

# A Single-Stage IPT Converter with Optimal Efficiency Tracking and Constant Voltage Output Against Dynamic Variations of Coupling and Load

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**Abstract**—Multistage design based inductive-power-transfer (IPT) systems have been carried out to cope with power regulation and optimal efficiency tracking against variations of load conditions and coupling coefficients in dynamic scenarios like roadway-powered electric vehicles. However, it is still challenging for a single-stage IPT converter to do so, due to the fixed transfer ratio and narrow optimal load range of the resonant tank. This paper aims to fill the gap by integrating active switches within the IPT converter to extend the modulation range. A new series-series compensated IPT converter is proposed, which, unlike the conventional ones, employs two switched-controlled capacitors (SCCs) and a semi-active rectifier (SAR). The secondary SCC and the SAR cooperate via an inner control loop to emulate a null secondary impedance and an optimal load resistance, while the primary SCC is responsible for the output power regulation via an outer control loop. The operating principle enables a constant voltage output and optimal efficiency tracking against dynamic variations of coupling coefficient and load condition, while fixed operating frequency and soft switching are also permitted. Experimental results validate the performance.

**Index Terms**—Constant Voltage, Dynamic, Inductive Power Transfer, Optimal Efficiency, Single-Stage.

## I. INTRODUCTION

The development of inductive power transfer (IPT) technology has enabled contactless power transfer over an air gap distance via magnetic coupling [1], [2]. IPT is suitable for power supply applications where physical contact is inconvenient or hazardous. In stationary scenarios, like wireless chargers for consumer electronics, mine appliances, and electric vehicles (EVs) in garages, IPT has shown significant benefits in providing safe, user-friendly, and maintenance-free operations and has become standardized in recent years [3]–[5]. In contrast to

applications in stationary scenarios, IPT performs favorably in dynamic scenarios, being able to truly free the power supply, for instance, in roadway-powered EVs [6], wireless excited synchronous machines [7], biomedical implants [8] and so on. However, using IPT systems in dynamic scenarios inevitably incurs more severe parameter fluctuation, which is challenging for efficient and stable power transfer.

The loosely-coupled transformer (LCT), which is formed by the coupled transmitter and receiver coils, is essential for power transfer in IPT systems. A variety of studies have been carried out to design and optimize the LCT, aiming at improving the efficiency of wireless power transfer. The transmitter and receiver coils should first have interoperable structures in order to couple well with each other [9]. Optimization of the LCT towards a high coupling coefficient ( $k$ ) and coil quality factors ensure efficient wireless power transfer [10], [11]. Design for misalignment tolerance has also been considered by using more coils to shape desired magnetic flux, such as double-D quadrature pad [12] and tripolar pad [13]. They incur additional material cost or higher manufacturing complexity.

For the misalignment issues in stationary applications,  $k$  usually varies in a small range and becomes fixed during the power transfer. In contrary, continuous and severe longitudinal, lateral, angular and rotational motions occur between the transmitter and receiver coils in dynamic applications, which leads to wide-range variation of  $k$  that be fully addressed via the design of LCT. Moreover, to achieve optimal efficiency, a well-compensated LCT should be loaded at a very narrow vicinity of load condition, which has been studied and known to be  $k$ -dependent [14]–[17]. Once the well compensated LCT is not optimally loaded, which may be caused by  $k$  variation or load change, it will suffer from the dramatic efficiency degradation. Therefore, an appropriate IPT converter topology and an effective control method are needed to achieve optimal efficiency and constant output against the variations of  $k$  and load condition.

Multistage designs have been first carried out to address the above issues [17]–[23]. The diagram shown in Fig. 1(a) illustrates the operating principle of these multistage designs. In these designs, an IPT converter is cascaded with a front-end DC-DC converter and a load-side DC-DC converter. Depending on the choice of compensation circuitry, the IPT converter can act as a voltage gain, transadmittance, current gain, or transresistance converter. For example, the four basic IPT

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TABLE I  
COMPARISON OF STATE-OF-THE-ART WORKS

| Desirable merits                      | Multistage<br>[17]–[25] | Hopping methods<br>[26]–[30] | Hybrid topologies<br>[31], [32] | Dual H-bridges<br>[33], [34] | FM+SAR<br>[36], [37] | SCC+SAR<br>[38] | Proposed |
|---------------------------------------|-------------------------|------------------------------|---------------------------------|------------------------------|----------------------|-----------------|----------|
| Constant output against load change   | ✓                       | ✓                            | ✓                               | ✓                            | ✓                    | ✓               | ✓        |
| Constant output against $k$ variation | ✓                       | ×                            | ✓ (Nearly)                      | ✓                            | ×                    | ×               | ✓        |
| Load matching for optimal efficiency  | ✓                       | ×                            | ×                               | ✓                            | ✓                    | ✓               | ✓        |
| Single-stage                          | ×                       | ✓                            | ✓                               | ✓                            | ✓                    | ✓               | ✓        |
| Soft switching                        | ×                       | ✓                            | ✓                               | ×                            | ✓                    | ✓               | ✓        |
| Fixed operating frequency             | ✓                       | depends                      | ✓                               | ✓                            | ×                    | ✓               | ✓        |

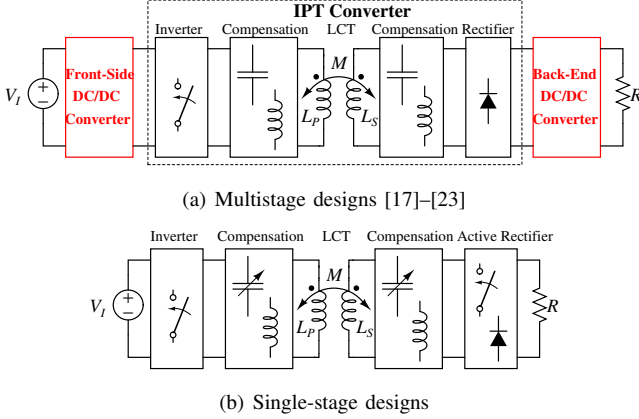


Fig. 1. Diagrams of IPT systems that can achieve optimal efficiency and constant output against variations of  $k$  and load condition.

converters with series-series, series-parallel, parallel-series and parallel-parallel compensation circuits have constant  $I_{\text{Out}}/V_{\text{In}}$ ,  $V_{\text{Out}}/V_{\text{In}}$ ,  $I_{\text{Out}}/I_{\text{In}}$  and  $V_{\text{Out}}/I_{\text{In}}$  ratios respectively [17]. Without any modulation, the efficiency and output of the IPT converter suffer from the variations of  $k$  and load condition. Thus, the load-side DC-DC converter adaptively controls the equivalent load observed by the IPT converter to track the optimal efficiency, while the input-side modulation is responsible for the front-end DC-DC converter to achieve output regulation. Different control schemes have been studied, such as the minimum input current tracking [18], the maximum efficiency tracking [19]–[22], and the voltage/current ratio control [17], [23]. Based on the above operating principle, the IPT converter is readily to track the optimal efficiency and maintain constant output under variations of  $k$  and load condition. However, additional losses and costs of the DC-DC converters are inevitable, and moreover, complex control is needed to ensure stability and fast response of the cascading DC-DC converters [24], [25].

To avoid the penalties brought by multistage designs, significant research efforts have been directed toward single-stage designs. Under a fixed  $k$ , hopping the operating frequencies [26], [27] or the compensation topologies [28]–[30] are two common ways to realize constant current (CC) or constant voltage (CV) outputs against a wide range of loads. Taking it one step further, hybrid IPT topologies, including input-parallel-output-parallel, input-parallel-output-series, input-series-output-parallel, and input-series-output-series, are constructed for a CC or CV output with misalignment tolerance [31], [32]. Nevertheless, without deep modulation in the

rectifier, the above methods cannot realize load matching and optimize efficiency performance against variations of  $k$  and load condition.

With additional active switches integrated within the IPT converter, as shown in Fig. 1(b), the modulation given by the additional DC-DC converter can be alternatively implemented by the inverter and the rectifier, such that the DC-DC converters can be eliminated. However, it is still challenging to achieve wide range modulation in the input and output terminals, while implementing soft switching in the inverter and the rectifier. In [33], [34], dual H-bridges are used, and hard switching loss is allowed to extend the modulation range of the inverter and rectifier, without changing the resonance condition of the resonant tank circuit, so as to achieve output regulation and load matching against variations of  $k$  and load condition. It is noteworthy that the efficiency degrades significantly with the increase of modulation depth. Soft-switching modulation of voltage-controlled and current-controlled semi-active rectifiers (SARs) has been investigated [35], [36]. These SARs transform the load into an equivalent impedance that includes not only a resistive component but also a reactive component. They provide control freedom for load matching but may also worsen the resonance condition in the secondary. To address this issue, frequency modulation (FM) in the primary cooperates with the load transformation to simultaneously realize load matching and a CC or CV output [36], [37]. Nevertheless, it cannot cope with the dynamic  $k$  variation or misalignment issues, and thus more control freedom is desired for the IPT systems. The switch-controlled capacitor (SCC) technique is found to be another way to cancel the undesired reactance generated by the SARs, so as to maintain the resonance condition in the secondary [38]. Without any modulation in the primary, the proposed wireless power supply in [38] can maintain maximum efficiency against load change but still fails to cope with the dynamic  $k$  variation or misalignment issues.

To the best of our knowledge, an IPT converter that features the following merits has not yet been developed, as summarized in Table I.

- 1) Single-stage design
- 2) CV output and optimal efficiency tracking against dynamic variations of  $k$  and load condition
- 3) Soft switching and fixed operating frequency

To fill the gap, this paper proposes a series-series compensated IPT converter, which, unlike the conventional ones, employs two SCCs and a semi-active rectifier. The operating principles, including  $k$ -independent criteria for optimal efficiency

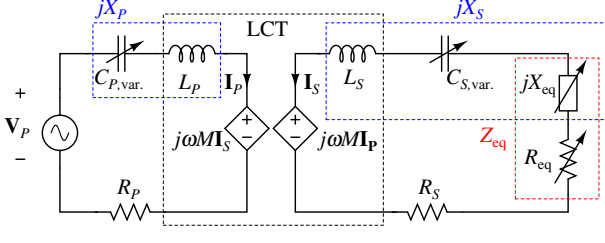


Fig. 2. General equivalent circuit model of the widely-used series-series compensated IPT converter.

and regulation of output power, are first introduced in Section II, based on which, the schematic, modulation, and control of the proposed IPT converter are detailed in Section III. The performances of the proposed IPT converter are experimentally verified in Section IV. Finally, Section V concludes this paper.

## II. OPERATING PRINCIPLE

An equivalent circuit model of the widely-used series-series compensated IPT converter is shown in Fig. 2. As usual, the loosely-coupled transformer has primary self-inductance  $L_P$ , secondary self-inductance  $L_S$  and mutual inductance  $M$ . The coupling coefficient is given by  $k = M/\sqrt{L_P L_S}$ .  $R_P$  and  $R_S$  are equivalent series resistors that represent losses in the primary and secondary respectively. For a more general discussion, the primary and secondary compensation capacitors are both variables, denoted as  $C_{P,var.}$  and  $C_{S,var.}$  respectively, while the load is an impedance that includes resistive and reactive components, given by  $Z_{eq} = R_{eq} + jX_{eq}$ . Implementation of variable  $C_{P,var.}$ ,  $C_{S,var.}$  and  $Z_{eq}$  will be detailed in subsequent sections.  $v_p$ ,  $i_p$ ,  $v_s$  and  $i_s$  are fundamental components, and  $\mathbf{V}_P$ ,  $\mathbf{I}_P$ ,  $\mathbf{V}_S$  and  $\mathbf{I}_S$  are vector representation of fundamental components respectively. Circuit equations are

$$\mathbf{V}_P = \mathbf{I}_P(jX_P + R_P) + jX_M \mathbf{I}_S, \text{ and} \quad (1)$$

$$-\mathbf{I}_S R_{eq} = \mathbf{I}_S(jX_S + R_S) + jX_M \mathbf{I}_P, \quad (2)$$

where

$$X_P = \omega L_P - \frac{1}{\omega C_{P,var.}}, \quad (3)$$

$$X_S = \omega L_S - \frac{1}{\omega C_{S,var.}} + X_{eq}, \text{ and} \quad (4)$$

$$X_M = \omega M. \quad (5)$$

### A. $k$ -Independent Criteria for Optimal Efficiency

The efficiency  $\eta$  can be calculated by (6). By substituting (1) to (5) into (6),  $\eta$  can be expressed by (7). As a high-quality IPT converter usually satisfies  $X_M^2/(R_P R_S) \gg 1$  and  $R_{eq}/R_S \gg 1$  [37], (7) can be further simplified as (8).

$$\eta = \frac{|\mathbf{I}_S|^2 R_{eq}}{|\mathbf{I}_S|^2 (R_{eq} + R_S) + |\mathbf{I}_P|^2 R_P} \quad (6)$$

$$= \frac{X_M^2 R_{eq}}{[(R_{eq} + R_S)^2 + X_S^2] R_P + X_M^2 (R_{eq} + R_S)} \quad (7)$$

$$\approx \frac{1}{\frac{R_{eq} + \frac{X_S^2}{R_{eq}}}{X_M^2} R_P + \frac{R_S}{R_{eq}} + 1} \quad (8)$$

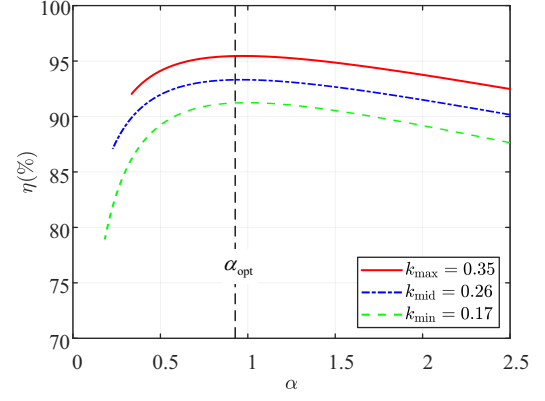


Fig. 3. Efficiency  $\eta$  versus current ratio  $\alpha$  under different values of  $k$ . (Parameters used for simulation are:  $L_P = 147.23 \mu\text{H}$ ,  $L_S = 148.5 \mu\text{H}$ ,  $R_P = 0.4 \Omega$ , and  $R_S = 0.49 \Omega$ . Unless specified, they will be used for the rest simulation.)

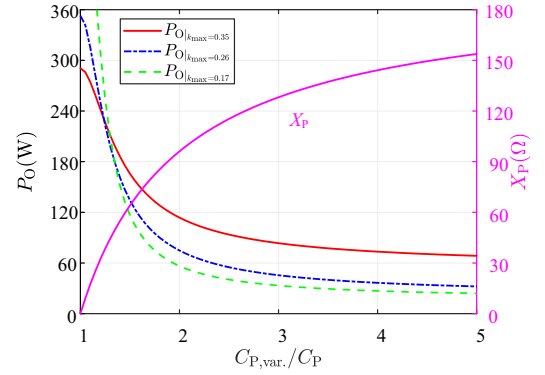


Fig. 4. Output power  $P_O$  and primary reactance  $X_P$  versus primary compensation capacitor  $C_{P,var.}$ .

Optimum values of  $R_{eq}$  and  $X_S$  that achieve maximum efficiency can be found from (8). If the operating frequency  $\omega$  is fixed, by solving the partial differentials  $\frac{\partial \eta}{\partial X_S} = 0$  and  $\frac{\partial \eta}{\partial R_{eq}} = 0$ , optimal efficiency  $\eta_{max}$  can be achieved as

$$\eta_{opt} = \frac{1}{\frac{2}{Q_P Q_S} + 1}, \text{ if} \quad (9)$$

$$X_{S,opt} = \omega L_S - \frac{1}{\omega C_{S,var.}} + X_{eq} = 0, \text{ and} \quad (10)$$

$$R_{eq,opt} = X_M \sqrt{\frac{R_S}{R_P}}. \quad (11)$$

where  $Q_P = (\omega L_P)/R_P$  and  $Q_S = (\omega L_S)/R_S$  are quality factor representation. Criteria for maximum efficiency given in (10) and (11) indicate that the secondary resonant tank should have null equivalent reactance and the equivalent load resistance should be matching.

From (10), null secondary reactance can be maintained by varying  $C_{S,var.}$  to offset  $X_{eq}$ . From (11), it can be observed that  $R_{eq,opt}$  is  $k$ -dependent and thus should be correspondingly varied with the change of  $k$  for optimal efficiency in dynamic scenarios. However, since it is difficult to directly detect  $k$  in practice, a viable condition for optimal efficiency should be derived. Given null secondary reactance can be satisfied, i.e.,

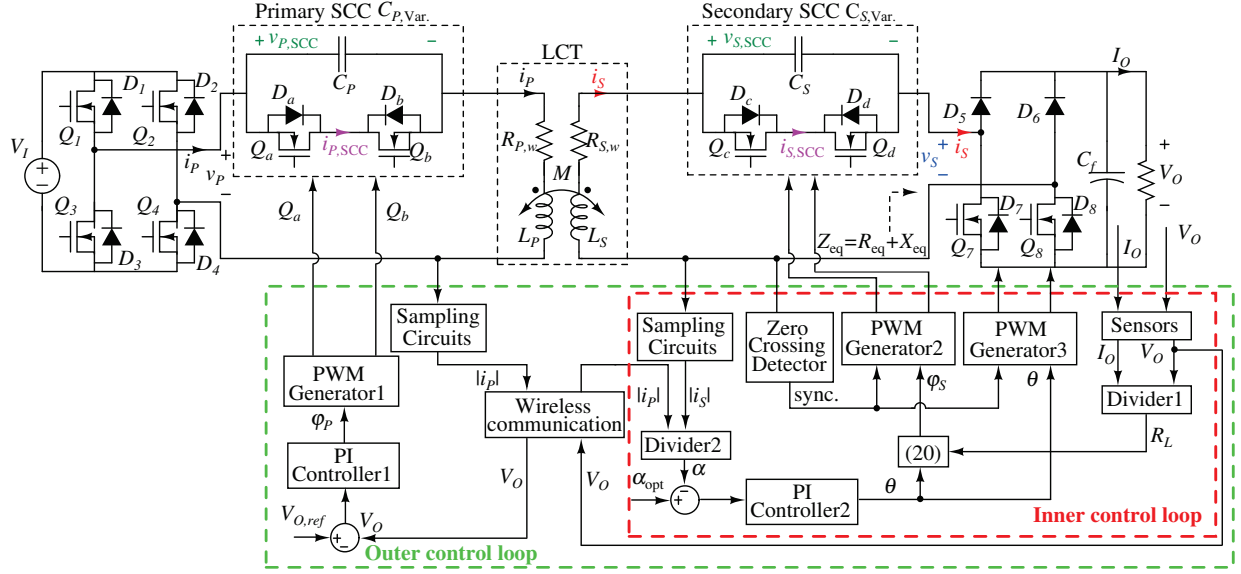


Fig. 5. Schematics and control diagram of the proposed IPT converter.

(10) is satisfied, the amplitude ratio of secondary-side current to primary-side current can be defined as

$$\alpha = \frac{|\mathbf{I}_S|}{|\mathbf{I}_P|} = \frac{X_M}{(R_{eq} + R_S)}. \quad (12)$$

With (6) and (12), typical curves of efficiency  $\eta$  versus current ratio  $\alpha$  are plotted in Fig. 3. It can be observed that there are efficiency peaks under different values of  $k$ . By substituting  $X_{S,opt}$  and  $R_{eq,opt}$  given by (10) and (11) into (2), the optimal  $\alpha$  can be derived as

$$\alpha_{opt} = \frac{|\mathbf{I}_S|}{|\mathbf{I}_P|} \approx \sqrt{\frac{R_P}{R_S}}. \quad (13)$$

Here, (13) represents the  $k$ -independent criterion that can achieve the optimal efficiency regardless of  $k$ . This means that direct detection of the coupling coefficient is not needed in the proposed scheme, which is an advantage over existing methods. As shown in Fig. 3, if  $\alpha$  is maintained at  $\alpha_{opt}$ , the IPT converter can always operate at its optimal efficiency under different  $k$ . Since the current amplitudes can be readily detected, (13) is a viable condition for tracking the optimal efficiency. It is worth noting that the proposed control scheme relies on the amplitude information, while the phase information of these currents is not necessary. Thus, it offers practical advantages and simplifies the implementation of current sampling.

### B. Regulation of Output Power

In addition to maintaining optimal efficiency, wide-range output power is desired in dynamic IPT. Since the modulation in the secondary has been responsible for the optimal efficiency, primary-side tuning will be investigated for the regulation of output power. Assuming  $R_P = R_S = 0$ , i.e.,

lossless power transfer, the output power can be derived with (1), (2), (10) and (11), and it is given by

$$P_O = \left| \frac{\mathbf{I}_S}{\sqrt{2}} \right|^2 R_{eq,opt} = \frac{|\mathbf{V}_P|^2 X_M}{2(X_P^2 + X_M^2)}. \quad (14)$$

If the operating frequency  $\omega$  is fixed,  $P_O$  can only be regulated via the change of  $X_P$ . According to (3), variable  $C_{P,var}$  is therefore required. Fig. 4 shows the output power  $P_O$  together with the primary reactance  $X_P$  along with the variation of primary compensation capacitor  $C_{P,var}$ . Wide-range output power can be readily achieved by varying  $C_{P,var}$ .

## III. SCHEMATIC, MODULATION AND CONTROL OF THE PROPOSED SINGLE-STAGE IPT CONVERTER

### A. Schematic

Based on the fundamental circuit model shown by Fig. 2 and the operating principle illustrated in Section II, a single-stage IPT converter is proposed as depicted in Fig. 5. It includes a full-bridge inverter, two switch-controlled capacitors (SCCs), an LCT and a semi-active rectifier (SAR). The SAR is used to transform the load resistance  $R_L$  that varies in a wide range into the optimal value  $R_{eq,opt}$  for the optimal efficiency  $\eta_{opt}$  according to (11). However, to realize soft-switching modulation in the SAR, which will be detailed in Section III-B, a variable equivalent impedance  $X_{eq}$  that violates (10) will be incurred with the variation of  $R_L$ . The secondary SCC helps to offset  $X_{eq}$ , such that (10) can be obeyed for optimal efficiency. Another control variable for output power regulation is introduced by the primary SCC.

### B. Soft-Switched Modulations and Models of SAR and SCC

The SAR has two diodes ( $D_5, D_7$ ) in the upper legs and two MOSFET switches ( $Q_6, Q_8$ ) in the lower legs. The switching sequences and the operating waveforms of the SAR are shown in Fig. 6.  $Q_6$  and  $Q_8$  are used to control the conduction angle

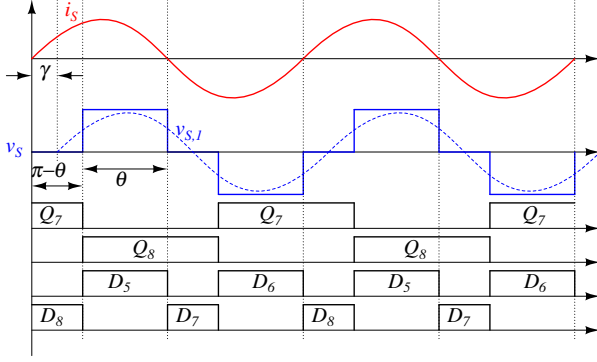


Fig. 6. Switching sequences and operating waveforms of the SAR.

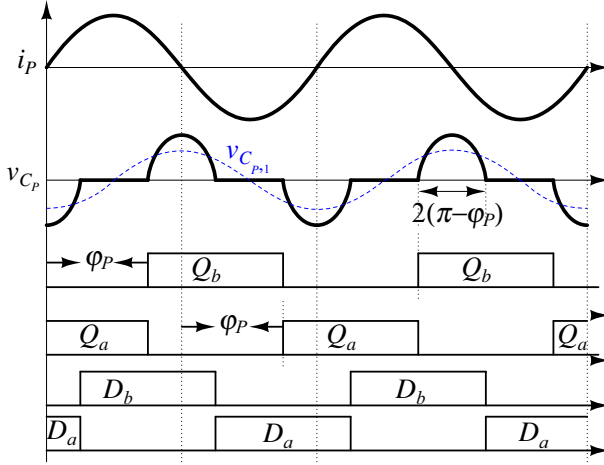


Fig. 7. Switching sequences and operating waveforms of the SCCs.

of  $v_S$  denoted by  $\theta$ , and they complement each other.  $Q_6$  is turned ON for half a cycle with a phase delay  $(\pi - \theta) \in [0, \pi]$  to the zero-crossing point where  $i_S$  commutates from negative to positive. Then,  $Q_8$  operates in a similar way. The zero-crossing detection can be implemented by using a current transformer and comparators, and the zero-crossing signal is then used to trigger the ePWM module of the TMS320F28335 DSP controller to synchronize the PWM generation. The ePWM module of the TMS320F28335 DSP controller uses a 150 MHz clock and can capture the 85 kHz zero-crossing signal. The conduction angle  $\theta$  of the SAR ranges from 0 to a maximum  $\pi$ , i.e.,  $\theta \in [0, \pi]$ . It can be observed that  $Q_6$  and  $Q_8$  are turned ON during the ON-time of their antiparallel diodes to enjoy zero-voltage switching (ZVS). The change of  $\theta$  will lead to a phase angle between  $v_S$  and  $i_S$ . As shown by the dashed blue curve in Fig. 6,  $v_{S,1}$  is the fundamental component of  $v_S$ , and it lags  $i_S$  with a phase angle given by  $\gamma = \frac{\pi - \theta}{2}$ . Therefore, the equivalent load is an impedance rather than a pure resistance, as given by (15)–(17). The derivation can be

found in [35], [37], [39].

$$Z_{eq} = R_{eq} + jX_{eq}, \text{ where} \quad (15)$$

$$R_{eq} = \frac{8}{\pi^2} R_L \sin^4\left(\frac{\theta}{2}\right), \text{ and} \quad (16)$$

$$X_{eq} = -\frac{8}{\pi^2} R_L \sin^3\left(\frac{\theta}{2}\right) \cos\left(\frac{\theta}{2}\right). \quad (17)$$

Considering the similarity in operational principles between the primary and secondary SCC modules, the primary SCC is employed as an example to illustrate the operating mechanism. The primary SCC consists of a capacitor  $C_P$  in parallel with two anti-connected MOSFET switches  $Q_a$  and  $Q_b$ . The switching sequences and the operating waveforms of the SCC are shown in Fig. 7.  $Q_a$  and  $Q_b$  are used to shape the terminal voltage of  $C_P$ , and they complement each other.  $Q_b$  has a phase shift  $\varphi_P \in [\frac{\pi}{2}, \pi]$  from the zero-crossing point where  $i_P$  commutates from negative to positive. Then,  $Q_a$  operates in a similar way.  $Q_a$  and  $Q_b$  are turned ON and OFF at zero voltage for low switching loss.  $v_{C_P}$  is shaped as part of a sine wave, and the effective charging/discharging time of  $C_P$  in half a cycle is  $2(\pi - \varphi_P)$ . As the dashed curve labeled with  $v_{C_P,1}$  shows, the increase of  $\varphi_P$  will decrease the peak of the fundamental component of  $v_{C_P}$ . Consequently, the equivalent capacitance  $C_{P,var.}$  of the SCC can be varied by the phase shift angle  $\varphi_P$ . Equivalent  $C_{P,var.}$  has been studied and given by (18) [40], and it can be simplified as a polynomial given by (19) for easy calculation.

$$C_{P,var.} = \frac{C_P}{2 - \frac{2\varphi_P - \sin 2\varphi_P}{\pi}} \quad (18)$$

$$\approx \frac{\pi^2}{4(\varphi_P - \pi)^2} C_P. \quad (19)$$

### C. Inner Loop Control for Optimal Efficiency

The objective of the secondary SCC is to cancel the equivalent load reactance  $X_{eq}$  generated by the SAR and maintain null reactance in the secondary. To satisfy (10),  $\varphi_S$  should be cooperatively varied with  $\theta$ . With (10), (17) and (19), the expression of phase-shift angle  $\varphi_S$  can be derived as (20).

$$\varphi_S = \pi - \frac{\pi}{2} \sqrt{C_S \left( \omega_0^2 L_S - \omega_0 \frac{8}{\pi^2} \sin^3 \frac{\theta}{2} \cos \frac{\theta}{2} R_L \right)} \quad (20)$$

With  $X_{eq}$  being canceled by the secondary SCC, the SAR is responsible for transforming  $R_L$  into  $R_{eq,opt}$  by varying the conduction angle  $\theta$  of the SAR. However, it is difficult to observe  $R_{eq}$  directly. Alternatively, as demonstrated in Section II-A, it is readily to observe the current ratio  $\alpha = |I_S/I_P|$ , and once  $\alpha$  is tightly regulated as  $\alpha_{opt} \approx \sqrt{\frac{R_P}{R_S}}$  given in (13),  $R_{eq,opt}$  can be achieved regardless of  $k$  variation. Therefore,  $\theta$  is controlled to achieve  $\alpha_{opt}$ . Only the amplitude of the high-frequency current is needed for the control, and the sampling rate is the same as the control frequency of the inner loop, which is far lower than the resonant frequency. Thus, there is no difficulty in the current measurement.

Moreover, by solving the partial differential given in (21), it can be found that  $\alpha$  monotonically decreases with the increase



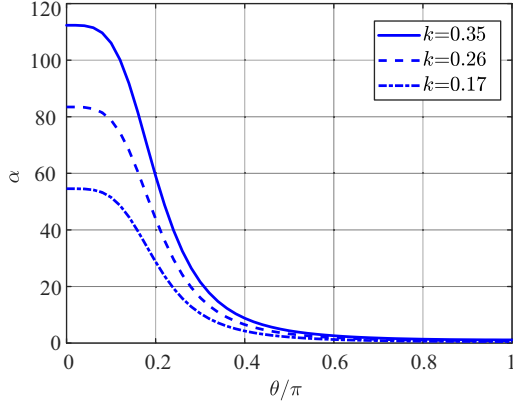


Fig. 8.  $\alpha$  versus  $\theta$  under different coupling coefficients.

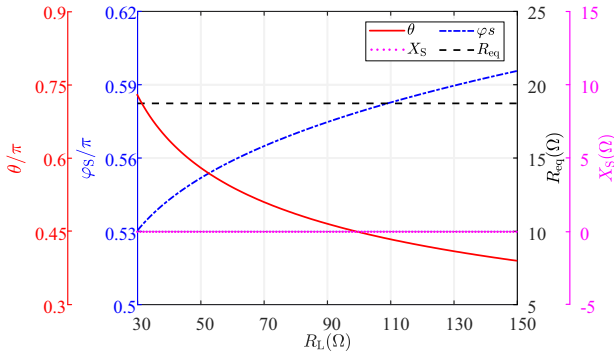


Fig. 9.  $\varphi_s$ ,  $\theta$ ,  $R_{eq}$  and  $X_S$  versus  $R_L$  ( $k = 0.26$ ).

of  $\theta$  as shown in Fig 8. Such that, a linear controller can be used to regulate  $\theta$ .

$$\frac{\partial \alpha}{\partial \theta} = -\frac{\pi^2 \omega_0 M}{16 R_L \sin^5(\frac{\theta}{2})} \cos(\frac{\theta}{2}) < 0, \text{ for } \theta \in [0, \pi]. \quad (21)$$

In practice,  $\alpha_{opt}$  should be precisely evaluated by including the losses of all switches into consideration. To make the key contents more concise, a detailed evaluation is given in the Appendix. It can be known that the practical  $\alpha_{opt}$  for our experimental prototype is around 0.9. Therefore,  $\alpha_{opt} = 0.9$  is used as the control reference.

As depicted in Fig. 9, the cooperation of the SAR and the secondary SCC meets the requirements of optimal efficiency under variations of  $k$  and load condition. The diagram of inner loop control can be designed and implemented as shown in the red box in Fig. 5. The output voltage  $V_O$  and output current  $I_O$  can be measured by sensors, and  $R_L$  can be calculated with a divider. Current transformers together with rectifiers are used to sample  $I_P$  and  $I_S$ , which are DC values. Wireless communication is used to transmit the value of  $I_P$  from the primary to the secondary for the calculation of  $\alpha$ . A simple PI controller is used to correct the difference between  $\alpha$  and  $\alpha_{opt}$ . The PI controller generates the control signal, i.e., the conduction angle  $\theta$ , for the SAR. Meanwhile, the phase-shift angle  $\varphi_s$  of the secondary SCC is generated according to (20).

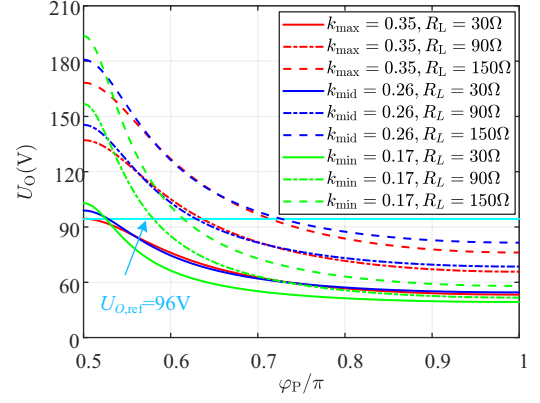


Fig. 10.  $V_O$  versus the phase-shift angle  $\varphi_P$  of the primary SCC.

#### D. Outer-Loop Control for Constant Voltage Output

The equivalent capacitance of the primary SCC is given by (22), where  $\varphi_P$  is the phase-shift angle of the primary SCC.

$$C_{P,var.} \approx \frac{\pi^2}{4(\varphi_P - \pi)^2} C_P \quad (22)$$

The primary reactance is therefore can be calculated by (23).

$$X_P = \omega L_P - \frac{4(\varphi_P - \pi)^2}{\pi^2 \omega C_P} \quad (23)$$

Under the cooperation of the SAR and the secondary SCC, the output power of the proposed system has be derived in (14). With (14) and (23), the output voltage of the proposed system can be calculated by (24).

$$V_O = \sqrt{P_O R_L} = \frac{2\sqrt{2}}{\pi} V_I \sqrt{\frac{R_{eq,opt} R_L}{X_P^2 + R_{eq,opt}^2}} \quad (24)$$

From (24),  $\varphi_P$  should be tightly regulated to maintain a CV output against the variations of  $k$  and load condition. By solving the partial differential (25),  $V_O$  monotonically decreases with the increase of  $\varphi_P$ , which is regardless of coupling coefficient  $k$  and load condition  $R_L$ .

$$\begin{aligned} \frac{\partial V_O}{\partial \varphi_P} &= \frac{16\sqrt{2}V_I X_P \sqrt{R_{eq,opt} R_L} (\varphi_P - \pi)}{\omega_0 \pi^3 C_P \sqrt{X_P^2 + R_{eq,opt}^2}} \\ &< 0, \text{ for } \varphi_P \in [\pi/2, \pi]. \end{aligned} \quad (25)$$

Fig 10 shows the simulated curves of  $V_O$  versus  $\varphi_P$ , and it also validates the monotonicity. Therefore, as shown in the green box in Fig. 5, another PI controller is readily implemented for the output voltage regulation in the outer control loop. The DC value of  $V_O$  is sampled in the secondary and wirelessly transmitted to the primary, and the PI controller corrects the difference between  $V_O$  and its reference  $V_{O,ref.}$ . The output of the PI controller is used to generates the phase-shift angle  $\varphi_P$  for the primary SCC.

#### IV. EXPERIMENTAL RESULTS

To verify the optimal efficiency tracking and CV output against variations of  $k$  and load condition, an experimental prototype is built as shown by Fig. 11. The system parameters

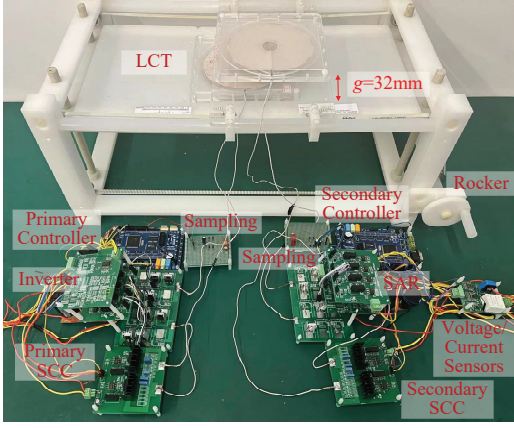


Fig. 11. Experimental prototype.

TABLE II  
SYSTEM PARAMETERS

| Parameters             | Symbols                       | Measured Values                                                        |
|------------------------|-------------------------------|------------------------------------------------------------------------|
| Input voltage          | $V_I$                         | 98 V                                                                   |
| Switches               | $Q_1$ - $Q_4$ , $Q_7$ , $Q_8$ | IPW65R190CFD with $R_{on} \approx 0.171\Omega$ and $V_F \approx 0.7$ V |
|                        | $Q_a$ , $Q_b$ , $Q_c$ , $Q_d$ | SCT4045DR with $R_{on} \approx 0.045\Omega$ and $V_F \approx 3.0$ V    |
| Diodes                 | $D_5$ , $D_6$                 | MBR20200CT with $V_F \approx 0.9$ V                                    |
| Number of turns        | $N$                           | 34                                                                     |
| Inner diameter         | $d_i$                         | 34 mm                                                                  |
| Outer diameter         | $d_o$                         | 152 mm                                                                 |
| Air gap distance       | $g$                           | 32 mm                                                                  |
| Self inductance        | $L_P$ , $L_S$                 | 147.23 $\mu$ H, 148.5 $\mu$ H                                          |
| Coil resistance        | $R_{P,w}$ , $R_{S,w}$         | 0.246 $\Omega$ , 0.3 $\Omega$                                          |
| Coupling coefficient   | $k$                           | 0.17 ~ 0.35                                                            |
| Primary compensation   | $C_P$                         | 25.68 nF                                                               |
| Secondary compensation | $C_S$                         | 25.4 nF                                                                |
| Operating frequency    | $\frac{\omega}{2\pi}$         | 85 kHz                                                                 |
| Load                   | $R_L$                         | 30 $\Omega$ -150 $\Omega$                                              |

are listed in Table II. The primary-side pad is fixed, while the horizontal displacement between the secondary-side pad and the primary-side pad is adjusted by rotating the rocker. The displacement ranges from 0 mm to 52 mm, and the coupling coefficient  $k$  varies from 0.35 to 0.17 correspondingly. An electronic load is used. Efficiency measurement is performed by Yokogawa WT1803E Precision Power Scope.

#### A. Steady-State Waveforms

Without loss of generality, the coupling coefficient is set at  $k = 0.26$ . Fig. 12 shows the steady-state waveforms under different load conditions. One 6-channel oscilloscope is used to capture the waveforms of the inverter, the primary SCC, the SAR, and the secondary SCC. Against the increase of load resistance, the phase difference between the primary current  $i_P$  and the secondary current  $i_S$  is fixed at  $0.5\pi$ , which means the secondary side is fully compensated. The conduction angle  $\theta$  of the SAR, phase-shift angle  $\varphi_S$  of the secondary SCC and phase-shift angle  $\varphi_P$  of the primary SCC can be observed. The waveforms are therefore consistent with the operating principle illustrated in Section III-B. Soft switching is permitted in the

primary SCC, the secondary SCC and the SAR. Since the proposed IPT converter generates a constant voltage (96 V) and operates over a wide range of load resistance (from 30  $\Omega$  to 150  $\Omega$ ), the range of output power is wide. The reactive power in the primary is utilized to facilitate the wide-range output power, but would not lead to an increase in voltage-ampere capacity of the IPT converter and hence would not harm the efficiency performance, as evidenced by (9)–(11).

#### B. Measured Efficiency and Current Ratio

The SAR and the secondary SCC are set to have close cooperation, which means the reactance generated by the SAR is canceled by the SCC via the control of  $\varphi_S$ . Such that, the secondary has null reactance, and there is only variation of equivalent load resistance  $R_{eq}$ . The variation of  $R_{eq}$  leads to different values of current ratio  $\alpha$ . Efficiency curves versus  $\alpha$  under different coupling coefficients are measured and plotted in Fig. 14. It can be observed that optimal efficiency can be achieved around  $\alpha = 0.9$  under all coupling coefficients and load conditions. It is consistent with the analysis in the Appendix, and  $\alpha = 0.9$  can be set as the control reference for the inner loop control. Furthermore, to clearly understand the practical distributed losses of the IPT system, the system power losses are measured and calculated for different cases. In Fig. 15, power loss breakdown of secondary-side SCC, SAR, coupling coils, primary-side SCC and inverter are represented by  $P_{S,SCC}$ ,  $P_{sar}$ ,  $P_i$ ,  $P_{P,SCC}$  and  $P_{inv}$ , respectively. The output voltage of the IPT system is maintained as 96 V, while the coupling coefficient and load resistance vary from 0.17 to 0.35 and from 30  $\Omega$  to 150  $\Omega$  respectively. It can be observed that the absolute power losses decrease with the decrease of the output power but it comprises a larger proportion, such that the maximum efficiency degrades in lighter load conditions.

#### C. Transient Waveforms Under Changing Coupling and Load

The transient waveforms under step load change for different values of  $k$  are shown in Fig. 16. Here, CH1 and CH2 are input AC voltage  $v_P$  and primary side current  $i_P$ , respectively. The DC output voltage  $V_O$ , output current  $I_O$  and secondary current  $i_S$  are measured in CH5 (brown), CH6 (orange) and CH7 (magenta), respectively. The load resistance is stepped between 30  $\Omega$  and 90  $\Omega$ . It can be observed that the current ratio is tightly regulated against step load change, and it is fixed at 0.9 via an inner loop control. The output voltage  $V_O$  is tightly regulated at 96 V via an outer loop control.

The transient waveforms under varying coupling coefficient in different load conditions are shown in Fig. 17. The coupling coefficient  $k$  is varied from 0.35 to 0.17 and back to 0.35 by slowly changing the displacement of the secondary pad to the primary pad. Similarly, CH1 (red) and CH2 (green) measure the input AC voltage  $v_P$  and primary side current  $i_P$ , respectively. The DC output voltage  $V_O$ , output current  $I_O$  and secondary current  $i_S$  are measured and shown in CH5 (brown), CH6 (orange) and CH7 (magenta), respectively. The current ratio  $\alpha$  and the output voltage  $V_O$  are still tightly regulated as 0.9 and 96 V against the variation of the coupling coefficient via an inner loop and an outer loop control, respectively.

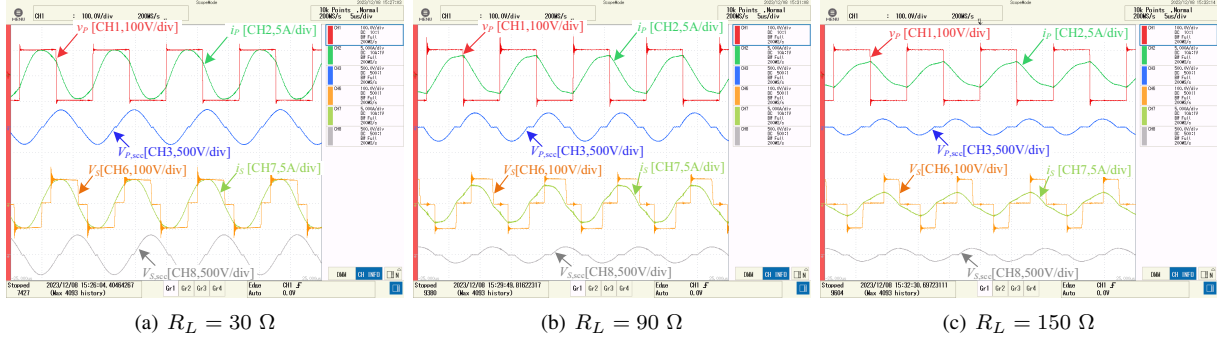


Fig. 12. Measured steady-state waveforms under different load conditions with coupling coefficient  $k = 0.26$ .

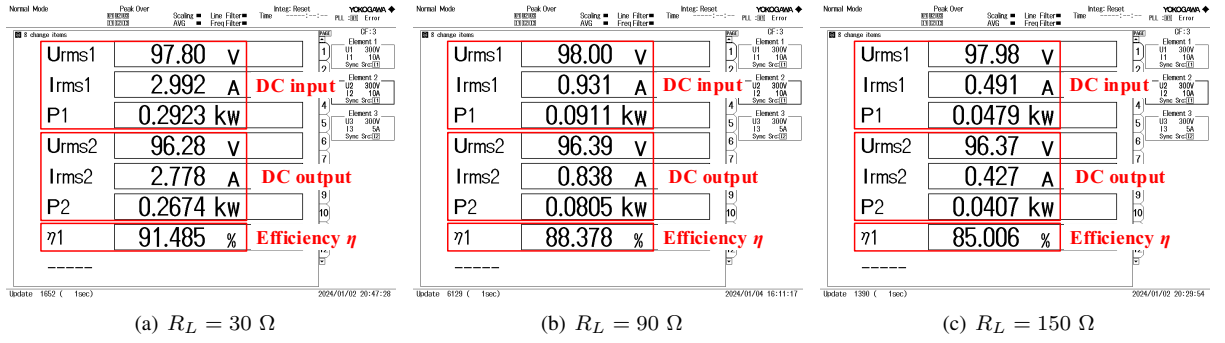


Fig. 13. Measured steady-state experimental input power, output power and efficiency under different load conditions with coupling coefficient  $k = 0.26$ .

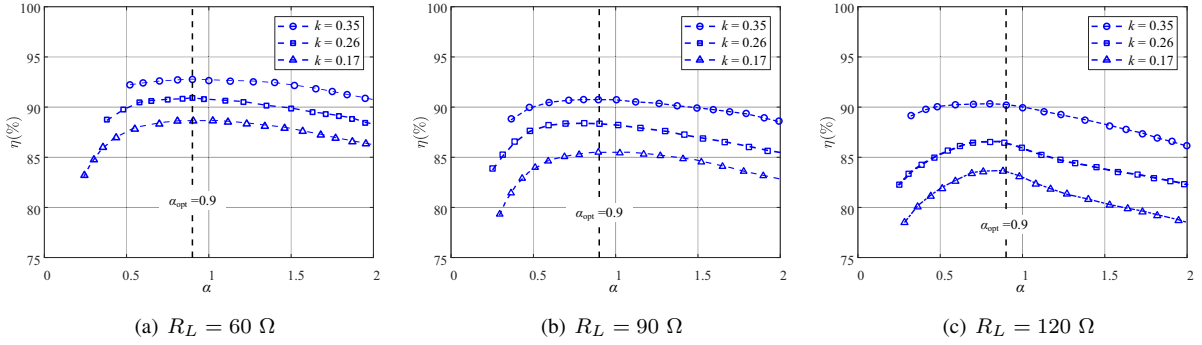


Fig. 14. Measured efficiency versus  $\alpha$  under different coupling coefficients and load conditions.

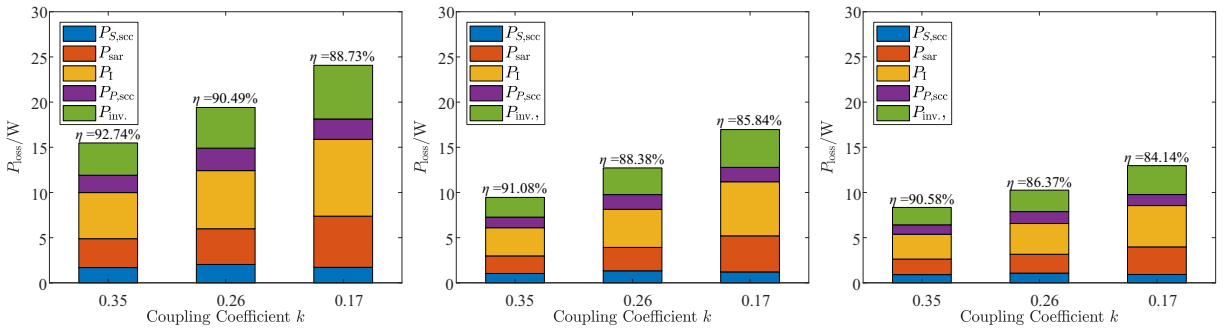


Fig. 15. Power loss breakdown of the proposed single-stage IPT.



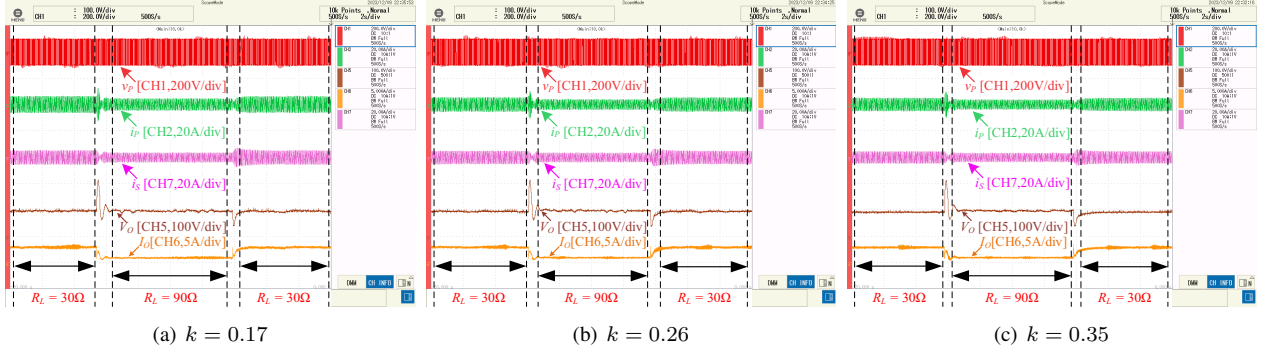


Fig. 16. Transient waveforms against step load change under different coupling coefficients.

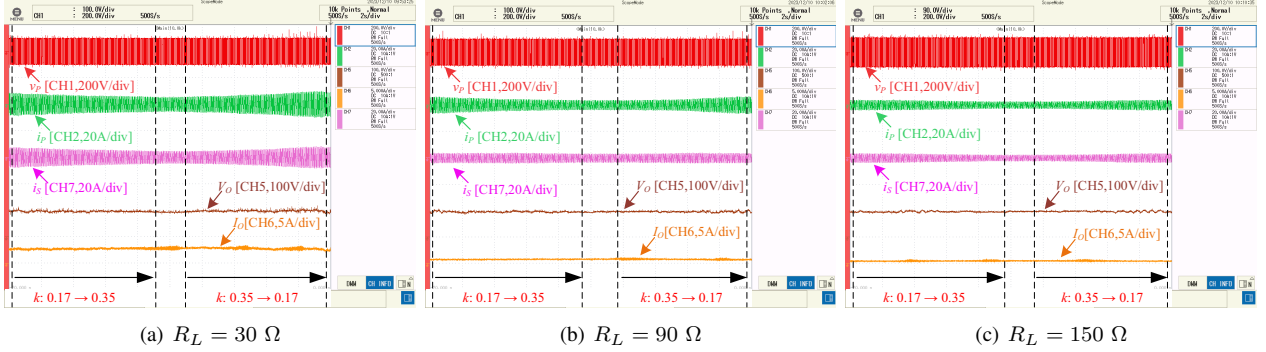


Fig. 17. Transient waveforms against variation of coupling coefficient under different load conditions.

Therefore, these three sets of experiments validate the optimal efficiency tracking and CV output against variations of  $k$  and load conditions.

## V. CONCLUSION

A single-stage series-series compensated inductive-power-transfer converter with two switch-controlled capacitors and a semi-active rectifier is proposed to cope with power regulation and optimal efficiency tracking against variations of  $k$  and load conditions. The operating principle is detailed and a two-loop control scheme is used to achieve a constant voltage output and track the optimal efficiency. Steady-state waveforms and transient responses are experimentally measured to validate the performance.

## APPENDIX

The power losses in the full-bridge inverter, the primary SCC and the primary coil should be considered to estimate  $R_P$ . Likewise,  $R_S$  represents losses in the secondary coil, the secondary SCC and the SAR. To estimate  $R_P$  and  $R_S$ , the power losses in the primary SCC, secondary SCC and the SAR should be calculated.

### A. Calculation of $R_P$

According to the operating waveforms and switching sequence shown in Fig. 7, since the primary SCC realize soft switching, only conduction loss will be considered and it can be calculated by

$$P_{P,SCC} = I_{P,SCC,RMS}^2 R_{on,2} + I_{P,SCC,Avg} V_{f,2}, \quad (26)$$

where  $R_{on,2}$  and  $V_{f,2}$  are on-resistance and body-diode forward voltage of the MOSFET switches  $Q_a$  and  $Q_b$ . Besides,  $I_{P,SCC,rms}$  and  $I_{P,SCC,avg}$  are the root mean square (RMS) and average value of the current flowing through primary SCC and they can be derived by

$$I_{P,SCC,RMS} = \sqrt{\frac{1}{\pi} \int_{\pi-\varphi_P}^{\varphi_P} (|I_P| \sin x)^2 dx}, \quad \text{and} \quad (27)$$

$$I_{P,SCC,Avg} = \frac{1}{\pi} \int_{\pi-\varphi_P}^{\varphi_P} |I_P| \sin x dx, \quad (28)$$

respectively.

Therefore, the  $R_P$  can be estimated by

$$R_P = 2R_{on,1} + \frac{P_{P,SCC}}{I_{P,RMS}^2} + R_{P,w}, \quad (29)$$

where  $R_{on,1}$  represents ON-resistance of the MOSFET switches ( $Q_1$ - $Q_4$ ) in the full-bridge inverter,  $i_{P,RMS} = |I_P|/\sqrt{2}$  and  $R_{P,w}$  represents the resistance of primary coil.

### B. Calculation of $R_S$

Similar to primary-side SCC, the conduction loss of secondary-side SCC can be given by (30) to (32).  $R_{on,3}$  and  $V_{f,3}$  are ON-resistance and body-diode forward voltage of the MOSFET switches  $Q_c$  and  $Q_d$ .  $I_{S,SCC,rms}$  and  $I_{S,SCC,avg}$  are

the root mean square (RMS) and average value of the current flowing through secondary SCC.

$$P_{S,SCC} = I_{S,SCC,RMS}^2 R_{on,3} + I_{S,SCC,Avg} V_{f,3} \quad (30)$$

$$I_{S,SCC,RMS} = \sqrt{\frac{1}{\pi} \int_{\pi-\varphi_S}^{\varphi_S} (|I_S| \sin x)^2 dx} \quad (31)$$

$$I_{S,SCC,Avg} = \frac{1}{\pi} \int_{\pi-\varphi_S}^{\varphi_S} |I_S| \sin x dx \quad (32)$$

Moreover, the SAR realizes soft switching as shown in Fig. 6, only the conduction loss of SAR is considered and calculated by

$$P_{SAR} = i_{S,RMS}^2 R_{on,4} + i_{S,Avg} V_{f,4} \quad (33)$$

where  $R_{on,4}$  is the ON-resistance of the MOSFET  $Q_5$  and  $Q_6$  in the SAR, and  $V_{f,4}$  is diode forward voltage of  $D_7$  and  $D_8$ .  $i_{S,RMS} = I_S/\sqrt{2}$  and  $i_{S,Avg} = \frac{2|I_S|}{\pi}$  are the RMS value and average value of  $i_S$  flowing into the SAR respectively.

Therefore,  $R_S$  can estimated by

$$R_S = \frac{P_{S,SCC} + P_{SAR}}{i_{S,RMS}^2} + R_{S,w} \quad (34)$$

$R_{S,w}$  is the secondary coil resistance.

### C. Calculation of Practical $\alpha_{opt}$

With the parameters listed in Table II,  $\alpha_{opt} = \sqrt{R_P/R_S}$  versus  $R_L$  under different coupling coefficients can be calculated. However, since a slight deviation from the optimal  $\alpha_{opt}$  will not affect the efficiency too much, as shown by Fig. 3,  $\alpha_{opt}$  can be set as 0.9 for simplicity.

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In 2005, 2010 and 2018, he was selected as an IEEE Distinguished Lecturer. In 2006 he chaired the IEEE CAS Technical Committee on Nonlinear Circuits and Systems. He serves and has served as Editor-in-Chief of *IEEE TRANSACTIONS ON CIRCUITS AND SYSTEMS II* (2016–2019), *IEEE Circuits and Systems Magazine* (2013–2016), *IEICE Nonlinear Theory and Applications* (since 2013); as associate editor of a few other IEEE journals; and on the Editorial Board of *IEEE Proceedings* (2021–2022). He has served on a number of IEEE committees including the IEEE Fellows Committee and the IEEE Awards Committee, and chaired the Steering Committee for IEEE Transactions on Network Science and Engineering.