Current Balancer design in Parallel Operated Inverter for Several-kW class MHz-band Wireless Power Transfer System

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***Abstract*— This paper proposes a parameter design method for a current balancer used in paralleled MHz-band inverters for wireless power transfer (WPT) systems. The balancer, composed of two transformers, equalizes inverter currents by adjusting the mutual inductance and external inductance. Analytical relationships between the mutual inductance and current unbalance rate are derived, and a design flowchart is presented. Simulation results at 6.78 MHz confirm that the proposed design achieves a 5.0% current unbalance rate with output powers of 2.87-kW and 3.57 kW for two inverters, yielding 6.44 kW total. The results demonstrate that the proposed current balancer effectively enables kW-class MHz-band inverter operation with balanced current sharing.**

Keywords*— Wireless Power Transfer; Current balance; Circular current; transformer.

I. Introduction

Reduction of the CO_2 emission is an urgent issue for humanity. One of the effective solutions for the CO_2 reduction is electrification, which enables the utilization of the renewable energy. Thus, the electrification of drive systems, which is motor-based system instead of engine-based system, is rapidly spreading especially in mobilities. An electric mobility (e-mobility) requires large batteries to extend a driving range except for the electric train system. Although e-mobility emit little CO_2 during driving, a large amount of CO_2 is emitted in the manufacturing process of batteries.

Wireless Power Transfer (WPT) system is promising solutions to reduce the batteries. In particular, Dynamic-WPT(D-WPT) system, which enables battery charging during driving, is effective method to dramatically extend the driving range while reducing the battery capacity. Specifically, MHz-band WPT system is actively studied [1]-[4] because it reduces the weight and the volume of the transfer coil.

However, the system power of the MHz-band WPT system is limited by the rated current of power devices on the transmission side inverter. Although parallel connection of power devices increases the system power, variations of each device and the components should be considered. Thus, a difficulty of the power increasing by the paralleling power devices is still high especially in MHz-band. On the other hand, parallel operation of the inverter with current balancer is proposed [5] to increase the output power of the transfer side of the MHz-band WPT system. Although [5] utilizes two transformers to achieve the current balance in paralleled inverters, the detail parameter design of transformers is not mentioned.

This paper proposes a parameter design of the current balancer for the paralleled inverters in MHz-band WPT systems. The current balancer consists of two transformers to achieve the current balance. The relation of the mutual inductance and current unbalance rate is investigated analytically. The current balance effect is verified by the simulation. Moreover, an essential experiment with the prototype balancer is shown in this paper.

II. Current Balancer

A. Circuit Configulation

Figure 1 illustrates the configuration of paralleled inverters with the current balancer [6] on the transmission side of the MHz-band WPT systems. The current balancer is connected between each inverter and a resonant load. The resonant load

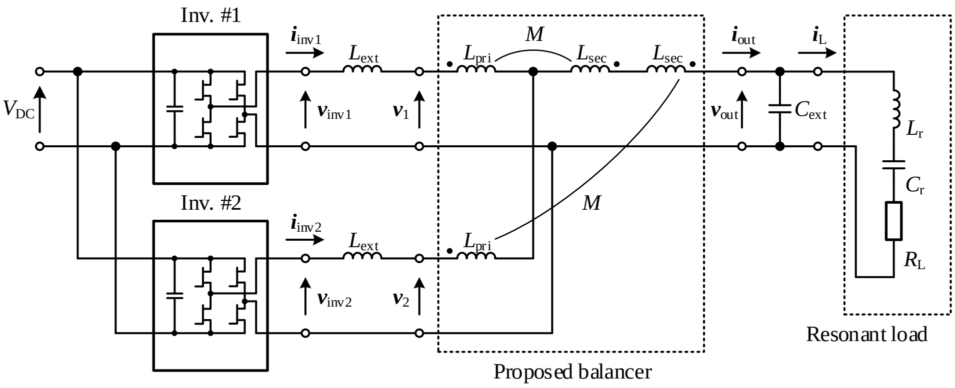


Fig. 1. Proposed balancer.

represents the characteristics of the secondary side of S-S type WPT systems. An external capacitor C_{ext} and two external inductors L_{ext} are connected to ensure an impedance matching between each inverter and the load. The external capacitor C_{ext} and external inductor L_{ext} are derived in [5] as

$$C_{ext} = \frac{1}{\omega R_L}\sqrt{\frac{2R_L}{r_{inv}} - 1} \quad (1),$$

$$L_{ext} = \frac{2C_{ext}R_L^{\,2}}{1+\left(\omega C_{ext}R_L\right)^2} + \frac{x_{inv}}{\omega} - 8L_{sec}\left(1-k\right) \quad (2),$$

where w is the angular frequency of inverter, R_L is the load resistance, r_{inv} is the real part of the output impedance at output terminal of each inverter, L_{sec} is the self-inductance on secondary winding of each transformer, and k is the coupling factor of each transformer. Then, the L_{ext} is also expressed as

$$L_{ext} = \frac{2C_{ext}R_L^{\,2}}{1+\left(\omega C_{ext}R_L\right)^2} + \frac{x_{inv}}{\omega} - 2l_1 \quad (3),$$

where l_1 is the leakage inductance on primary winding of each transformer. The leakage inductance l_1 is expressed by

$$l_1 = L_{pri}\left(1-k\right) \quad (4),$$

where L_{pri} is the self-inductance on the primary winding of each transformer.

B. Current Balancer and Equivalent Circuit

The current balancer consists of two transformers. The primary winding is connected to each output of the inverter. The secondary winding is connected in series and connected to a resonant load. The turns ratio to achieve the current balance is derived theoretically in [5] as

$$\frac{N_1}{N_2} = 2 \quad (5),$$

where N_1 and N_2 are the turns number on the primary and secondary winding respectively.

Figure 2 shows the equivalent circuit of current balancer using ideal transformer and mutual inductor. The current balance effect by two transformers is analyzed using equivalent circuit. Then, l_2 is the leakage inductance on secondary winding and M is a mutual inductance. Each parameter of two transformers are assumed to be identical.

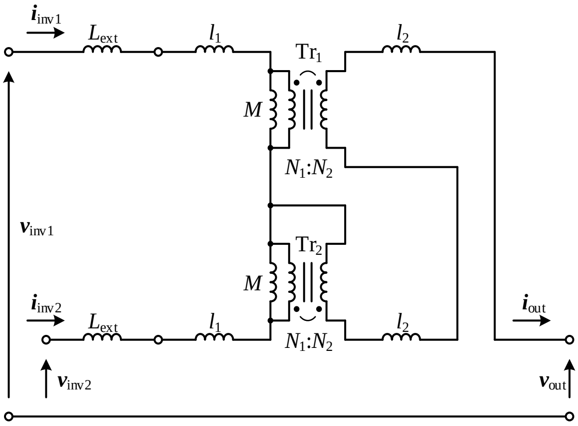


Fig. 2. Equivalent circuit of proposed balancer.

C. Current Balance Effect by Mutual Inductance

A fundamental component, which is resonant frequency of the load, just relates to the output power because of the resonant load. Thus, each voltage and current are assumed to be sinusoidal in the following derivation.

The voltage equations of each transformer are expressed by

$$\begin{aligned} \boldsymbol{v}_{inv1} &= V_{rms}\left\{\cos\left(\frac{\phi}{2}\right) + j\sin\left(\frac{\phi}{2}\right)\right\} \\ &= j\omega\left(L_{ext}+l_1+M\right)\boldsymbol{i}_{inv1} - j\omega\frac{N_2}{N_1}M\left(\boldsymbol{i}_{inv1}+\boldsymbol{i}_{inv2}\right) + j2\omega l_2\boldsymbol{i}_{out} + \boldsymbol{v}_{out} \end{aligned} \quad (6),$$

$$\begin{aligned} \boldsymbol{v}_{inv2} &= V_{rms}\left\{\cos\left(\frac{\phi}{2}\right) - j\sin\left(\frac{\phi}{2}\right)\right\} \\ &= j\omega\left(L_{ext}+l_1+M\right)\boldsymbol{i}_{inv2} - j\omega\frac{N_2}{N_1}M\left(\boldsymbol{i}_{inv1}+\boldsymbol{i}_{inv2}\right) + j2\omega l_2\boldsymbol{i}_{out} + \boldsymbol{v}_{out} \end{aligned} \quad (7),$$

where V_{rms} is the RMS value of the fundamental component of the inverter output voltage, φ is a voltage phase difference between two inverter outputs. The RMS value is assumed to be same value between two inverters.

Figure 3 shows the equivalent circuit of load with the external capacitor C_{ext}. The resonant inductor and capacitor are ignored because the reactance of the resonant tank is zero at the resonant frequency. Then, the output voltage v_{out} is expressed as

$$\boldsymbol{v}_{out} = \left(R_{out} + jX_{out}\right)\boldsymbol{i}_{out} \quad (8),$$

$$R_{out} = \frac{R_L}{1+\left(\omega C_{ext} R_L\right)^2} \quad (9),$$

$$X_{out} = -\frac{\omega C_{ext} R_L{}^2}{1+\left(\omega C_{ext} R_L\right)^2} \quad (10).$$

where R_{out} and X_{out} are real part and imaginary part of load-side impedance including external capacitor C_{ext} respectively. Then, the amplitude of output current i_{out} is derived using (5), (6), (7), and (8) by

$$\left|\boldsymbol{i}_{out}\right| = \frac{V_{rms}}{\sqrt{R_{out}{}^2 + \left\{X_{out} + \omega\left(\frac{L_{ext}}{2} + l_1\right)\right\}^2}}\cos\left(\frac{\phi}{2}\right) \quad (11).$$

The amplitude of each output current is also expressed as

$$\left|\boldsymbol{i}_{inv1}\right| = \sqrt{\left(A+B\right)^2 + C^2} \quad (12),$$

$$\left|\boldsymbol{i}_{inv2}\right| = \sqrt{\left(A-B\right)^2 + C^2} \quad (13),$$

$$\left(\begin{aligned} A &= \frac{R_{out}}{R_{out}{}^2 + \left\{X_{out} + \omega\left(\frac{L_{ext}}{2} + l_1\right)\right\}^2}\frac{V_{rms}}{2}\cos\left(\frac{\phi}{2}\right) \\ B &= \frac{V_{rms}}{\omega\left(L_{ext} + l_1 + M\right)}\sin\left(\frac{\phi}{2}\right) \\ C &= \frac{X_{out} + \omega\left(\frac{L_{ext}}{2} + l_1\right)}{R_{out}{}^2 + \left\{X_{out} + \omega\left(\frac{L_{ext}}{2} + l_1\right)\right\}^2}\frac{V_{rms}}{2}\cos\left(\frac{\phi}{2}\right) \end{aligned}\right) \quad (14).$$

The output current $\boldsymbol{i}_{out}$ is regulated by load side impedances, external inductor L_{ext}, and on primary-side leakage inductance l_1. On the other hand, output currents of each inverter are regulated by load side impedances, external inductor L_{ext}, primary-side leakage inductance l_1, and mutual inductance M. Although the mutual inductance M does not affect to the load current i_{out}, the mutual inductance M suppresses each output current.

D. Current Unbalance Rate

The suppression effect of the current difference between each output current is expressed as current unbalance rate. The current unbalance rate a is defined as

$$a = \frac{\left|\boldsymbol{i}_{inv1}\right| - \left|\boldsymbol{i}_{inv2}\right|}{\left|\boldsymbol{i}_{out}\right|} \times 100 \quad (15).$$

Table 1 shows parameters, which are used in the calculation of the unbalance rate. The switching frequency is set to 6.78 MHz, which corresponds to that of the target WPT system. The output impedance of each inverter is determined by the output capacitance of the GaN devices in each inverter.

Figure 4(a) illustrates the current unbalance rate with the mutual inductance variation. The mutual inductance is changed by decreasing the coupling factor k of transformer. Then, the leakage inductance of transformer is also changed by coupling factor k. Thus, the external inductor is modified based on (3) to ensure the impedance matching. The unbalance rate is suppressed by increasing inductance. Figure 4(b) shows the current unbalance rate with the voltage phase variation between inverter#1 and inverter#2. The unbalance rate increases with the voltage phase difference.

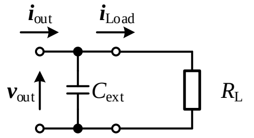


Fig. 3. Proposed balancer.

Table 1. Calculation parameters.

Circuit		
Switching frequency	f_s	6.78 MHz
Output reistance of each inverter	r_{inv}	10 Ω
Output leactance of each inverter	x_{inv}	10 Ω
Load resistance	R_L	50 Ω
Transformer		
Self inductance on primary side	L_{pri}	2.0 μH
Self inductance on secondary side	L_{sec}	0.50 μH

E. Mutual Inductance Design

The mutual inductance M required to achieve the target current unbalance rate is designed using flowchart. Figure 5 shows the proposed flowchart to determine the mutual inductance M. The proposed flowchart also determines the external inductance L_{ext}. The target unbalance rate is defined as a_{set}. The target unbalance rate a_{set}, switching frequency f_s, output impedances of each inverter, load resistance R_L, voltage phase difference φ, and coupling factor k are set as the design requirement.

Table 2 shows the example of the design requirement. The inductance step is set to 85 nH in this paper. As a result, the required mutual inductance M is 850 nH and the external inductance L_{ext} is 339 nH based on the proposed flowchart and design requirement. Then, the current unbalance rate is 5.0%.

III. SIMLATION

The current balance performance of the proposed balancer is evaluated through simulation. Table 1 lists the simulation parameters. The designed mutual inductance M and the external inductance L_{ext} are applied to the simulation. The load parameters are calculated based on the resonant frequency, which is equal to the switching frequency. The simulation is conducted using PLECS (Plexim Inc.).

Figure 6 shows the simulation result of parallel operation of high-frequency inverters at 6.78 MHz with proposed balancer. The current RMS value of the inverter#1 and inverter#2 are 18.39 A and 17.50 A respectively. Then, the RMS value of the output current is 35.88 A. Thus, the current unbalance rate is 5.0 % based on (15). The designed current unbalance is

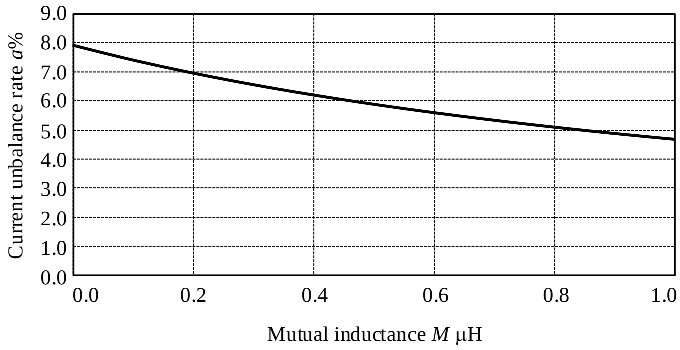
Table 2. Simulation parameters

Main circuit		
DC link voltage	V_{DC}	300 V
Switching frequency	f_s	6.78 MHz
Deadtime	t_d	25 ns
Load		
Resonant inductance	L_r	5.3 μH
Resonant capacitance	C_r	104 pF
Road resistance	R_L	50 Ω
Balancer		
Inductance of primary side	L_{pri}	2.0 μH
Inductance of secondary side	L_{sec}	0.50 μH
Cupling factor of transformer	k	0.85
Mutual inductance	M	850 nH
Impedance matching		
Inductance of primary side	L_{ext}	339 nH
Inductance of secondary side	C_{ext}	1.12 nF

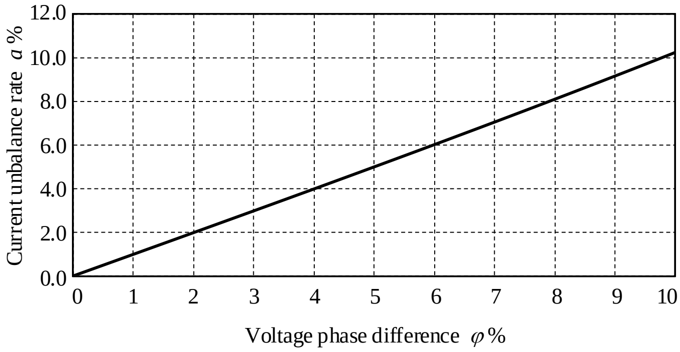
achieved by proposed balancer and parameter design flowchart. Then, the output power of inverter#1 and inverter#2 are 2.87-kW and 3.57 kW respectively. The total output power is 6.44-kW. Thus, the kW-order MHz band inverter is achieved by proposed balancer.

IV. Conclusion

This paper proposed a parameter design method for a current balancer used in paralleled MHz-band inverters for WPT systems. The current balancer, consisting of two transformers, was designed to achieve current sharing between inverters. The relationship between the mutual inductance and the current unbalance rate was analytically derived, and a design flowchart for determining the mutual inductance M and the external inductance L_{ext} was proposed. The effectiveness of the proposed design was verified through simulation. Using the designed parameters, the current unbalance rate was suppressed to 5.0% at an operating frequency of 6.78 MHz, resulting in a total output power of 6.44 kW.



(a) with variation of mutual inductance



(b) with variation of voltage phase difference

Fig. 4. Current unbalance rate.

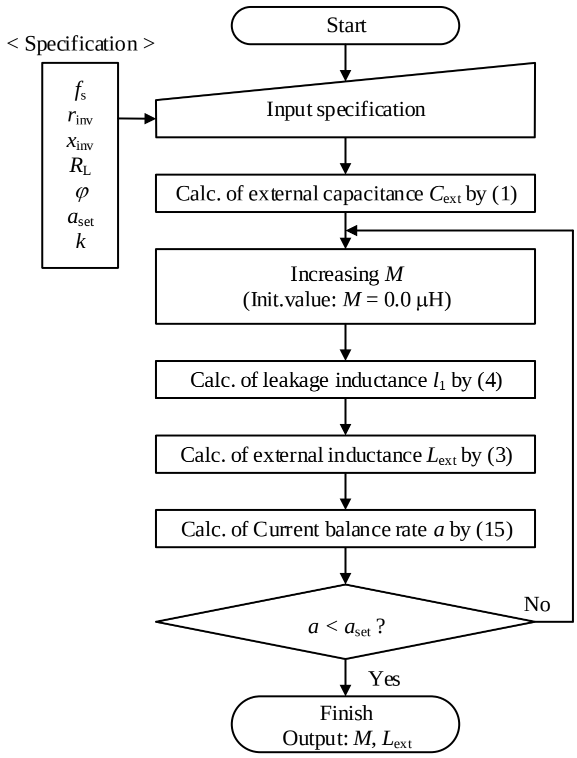


Fig.5. Parameter design flowchart.

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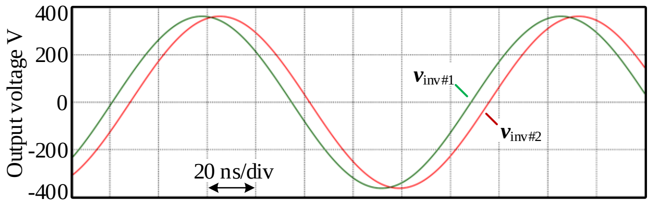
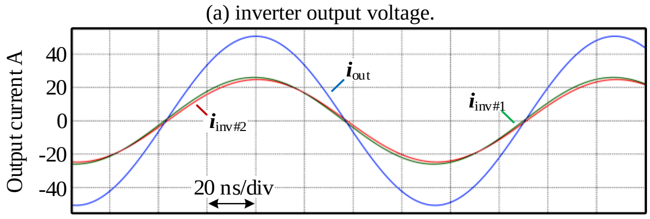


(a) inverter output voltage.



(b) each output current.

Fig. 6. Simulation result with phase difference ($\varphi = 5\%$).