that the dynamic models are different in CC and CV modes, a fourth-order and a fifth-order models are developed, respectively, to prevent the weakening of the performance of the system.

(2) The static gains are used to evaluate the couplings between the inputs of the charging and ZVS operating controllers. By designing decoupled compensators, the multivariable CC and CV models are equivalent to two single variable models and, in turn, they become easier to design controllers to regulate the dynamical behaviours of the decoupled systems.

(3) To optimize the system performance, we chose PI controllers, and their parameters are obtained by calculating the settling time and overshoot of the close-loop dominant poles according to the requirements of the control performance, which is especially useful when the control system is required to be designed tightly.

The rest of this paper is organized as follows: Section II analyses the input impedance, ZVS angle and outputs of the WCS and proposes a decoupled control strategy. Section III focuses on the dynamic modeling and linearization of the CC and CV modes. To optimize control performance, a
decoupled control method including compensators and PI controllers are proposed in section IV. A physical prototype is set up in Section V, and the recorded response curves and transient waveforms demonstrate that the proposed decoupled control method brings better performance to the system.

II. ANALYSIS OF THE SYSTEM STRUCTURE

A. STRUCTURE OF THE BASIC S-S WCS

A common series–series (S-S) compensated topology for WCS, as shown in Fig. 1, is composed of a controllable DC source, a high-frequency half-bridge inverter, a symmetrical S-S resonant network, a full-bridge rectifier and a battery equivalent series resistance (ESR). The DC source converts a three-phase AC voltage into a controllable DC voltage source \( V_1 \) by regulating the driver of the converter. Then, the resonant network and the full-bridge rectifier are activated by the high-frequency half-bridge inverter to charge the battery wirelessly. Each resonant tank includes a series capacitance, an inductance and its ESR, respectively. The symbol \( i_p \) and \( i_s \) are the resonance current through the coils; \( v_b \) and \( v_s \) are the voltage across the capacitors. \( M \) is the mutual inductance between two coils.

To realize the charging control and the ZVS operation control, the controllable \( V_1 \) and the operation frequency \( f_o \) are adjusted by the proposed decoupled control method, as shown in Fig.1. In detail, by detecting the zero-crossing of \( i_p \) and the rising edge of the gate-driving signal of the inverter, the phase difference between \( i_s \) and \( v_{ab} \) can be obtained to predict the state of the ZVS operation. In order to implement the CC and CV charging mode, the output current and voltage are detected to form a current and a voltage feedback control-loops, respectively, by using a common wireless communication module. The parameters of the PI controllers in CC and CV modes are optimized based on the respective dynamic models and performances to overcome the perturbation caused by the change of the operating point. In addition, the control performance of the controllers is improved because the couplings between the ZVS control-loop and the charging control-loop is cancelled by the decoupled compensators.

B. ANALYSIS OF THE CIRCUIT

Assuming that:

1. The resonant currents, \( i_1 \) and \( i_2 \), are pure sinusoidal.
2. The filter capacitance of the full-bridge rectifier is big enough.

The fundamental components of the inverter output voltages \( v_{ab} \) and the ESR \( R_{cd} \) of the full-bridge rectifier can be replaced by

\[
\begin{align*}
\text{v}_{ab} &= \left(2V_i \cos \omega t\right)/\pi, \\
R_{cd} &= 8R_b / \pi^2, 
\end{align*}
\]

where \( \omega \) and \( R_b \) are the angular switching frequency of the bridge inverter and the ESR of the battery, respectively. Therefore, the main circuit of the WCS can be equivalent to a fundamental harmonic equivalent circuit, as depicted in Fig. 1. The charging voltage and current of the battery are labelled as \( V_o \) and \( I_o \) respectively. The input impedance \( Z_{in} \) of the equivalent circuit can be expressed per:

\[
Z_{in} = Z_p + (\omega_i M) \left(1 - D_{inv}\right),
\]

where \( Z_p \) and \( Z_s \) stand for the primary and secondary resonant network impedance, respectively. According to [7], the ZVS angle \( \phi_{ZVS} \) can be depicted as

\[
\phi_{ZVS} = \Delta \phi - \phi_d = \text{arctan} \left(\frac{\text{Im}[Z_{in}]}{\text{Re}[Z_{in}]}\right) - \frac{(1 - D_{inv})\pi}{2},
\]

where \( \Delta \phi \) is the phase angle of the input impedance \( Z_{in} \), \( \phi_d \) is the phase angle of the dead-time of the inverter, and \( D_{inv} \) denotes the duty cycle of the square wave drive signal of the inverter.

According to the equivalent circuit, one obtains that the equivalent input voltage source \( v_{cd} \) of the full-bridge rectifier [35] is as follow:
\[ V_{cd} = j\omega M \frac{V_{ab}}{Z_{in}} + i_s Z_s . \]

(5)

It can be observed that the magnitude of \( V_{cd} \) is equal to the charging voltage \( V_o \), and the direction is decided by the sign of \( i_s \). In addition, the equivalent DC current source \( I_0 \) of the \( R_{battery} - C_f \) circuit input is a function of the rectifier input AC current \( i \), [35], which can be represented by the following expression:

\[ I_0 = I_p - i_{cf} = |i_s| - i_{cf} . \]

(6)

Considering (1)-(6), one can draw a conclusion that the charging voltage \( V_o \) and the current \( I_0 \) can be controlled by adjusting \( V_1 \). Meanwhile, the ZVS angle \( \phi_{zvs} \) can remain a constant by regulating the angular switching frequency \( \omega_0 \). Notably, due to the couplings of the multivariable systems, the charging voltage and current are fluctuated with \( \omega_0 \) at this time. Therefore, the decoupled control strategy for the battery charging and system ZVS operation is proposed in Fig. 2. The parameters of the PI controllers in CC and CV modes are optimized based on the respective dynamic models and performances to overcome the perturbation caused by the change of the operating point. Besides, the compensators are designed based on the static gains and are utilized to neutralize the coupleings of the system.

III. DYNAMIC MODELING OF THE SYSTEM

In this section, the dynamic models of the CC and CV charging modes covering the states of the ZVS angle, charging current and charging voltage are developed and linearized.

\[ \begin{align*}
\dot{i}_{p,s} &= \sqrt{2I_{p,s} \cdot a_{p,s} \cos(\omega t + \theta_{p,s})} \\
\dot{v}_{p,s} &= \frac{1}{2C_{p,s}} \cdot a_{p,s} \sin(\omega t + \theta_{p,s}) .
\end{align*} \]

(8)

Thus, one obtains a nonlinear time-invariant averaged model of the WCS given as following:

\[ \begin{align*}
\dot{x} &= f(x) + g(x)u \\
y &= h(x),
\end{align*} \]

(9)

where \( x=[a_p, \theta_p, a_s, \theta_s]^T \) and \( u=[I_o, V_1]^T \) are the state vector and inputs of the model, respectively; \( h_{wcs}(x) \) is the output of the ZVS angle \( \phi_{wcs} \); \( h_c(x) \) and \( h_v(x) \) are the outputs of the charging current and voltage, respectively, which will be detailed subsequently in the next subsections; \( f(x) \) and \( g(x) \) are smooth nonlinear functions given by

\[ f(x) = \begin{pmatrix}
K_1 \left(-K_1 S_{x_1} + K_2 M S_{x_2} x_3 - MK_3 L_{x_1}\right) \\
\frac{\omega_p}{2} \frac{K_1}{a_p} \left(K_1 S_{x_1} + K_2 M S_{x_2} x_3 - K_3 L_{x_1}\right) \\
\frac{\omega_p}{2} \frac{K_2}{a_s} \left(K_2 S_{x_2} + K_3 M S_{x_3} x_4 - K_4 L_{x_2}\right) \\
\frac{\omega_p}{2} \frac{K_3}{a_t} \left(K_3 S_{x_3} + K_4 M S_{x_4} x_5 - K_5 L_{x_3}\right)
\end{pmatrix} , \]

(10)

\[ g_{v_1}(x) = \begin{pmatrix}
K_1 L_{x_1} \cos x_2 \\
-x_1 \\
-K_3 M \cos x_4 \\
K_4 M \sin x_4
\end{pmatrix} , \quad g_{v_2}(x) = \begin{pmatrix}
0 \\
-1 \\
0 \\
1
\end{pmatrix} . \]

(11)
respectively, where \( S_1, S_2 \) and the variables \( K_1-K_8 \) are listed in the Appendix. For the control of the ZVS operating, the angle \( \varphi_{ZVS} \) should be a constant no matter the system is in the CC or CV charging mode, so the output of ZVS angle can be defined as

\[
h_{ZVS}(x) = -\frac{180}{\pi} \left( x_2 + \frac{(1-D_{on})}{2} \right),
\]

(12)

Since the operating point of the system will change as the charging mode switches, the dynamic models of the CC and CV charging modes need to be derived, respectively.

**B. CC AND CV NONLINEAR DYNAMIC MODEL**

Taking the average values during a switching period, the equation (6) yields the DC component of the charging current [35]:

\[
I_o = \frac{1}{2\pi} I_{sm} \cos \omega_t, \quad (13)
\]

which is proportional to the amplitude \( I_{sm} \) of the secondary current. The output of the charging current in the CC mode with the state vector \( x_{CC} = [\alpha_{ccp}, \theta_{ccp}, \alpha_{ccs}, \theta_{ccs}]^T \) can be defined as

\[
h_{cc}(x_{cc}) = \frac{2}{\pi} \sqrt{L_s} I_{sm} \cos \omega_t.
\]

(14)

Thus, the nonlinear dynamic model of the CC mode can be derived by combining (9)-(12) and (14). Moreover, the energy storage element \( C_i \) in Fig. 1 can be described by an additional differential equation

\[
\frac{dv}{dt} = I_r - \frac{v}{R_b}.
\]

(15)

By substituting (14) into (7), a fifth-order nonlinear time-invariant averaged model of the CV mode is obtained as follow:

\[
\begin{align*}
\dot{x} &= f(x) + g(x)u \\
\frac{dx}{dt} &= \frac{1}{C_i} \left( \frac{2}{\pi} \sqrt{L_s} x_3 - x_1 \right),
\end{align*}
\]

(16)

where the state vector of the CV mode is \( x_{CV} = [\alpha_{cvp}, \theta_{cvp}, \alpha_{csv}, \theta_{csv}, V_L]^T \), and the output of the charging voltage in the CV mode with the above state vector can be defined as

\[
h_{cv}(x_{cv}) = V_o.
\]

(17)

**C. MODEL LINEARIZATION**

The nonlinear time-invariant averaged models of the CC and CV modes represent the envelope of the charging current, voltage and ZVS angle of the WCS. In other words, these models contain both the steady-state operating points and the small-signal models. Thus, each state variable in (9)-(12) and (14)-(16) can be written as the sum of the steady-state value and a small perturbation around it [29][36]

\[
x = x_0 + \Delta x.
\]

(18)

Therefore, the nonlinear models can be linearized at the steady-state operation point and written in the state-space form as follows:

\[
\begin{align*}
\Delta \dot{x} &= A \cdot \Delta x + B \cdot \Delta u \\
\Delta y &= C \cdot \Delta x,
\end{align*}
\]

(19)

where A, B and C are the partial derivative matrices of (10)-(12) and (14)-(17). To obtain the linearized small-signal models, the steady-state operating point should be calculated by setting the derivative terms in (9) and (16) to zero.

**TABLE 1. Main parameters of the system**

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Quantity</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( L_p )</td>
<td>Resonant inductance on primary side</td>
<td>33.1 ( \mu )H</td>
</tr>
<tr>
<td>( C_p )</td>
<td>Resonant capacitance on primary side</td>
<td>122.1 pF</td>
</tr>
<tr>
<td>( R_p )</td>
<td>The ESR at primary side</td>
<td>0.5 ( \Omega )</td>
</tr>
<tr>
<td>( L_s )</td>
<td>Resonant inductance on secondary side</td>
<td>34.3 ( \mu )H</td>
</tr>
<tr>
<td>( C_s )</td>
<td>Resonant Capacitance on secondary side</td>
<td>121.9 pF</td>
</tr>
<tr>
<td>( R_s )</td>
<td>The ESR at secondary side</td>
<td>0.5 ( \Omega )</td>
</tr>
<tr>
<td>( M )</td>
<td>Mutual inductance between two coils</td>
<td>7.33 ( \mu )H</td>
</tr>
<tr>
<td>( C_f )</td>
<td>Filtering capacitance</td>
<td>300 ( \mu )F</td>
</tr>
<tr>
<td>( D_{dv} )</td>
<td>The duty cycle of drive signal for inverter</td>
<td>45 %</td>
</tr>
<tr>
<td>( f_s )</td>
<td>The range of the operating frequency</td>
<td>75-90 kHz</td>
</tr>
<tr>
<td>( V_i )</td>
<td>The range of the input DC voltage source</td>
<td>50-750 V</td>
</tr>
</tbody>
</table>

**IV. SIMULATION ANALYSIS AND DESIGN OF THE DECOUPLED CONTROLLERS**

In this paper, the reference constant charging current and voltage are set as 10 A and 50 V, respectively. Considering the analysis of the total loss [7] and the minimum phase angle [35] of the system, the reference ZVS angle is selected to be 20 °. Since the battery resistance changes from 2.5 \( \Omega \) to 5 \( \Omega \) in the CC mode, and from 5 \( \Omega \) to 15 \( \Omega \) in the CC mode [5], the range of the load resistance in the CC and CV dynamic models are selected as 3.5-10 \( \Omega \). In order to achieve the expected charging curve and maintain the ZVS angle at the reference angle of 20 °, the initial DC input voltage and operating frequency of the WCS in the CC and CV modes are set to be: 55 V and 88.5 kHz, 66 V and 82 kHz, respectively. According to the main parameters of the system in Table 1, one can obtain the operating points for the CC and CV charging modes as

\[
\begin{align*}
x_{CC0} &= [0.05, -0.35, 0.04, -2.9]^T \\
x_{CV0} &= [0.07, -0.35, 0.03, -2.13, 50]^T.
\end{align*}
\]

(20)

By substituting (20) and Table 1 into the CC and CV nonlinear models, the partial derivative matrices for the linearized small-signal state-space models are obtained:

\[
A_{CC} = \begin{bmatrix}
-7856 & 2332 & -4.1e4 & -1568 \\
-9e5 & -8025 & 7.5e5 & -3.3e4 \\
3.2e4 & -2346 & -5.1e4 & 2210 \\
1.1e6 & 3.9e4 & -1.3e6 & -5e4
\end{bmatrix}.
\]

(21)
The block diagram of the (a) decoupled compensator and (b) decoupled system.

FIGURE 3. Unit step responses of the linearized MIMO CC/CV models with/without the decoupled compensators.

FIGURE 4. The block diagram of the (a) decoupled compensator and (b) decoupled system.

plotted in Fig. 3. It can be seen that the input \( u_2 \) has little static deviation on the output \( y_1 \), while the output \( y_2 \) fluctuates with the step change of \( u_1 \). Notably, although the static deviation from DC input voltage to ZVS angle is small enough, the dynamic fluctuation in CV mode cannot be ignored. To eliminate these deviations and fluctuations, the feedforward compensators are adopted in this paper. The block diagram of the decoupled system is shown in Fig. 4(a). Based on the diagonal matrix method, the compensation matrices \( W_d(s) \) can be calculated per

\[
W_d(s) = G(s)^{-1} N_L(s),
\]

where \( G(s) \) is the transfer function models of the CC and CV modes, and \( N_L(s) \) is the diagonal elements of \( G(s) \). Due to the high order of \( W_d(s) \), the balanced truncation method [36] is used to reduce the order of the compensators. Hence, the compensation transfer functions from \( u_1 \) to \( y_1 \), and from \( u_2 \) to \( y_1 \) for the CC and CV linearized models can be obtained as

\[
W_{dc12}(s) = \frac{-1163s + 1.2e10}{s^2 + 2.8e4s + 1.3e10},
\]

\[
W_{dc21}(s) = \frac{-1163s + 1.2e10}{s^2 + 2.8e4s + 1.3e10},
\]

\[
W_{dc12}(s) = \frac{-7.1s^3 + 6.7e9 + 1.3e14}{s^4 + 4.6e5s^3 + 3e9s + 2.4e13},
\]

\[
W_{dc21}(s) = \frac{-7.1s^3 + 6.7e9 + 1.3e14}{s^4 + 4.6e5s^3 + 3e9s + 2.4e13}.
\]
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FIGURE 5. Unity feedback root locus of (a) \( N_{CC11}(s) \), (b) \( N_{CC22}(s) \), (c) \( N_{CV11}(s) \) and (d) \( N_{CV22}(s) \).

The unit step responses of the CC and CV linearized models with the decoupled compensations are plotted in Fig. 3. Compared to the curves without the compensators, one can see that the fluctuations and the static deviation are suppressed and compensated to zero. Therefore, the multivariable models are simplified to two single variable models with the diagonal elements of \( G(s) \) as shown in Fig. 4(b).

B. DESIGN OF THE EV CHARGING AND ZVS OPERATING LOOP CONTROLLERS

The diagonal transfer function models \( N_i(s) \) of the CC and CV linearized models are

\[
N_{CC11}(s) = \frac{3.6e4s^4 + 2.7e9s^3 + 3.7e14s + 1e19}{s^4 + 1.2e5s^3 + 1.5e10s + 8e14s + 1.3e19}
\]

\[
N_{CC22}(s) = \frac{1315s^4 + 1.5e8s^3 + 4.6e13s + 1.5e18}{s^4 + 1.2e5s^3 + 1.5e10s^2 + 8.1e14s + 1.3e19}
\]

\[
N_{CV11}(s) = \frac{3.6e4s^4 + 5e9s^3 + 4e14s^2 + 3e19s + 4e23}{s^4 + 2e5s^3 + 1.6e10s^2 + 1e15s^2 + 4e19s + 5.5e23}
\]

\[
N_{CV22}(s) = \frac{1.8e8s^3 + 9.5e13s^2 + 1e19s + 4.2e23}{s^4 + 2e5s^3 + 1.6e10s^2 + 1e15s^2 + 4e19s + 5.5e23}
\]

(30)

where \( N_{CC11}(s) \) and \( N_{CC22}(s) \) are the transfer functions of the CC charging mode from \( u_1 \) to \( y_1 \), and from \( u_2 \) to \( y_2 \), respectively; \( N_{CV11}(s) \) and \( N_{CV22}(s) \) are the transfer functions of the CV charging mode from \( u_1 \) to \( y_1 \), and from \( u_2 \) to \( y_2 \), respectively. The unity feedback root locus of (30) is plotted in Fig. 5. By analyzing the distribution of the closed-loop zero-pole, the following conclusion can be obtained:

1. The stability of (30) is available because the closed-loop poles are all located in the left s-plane. However, the plural poles of \( N_{CV22}(s) \), as depicted in Fig. 5 (d), indicate that the stability of the charging voltage in the CV mode is conditional.

2. Since the dynamic performance of the model mainly depends on the real part and attenuation rate of the dominant pole, a dipole can be ignored if the distance between its zero and pole is much smaller than its distance from the origin.

3. The proportion and integration elements of the PI controllers in each control-loop are equivalent to add a new closed-loop zero and a new dominant pole to the models. Therefore, the parameters of the controllers can be calculated by arranging the distribution of the closed-loop dominant pole with reference to the performance requirements of the settling time and overshoot.

The PI controller has the following form:

\[
G_i(s) = K_p \left(1 + \frac{1}{T_i s}\right),
\]

(31)

where \( K_p \) and \( T_i \) are the proportional coefficient and the integral time, respectively. According to the decoupled system, as shown in Fig. 4 (b), the closed-loop transfer function with the PI controller of each control-loop can be expressed as

\[
\Phi(s) = \frac{G_i(s)N_i(s)}{G_i(s)N_i(s)+1} \approx \frac{\omega_n^2}{z} \frac{(s+z)}{s^2 + 2\xi\omega_n s + \omega_n^2},
\]

(32)

where \( \omega_n \) and \( \xi \) are the natural frequency and damping ratio of the closed-loop dominant pole brought by the PI controllers. Thus, the closed-loop characteristic equation of (32) is

\[
s^2 + 2\xi\omega_n s + \omega_n^2 \approx K_p (s + K_i) \prod_{i=1}^{m} (s - z_i) + 1 = 0,
\]

(33)

Assuming that the performance requirements of the settling time and overshoot are \( t_s = 200 \) ms and \( \sigma\% = 5\% \) with an error band of 5\%, the distribution of the dominant pole can be calculated by

\[
t_s = 3.5 / \xi \omega_n, \quad \sigma\% = e^{-2\xi\sqrt{\omega_n^2 - \xi^2}} \times 100\%.
\]

(34)

Using MATLAB, the final parameters of the CC and CV PI controllers are calculated as: \( K_{pCC} = 0.8, \quad T_{iCC} = 0.003; \quad K_{pCV} = 0.003, \quad T_{iCV} = 0.03 \); \( K_{pCCV} = 0.1, \quad T_{iCCV} = 0.03; \quad K_{pCV} = 0.1, \quad T_{iCV} = 0.003; \quad K_{pCVZVS} = 0.15, \quad T_{iCVZVS} = 0.026. \)

V. EXPERIMENTAL RESULTS

A. THE WCS PROTOTYPE

Fig. 6 shows the physical experimental setup in terms of the topology circuit in Fig. 1. The experimental setup consists of a controllable DC source, a PCB including a half-bridge inverter, MCUs and a wireless communication module, an S-


FIGURE 6. Physical experimental setup of the S-S compensated wireless charging system.

S resonant network, a full-bridge rectifier and four switchable power resistors. The parameters of the components are shown in Table 1. Two IPW65R041CFD MOSFETs are used for the half-bridge inverter, and the controllers are implemented with two MCUs, namely, an FPGA (XC6SLX9-3TQG144I) and an ARM (STM32F407VG144). The type of the controllable DC source is INFY POWER REG75030, and it is given in the data sheets that the control interval and the settling time of the output voltage are about 40 ms and 6 ms, respectively. The range of the output voltage is listed in Table 1. The ARM is used to calculate the operating frequency and input DC voltage of the inverter, while the FPGA and DC source regulate the inverter driving signals and input DC voltage with a control period of 40 ms and 20 ms, respectively, according to the ZVS angle and charging commands from the ARM. The wireless communication for the charging current and voltage data are set up with two nRF24L01 modules. The dimensions of the two coaxial coils are the same, whose rectangular outer ring is 90 cm x 70 cm. The distance between the coils is 20 cm. Each coil is built with four turns of Litz-wire. The diameter of the Litz-wire consisting of 2000 isolated strands is 6 mm, and the diameter of each strand is 0.1 mm. At the secondary side, a class-D full-bridge rectifier is constructed using four VS-UFB280FA40 diodes. For the load, different resistance values are obtained by connecting four 10 Ω resistors in series or in parallel with an air switch.

B. EXPERIMENT OF CC AND CV CHARGING

To validate the improvement of the control performance of the system with the proposed decoupled control method, three cases of experiments are set up in this paper, namely, (i) decoupled control, (ii) coupled control, and (iii) traditional closed-loop feedback control. According to the range of the ESR of the lead battery during the charging process [5]-[7], two step changes of the load resistance for the CC and CV charging modes, namely, from 2.5 Ω to 5 Ω and from 7.5 Ω to 10 Ω, respectively, are considered as the disturbances in the experiment.

By step change of the resistor bank, the dynamic responses of the charging current and voltage in case (i) are captured by an oscilloscope (Tektronix DPO 2004B) as shown in Fig. 7(a) and (b). It can be seen that both the charging current and the voltage can reach the constant references, namely, 10 A and 50 V, within 200 ms which are in good agreement with the preset performance requirements. In addition, the dynamic waveforms for case (ii) and case (iii) in each charging mode are depicted in Fig. 7(c), (d), (e) and (f), respectively. In case (ii), when the load changes, the settling times of the CC and CV modes are all 235 ms. Meanwhile, the charging current and voltage in case (iii) can reach the reference within 435 ms and 420 ms, respectively. What is worth mentioning, comparing with the case (i), the settling times of the case (ii) and case (iii) are longer than the requirements. In addition, the same tendency can also be observed in the overshoots. Therefore, the following conclusions can be obtained:

1. The charging current and voltage in both CC and CV modes can reach the reference with the above control methods, namely, cases (i)-(iii).

2. The proposed decoupled control method (case (i)) can obtain a better control performance than other control methods. This is because that the couplings between the charging loop and the ZVS operation loop are accurately calculated in the solving process of the compensation transfer functions, therefore, the compensators can eliminate the fluctuations and improve the control performance. In contrast, the controllers for case (ii) and (iii) need more time to deal with the couplings of the system and the change of the operation point.

3. In Fig. 7 (c) and (d), although the charging current and voltage need more time to return to the reference compared to case (i), they respond faster than that in case (iii). This proves that the PI controllers which designed on the basis of the corresponding models are more robust to the change of the operating point.

For ZVS operation control, the phase difference between the output voltage and current of the inverter is measured by the FPGA and current sensor. In this paper, the data of the
FIGURE 7. The Dynamic waveforms of the charging current and voltage when the load resistance changes. (a) and (b) are the charging current and voltage in the case (i); (b) and (c) are the charging current and voltage in the case (ii); (e) and (f) are the charging current and voltage in the case (iii).

FIGURE 8. Dynamic waveforms of the ZVS angle in the (a) CC and (b) CV charging modes when the load resistance changes.

ZVS angle is calculated and saved by the MCUs and RS232 communication circuit at a period of 20 ms. Therefore, the waveforms of the ZVS angles in the CC and CV charging modes with different control methods (case (i)-(iii)) are plotted in Fig. 8 (a) and (b). Although the ZVS angles rise rapidly when the load changes, they gradually return to the reference of 20°, and no overshoots are observed in each case. Similar to the charging control-loop, the ZVS angles of the CC and CV modes in case (i) can obtain a better control performance, and reach the steady-state within 200 ms.

Since the system is semi-coupled, compared with the waveforms in Fig. 7, the proposed decoupled compensator has less improvement to the performance of the ZVS operation. In addition, since the dynamic fluctuation in the CV mode is larger than that in the CC mode, as depicted in Fig. 3, the decoupled compensator plays a more significant role in the CV charging mode. Furthermore, it can be seen that the settling times of the case (ii) are shorter than the case (iii), which also proves that the PI controllers based on different dynamic models can prevent the degradation of the performance caused by the change of the operating point.

The steady-state waveforms of the inverter output voltage and current in the CC and CV modes are captured by an oscilloscope and shown in Fig. 9 (a) through (d). One can see...
from Fig. 9 (a) that the ZVS angle and operating frequency in the CC mode are about 20° and 84.91 kHz, respectively. When the load resistance step rises to 5 Ω, as shown in Fig. 9 (b), the steady-state ZVS angle can be maintained at the reference of 20°, and the operating frequency is adjusted to 89.43 kHz. The same tendency can be observed in CV mode in Fig. 9 (c) and (d). When a disturbance is given to change the load resistance from 7.5 Ω to 10 Ω, the operating frequency is regulated from 80.71 kHz to 80.32 kHz to maintain the ZVS angle at constant. The above two sets of waveforms all illustrate that the ZVS operation can be achieved by changing the operating frequency of the inverter.

VI. CONCLUSION

In this paper, we propose a decoupled control method to improve the performance of the charging and ZVS operation controllers in the WCS. Since the operating point of the CC and CV modes are different, two dynamic models for the corresponding modes are built and linearized. By analyzing the couplings and distributions of the zero-pole, the decoupled compensators are designed to neutralize the deviations and fluctuations. Meanwhile, to optimize the system performance, we chose PI controllers, and their parameters are obtained by calculating the settling time and overshoot of the close-loop dominant poles according to the requirements of the control performance, which is especially useful when the control system is required to be designed tightly. To verify the improvement of the performance with the decoupled control method, a 500W S-S inductive power transfer system prototype is built and tested. The recorded response curves and steady-state waveforms demonstrated that the charging current, voltage and ZVS angle can reach the references in the whole charging period. Furthermore, the proposed decoupled controllers can achieve a better control performance compared with the coupled control method and the traditional closed-loop feedback control method.

APPENDIX

\[ K_1 = \frac{\sqrt{L_p/2}}{L_pL_s-M^2} \]
\[ K_2 = \frac{\sqrt{L_p/2}}{L_pL_s-M^2} \]
\[ K_3 = \frac{M}{\sqrt{2C_s}} \]
\[ K_4 = \frac{M}{\sqrt{2C_p}} \]
\[ K_5 = \frac{M}{\sqrt{2L_p}} \]
\[ K_6 = \frac{MR}{\sqrt{2L_p}} \]
\[ K_7 = \frac{L_i}{\sqrt{2C_p}} \]
\[ K_8 = \frac{L_p}{\sqrt{2C_s}} \]
\[ S_1 = \sin(x_1-x_3) \]
\[ S_2 = \cos(x_2-x_4) \]

REFERENCES


