Frequency tracking and tuning control of wireless power transfer system

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Abstract. A tuning control method with frequency tracking function is proposed to improve the degradation of efficiency caused by coil detuning in magnetically-coupled resonant wireless power transfer. Firstly, the detuning mechanism is studied by combining the AC impedance characteristics of the series resonant circuit, and the feedback control circuit is constructed by using a modified phase-locked loop, which outputs a variable frequency PWM wave to regulate the operating frequency of the high frequency inverter and maintains the phase difference between the inverter output voltage and the original side current within the error range. The Matlab/Simulink simulation results show that the design can successfully transfer the system to a new resonant state with short regulation period and high control accuracy, which can effectively improve the transmission efficiency and load power of the system.

1. Introduction

In recent years magnetically-coupled resonance wireless power transfer (MCR-WPT) systems are gradually replacing the traditional power supply [1] in cell phones, robots, and electric vehicles by virtue of their advantages such as freedom from physical connection and flexibility of power supply. To ensure the stability and efficiency of the MCR-WPT system [2], its operating conditions need to be maintained in the resonant state [3]. However, changes in transmission distance and system parameters in the actual environment will cause the resonant frequency of the system to shift [4], that is, detuning will occur, thus affecting the transmission performance of the system. Therefore, the adaptation of the resonant frequency is a key technology for the MCR-WPT system [5].

The frequency tracking and tuning control proposed in this paper avoids the reduction of efficacy due to long-term operation of the system by controlling the on-off of the switch tubes of the high-frequency inverter link so that its output frequency always adapts to the changing system parameters with the same inherent resonant frequency in the transmitter. The system parameters are determined through simulation, and the modified phase-locked loop (PLL) maintains the inverted square wave voltage in the same frequency and phase as the transmitter current so that the input impedance of the system is purely resistive during resonance, which can provide a theoretical basis and solution for studying the MCR-WPT technology with frequency regulation.

2. Magnetically-coupled resonant wireless power transfer system model

2.1. MCR-WPT system structure

Figure 1 shows the MCR-WPT system using a series-series (S-S) topology.
Figure 1. S-S type mutual inductance model.

$U_s$ is the fundamental component of the high-frequency power supply, and $R_s, R_1, R_2$ are internal resistances. $L_1, L_2$ are the inductance of the coil at the transmitting end and the receiving end respectively; $M$ is the mutual inductance; $C_1, C_2$ are the compensation capacitors; $R_L$ is the load.

When the high frequency power supply operates at $\omega$, the primary equivalent impedance $Z_1$, the secondary equivalent impedance $Z_2$ can be expressed as:

$$Z_1 = R_s + j\omega L_1 + \frac{1}{j\omega C_1} + R_1$$
$$Z_2 = R_2 + j\omega L_2 + \frac{1}{j\omega C_2} + R_L$$

where $M = k\sqrt{L_1L_2}$ (k is the coupling coefficient).

The Kirchhoff’s voltage law (KVL) equation is written for the mutual inductance model:

$$\begin{align*}
Z_1 I_1 - j\omega M I_2 - U_s &= 0 \\
I_2 Z_2 - j\omega M I_1 &= 0
\end{align*}$$

This leads to the system output power and efficiency through equation:

$$\eta = \frac{P_{\text{out}}}{P_{\text{in}}} = \frac{I_2^2 R_L}{U_s^2 \Re(I_1)} = \frac{\omega^2 M^2 R_L}{Z_2^2 Z_1 Z_2 + (\omega M)^2}$$

The optimal efficiency point corresponding to the frequency $\omega_0$ can be found by making $\frac{d\eta}{d\omega} = 0$:

$$\omega_0 = \frac{1}{\sqrt{L_1C_1}} = \frac{1}{\sqrt{L_2C_2}}$$

At this point $\eta$ takes the maximum value $\eta_{\text{max}}$:

$$\eta_{\text{max}} = \frac{(\omega_0 M)^2 R_L}{(R_L + R_1)^2 (R_L + R_1) + (R_L + R_2)(\omega_0 M)^2}$$

Looking in from the power side of Figure 1, introduce the reflected impedance $Z_r = \frac{(\omega M)^2}{Z_2}$, the MCR-WPT system can be equated to a passive one-port network with an input impedance of:

$$Z_{in} = Z_1 + Z_r = \Re(Z_{in}) + j\Im(Z_{in})$$

It is known that $\Im(Z_{in}) = 0$, so impedance angle $\phi = \arctan\frac{\Im(Z_{in})}{\Re(Z_{in})} = 0$ in (6), which means that $Z_{in}$ presents a purely resistive form[6], the system resonates at frequency $\omega_0$, and the most significant feature is that the voltage and current at the transmitter are in the same frequency and phase[7].

2.2. Influence of detuning on transmission characteristics

In actual operation, due to the error of the coil inductance value and other factors, the resonant frequency will gradually shift, if $\omega_0$ is maintained as the original value, the complete resonant state is destroyed, that is, detuning occurs, the formula (6) is no longer a pure real number. The efficiency and power curves under detuning condition are shown in Fig.2. When the primary side is detuning, $\eta_{\text{max}}$ occurs at a frequency significantly shifted away from the fully resonant frequency of 300KHz; relatively, the detuning phenomenon has a less obvious effect on the power.
As can be seen, the change of inductance at the transmitter \( \Delta L_1 \) is the main factor causing the frequency shift. Therefore, \( \omega_0 \) needs to satisfy the following condition when the system resonates:

\[
\omega_0 = \omega_1 = \frac{1}{\sqrt{L_1 C_1}}
\]  

(7)

\( \omega_1 \) is the intrinsic resonant frequency of \( L_1 \). The curve of \( \phi \) shifted with the operating frequency is shown in Fig. 3. When \( L_1 \) occurs incremental or decremental detuning, \( \phi(\omega_0) \neq 0, \omega_1 \) changes to \( \omega_1' \) and no longer satisfies the formula (7).

### 3. Frequency track control method

In order to solve the detuning problem, this paper designs a frequency tracking control method, when \( \omega_1 \) is shifted to \( \omega_1' \), the operating frequency of HF inverter is changed online to \( \omega_0' \), so that the adjusted system conforms to (8) and ensures a new resonant working state:

\[
\omega_0' = \omega_1'
\]  

(8)

In this paper, we propose a frequency tracking and tuning control method based on an improved PLL. The control block diagram is shown in Figure 4. The primary current sampling signal \( u_{in}(\theta_{in}) \) and the feedback control signal are processed by the phase and frequency detector (PFD) composed of second-order generalized integrator (SOGI) to obtain the impedance angle information \( u_q \), \( u_q \) is calculated by the loop controller (ARC) to obtain the control input \( u_c \), according to which the voltage controlled oscillator (VCO) generates the feedback control signal \( u_{out}(\theta_{out}) \) whose frequency varies linearly with \( u_c \) and the pulse frequency modulation (PFM) signal generator generates the frequency adjustable PWM waves \( V_1-V_4 \) under the action of \( u_{out} \), which completes the frequency tracking control of the high frequency inverter through the driving circuit. The control system is shown in Figure 5. The implementation method of each main part is described below.

Figure 4. The proposed frequency tracking tuning control.

#### 3.1. phase and frequency detector with second-order generalized integrator

The model of SOGI-PFD is shown in Figure 5.
When the primary signal is $u_{in} = U_{in} \sin \theta$, set actual phase as $\hat{\theta}$, SOGI extracts two orthogonal sinusoidal signals from it as:

$$
\begin{align*}
\hat{u}_\alpha &= U_{in} \sin \hat{\theta} \\
\hat{u}_\beta &= U_{in} \cos \hat{\theta}
\end{align*}
$$

(9)

After the Park coordinate transformation, we get $u_q = U_{in} \sin (\hat{\theta} - \theta)$. When the system is stable, the difference between $\hat{\theta}$ and $\theta$ is very small, according to the equivalence infinitesimal principle $\lim_{\hat{\theta} \to \theta} U_{in} \sin (\hat{\theta} - \theta) = U_{in} \sin (\hat{\theta} - \theta)$. $u_q$ can be approximated as the phase difference value.

The SOGI structure introduced in this paper is not affected by the original edge, and high dynamic response can be achieved by designing reasonable parameters.

3.2. loop controller and voltage controlled oscillator

In this paper, a second-order active proportional integral filter is used for loop control. Set $\tau_1, \tau_2$ as the parameters to be designed. Let the transfer function of ARC be:

$$
F(s) = \frac{1 + s\tau_2}{s\tau_1}
$$

(10)

VCO is an integral section in the S domain and the transfer function is:

$$
N(s) = \frac{K_{VCO}}{s}
$$

(11)

$K_{VCO}$ is the gain, and to ensure the sensitivity of VCO, $K_{VCO} = 20000\text{Hz/v}$.

Introduce the intrinsic frequency and damping coefficient:

$$
\begin{align*}
\omega_n &= \sqrt{\frac{K}{\tau_1}} \\
\xi &= \frac{\tau_2}{2}\sqrt{\frac{K}{\tau_1}}
\end{align*}
$$

(12)

The PLL closed-loop transfer function is:

$$
H(s) = \frac{K_r F(s) N(s)}{1 + K_r F(s) N(s)} = \frac{2\xi \omega_n s + \omega_n^2}{s^2 + 2\xi \omega_n s + \omega_n^2}
$$

(13)

The loop gain $K = K_{VCO} K_r$. In order to obtain a better transient response, the damping factor $\xi = 0.707$ is taken; 0.5 times the fundamental frequency is chosen as the bandwidth to effectively filter out high frequency noise.

$$
B_l = \int_{0}^{\infty} |H(j2\pi F)| dF \approx \frac{\omega_n}{8\xi} \left(1 + 4\xi^2\right)
$$

(14)

Let $B_l = 300000\pi \text{ rad/s}$, so $\omega_n \approx 592384.4$ which substituted into (14) yields the parameters of the ARC of $\tau_1 = 2.39e-6, \tau_2 = 1.181e-8$.

Figure 6 shows the second-order loop bode diagram of the PLL. When $\xi = 0.707$, the phase margin $\gamma = 65^\circ$, the loop can work stably and maintain a strong immunity to interference. In addition to this, when there is a frequency difference in the input signal, the phase difference at steady state can be guaranteed to be 0 due to the presence of a near-ideal product link in its transfer function (10).
4. Simulation experiments and results

4.1. Space considerations

In this paper, the system is tested using Matlab/Simulink, and the circuit parameters are shown in Table 1. The system is detuned twice in the operation. The input impedance angle is shown in Figure 7. At 0.002s, $\phi$ first undergoes a short overshoot phase and then frequency tracking control can quickly achieve tuning; at 0.004s, the PLL can still precisely control the fluctuation range of $\phi$.

| Symbol | Value |
|--------|-------|
| $L_1$  ($L_2$) | 24.7e-6 |
| $C_1$  ($C_2$) | 11.4e-9 |
| $R_1$  ($R_2$) | 1.1 |
| $R_L$  | 10 |
| $U_{DC}$ | 40 |
| $M$    | 3e-6 |

The system voltage and current waveform are shown in Fig 8. In 8(a), at 0.002s, the resonant frequency drifts, and the figure shows that $U_1$ and $I_1$ are no longer in phase; then the PLL circuit starts frequency tracking and the phase difference gradually decreases; finally, the system shifts to the new resonant frequency $f = 273540$Hz.

In 8(b), when the system is adjusted, the $U_2$ waveform is more stable and the amplitude increases compared with that of before detuning, indicating that the frequency tracking effect of the PLL circuit can improve the quality of the load voltage waveform while increasing the transmission efficiency.
4.2. Simulation Data Analysis

Based on the simulation results, the output power $P_{out}$ and the system efficiency $\eta$ are plotted with the degree of detuning (time), as shown in Figure 9.

![Graph showing the variation of output power and efficiency with detuning](Image)

(a) $P_{out}$ varies with the degree of detuning  
(b) $\eta$ varies with the degree of detuning

Figure 9. Transmission characteristics.

Taking the open-loop system as the reference, the occurrence of detuning has an irreversible influence on the system - the frequency is shifted from the global optimal resonance point and leads to power and efficiency degradation, while the tuning method proposed in this paper cannot completely eliminate the adverse effect, but the local optimal resonance point can be found after $3e-4s$ of frequency tracking, thus improving the system characteristics and substantially recovering $P_{out}$ and $\eta$.

5. Conclusion

The MCR-WPT system needs to change the driving frequency to improve the transmission performance due to the detuning caused by the change of coil parameters in actual operation. To address this problem, this paper proposes a frequency tracking control tuning method through mathematical modeling of the detuning mechanism, which quickly reverts the phase difference between the inverter output voltage and the transmitting coil current within the error range after the system detuning, and actively adjusts the system to a new resonant state with high control accuracy. It can be verified by simulation that the transmission efficiency and load power of the MCR-WPT system can be effectively improved after adopting frequency track.

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