A Compact Isolated 6.78-MHz Class E^2 Converter via Wireless Inductive Coupling

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***Abstract*—Compact, isolated DC-DC converters are required in many applications such as communication terminals. This paper presents a compact, planar Class E^2 DC-DC converter via wireless inductive coupling. For high frequency converters working at megahertz (MHz), it is challenging to design the system parameters under the condition of different final loads and voltage ratios. In this paper, a design method is proposed to adjust the voltage ratio of a Class E^2 DC-DC converter without changing the turn ratio of the coils. An experimental 6.78 MHz Class E^2 converter system with flexible voltage ratio is built, with the coupling coefficient *k*=0.33 and overall efficiency above 80%.**

***Index Terms*—DC-DC power converters, compact, planar, voltage ratio, Class E, isolation.**

I. Introduction

In the past few years, the industry has placed high demands on the safety and miniaturization of the power conversion system. In such systems, compact and galvanic isolation characteristics are required. Power conversion circuits with high operation frequency own the potential to be lighter and more compact [1], [2], which are also applied to wireless power transfer system [3], [4].

Typically, an isolated DC-DC converter includes an inverter (amplifier), a coupling coil, and a rectifier. Class E power amplifier (PA) is a good candidate high frequency converter due to its circuit simplicity and high efficiency [5], [6], which is designed under zero voltage switching (ZVS) and zero voltage derivative switching (ZVDS) conditions. The Class E rectifier was first proposed in 1988 [7], with various Class E topologies developed later. The Class E full-wave current-driven rectifier is a promising candidate for low-harmonic-contents and high-efficiency rectification at MHz [3]. In this paper, a Class E^2 DC-DC converter consisting of a Class E PA, coupling coils, and a Class E full-wave rectifier is used.

Compared to conventional DC-DC converter typologies such as boost or buck converters, a Class E^2 DC-DC converter with coupling coils owns the advantage of galvanic isolation, weight reduction and efficiency improvement. Besides, the using of coupling coils without an iron core provides possibility for the converter to be planar and more compact. Conventional converters is able to provide a desired DC-DC voltage ratio preciously, while for resonant converters as Class E^2 converters it is more challenging to achieve that. Most researchers tend to design converters with the required output voltage by choosing the appropriate turn ratio of the coils [1], [2], [8], which adds complexity to the design and manufacturing. The article [8] proposes an equivalence approach to transform the Class E^2 DC-DC converter to a new typology. The method is able to reduce the number of inductors and achieve the desired voltage ratio, which is practical in system design. The method has two main disadvantages, however. First, an original design with a known voltage ratio is required to perform circuit transformation. Second, when changing different voltage ratios, the inductors of both primary and secondary sides need to be replaced. Besides, the height of the converter in that paper makes it not compact enough. Actually, if a planar design of the converter is required, the coupling coefficient k and coil inductance can not be used as flexible design variables.

In this paper, a compact, planar Class E^2 DC-DC converter via wireless inductive coupling is proposed. Flexible voltage ratio is realized through changing the compensation capacitors of the primary and second sides. In the proposed design procedure, the inductors of the system remain unchanged, which dramatically reduce the complexity to design and manufacture such a system. Moreover, the performance of the Class E PA is analyzed in particular to derive an approach to get a higher DC-DC voltage ratio.

In order to achieve high compactness of the converter, the following measures have been taken: 1) The coils are loosely coupled, i.e., the coupling coils of the converter have an air core. 2) The AC inductor of the LC matching network is designed to be absorbed into the primary coil. 3) The coupling coils are designed with embedded structure to reduce the thickness.

II. System Analysis

A. System Configuration

The system configuration of the Class E^2 DC-DC converter is shown in Fig. 1 (a). It consists of a Class E PA, coupling coils and a Class E full-wave rectifier. In conventional design, there is always a LC matching network of Class E PA [9]. However in this paper, the matching network is combined with the transmitting coil L_{tx} and its compensation capacitor C_{tx}, which makes the system more compact. In the figure, P_{dc} is the input power of the PA, P_{in} is the input power of the coils, P_{rec} is the input power of the rectifier and P_o is the power dissipated on the DC load. Assuming that the secondary side of the circuit is compensated to pure resistive, i.e., the imaginary parts of the input impedance of the rectifier and receiving coil are compensated by C_{rx}. In that case, the system in Fig. 1 (a) can be transformed [10]. The input impedance Z_{in} of the coupling coils is formulated as follows, which is pure resistive.

$$Z_{in} = r_{tx} + R_{re} = r_{tx} + \frac{\omega^2 M^2}{R_{rec} + r_{rx}}. \tag{1}$$

Where M is the mutual inductance calculated by $M = k\sqrt{L_{tx}L_{rx}}$. In order to change the DC-DC voltage ratio, some new design variables should be added. Here, the capacitor C_{rx} is adjusted so as to generate reactance ΔX_{rx} on the secondary side of the converter (Fig. 1 (b)). Note that C_{tx} is adjusted to $C_{tx}^{'}$ to compensate the imaginary part of Z_{re}.

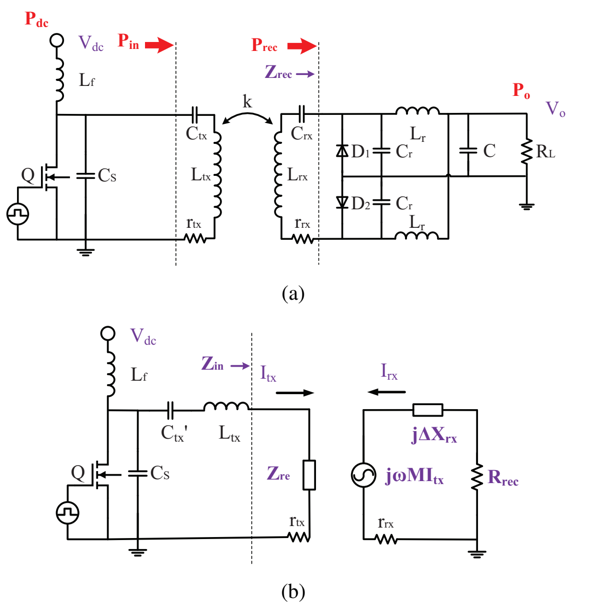


Fig. 1. System equivalence of the Class E^2 DC-DC converter. (a) Original system. (b) Equivalent system.

B. Impedance Transformation

In order to achieve high operation efficiency, optimum design needs to be performed for every part of the system. Meanwhile, the system should also be able to work with desired DC-DC voltage ratio, which will improve the complexity of system design. For such a nonlinear Class E^2 system, it will be more reliable to analyze the voltage ratio through impedance transformation. The optimum design of the rectifier is only related to the DC load R_L, where (2) can be used to give the optimum design of C_r [3].

$$C_r = \frac{0.1756}{\omega R_L}. \tag{2}$$

The input impedance Z_{rec} of the rectifier is calculated in [93]. Z_{rec} can be represented as

$$Z_{rec} = R_{rec} + jX_{rec}, \tag{3}$$

where

$$\begin{aligned} R_{rec} = &-\frac{2\sin(\phi+2\pi D)}{\pi\omega C_r}\ [\sin\phi - \sin(\phi+2\pi D) \\ &- 2\pi(1-D)\cos\phi + \frac{\cos\phi - \cos(\phi+2\pi D)}{\tan(\phi+2\pi D)} \\ &+ \frac{\cos(2\phi+4\pi D) - \cos 2\phi}{4\sin(\phi+2\pi D)}], \end{aligned} \tag{4}$$

$$\begin{aligned} X_{rec} = &-\frac{2\sin(\phi+2\pi D)}{\pi\omega C_r}[\cos\phi + 2\pi(1-D)\sin\phi \\ &- \frac{\sin\phi}{\tan(\phi+2\pi D)} - \frac{\cos(\phi+2\pi D)}{2} \\ &+ \frac{2\pi(1-D) + \sin\phi\cos\phi}{2\sin(\phi+2\pi D)}], \end{aligned} \tag{5}$$

$$\tan\phi = -\frac{\pi(1-D)\sin(2\pi D) + \sin^2(\pi D)}{\pi(1-D)\cos(2\pi D) + \sin(\pi D)\cos(\pi D)}. \tag{6}$$

In (4)-(6), ω is the operating frequency, C_r is the capacitance of the parallel capacitors and D is the duty cycle of the rectifier diodes. The Class E rectifier is usually designed to work at 50% duty cycle. Since the overlapping of the diode conduction exists when $D > 0.5$, D is chosen as 0.49 in this paper to avoid the overlapping. Substituting $D = 0.49$ in (4)-(6), it can be seen that Z_{rec} is only a function of R_L.

Suppose the initial value of C_{rx} is designed to compensate the the imaginary part of the secondary side,

$$j\omega L_{rx} + \frac{1}{j\omega C_{rx}} + jX_{rec} = 0, \tag{7}$$

thus, the initial value of C_{rx} is formulated as

$$C_{rx} = \frac{1}{\omega^2 L_{rx} + \omega X_{rec}}. \tag{8}$$

In Fig. 1 (b), the capacitance of C_{rx} is changed to generate ΔX_{rx}, therefore, the new capacitance of C_{rx} can be calculated as

$$C_{rx}^{'} = \frac{C_{rx}}{1 - \omega C_{rx}\Delta X_{rx}}, \tag{9}$$

and the input impedance of the coupling coils can be formulated as

$$\begin{aligned} Z_{in} = R_{in} + jX_{in} = Z_{re} + r_{tx} = \frac{\omega^2 M^2}{j\Delta X_{rx} + r_{rx} + R_{rec}} + r_{tx} \\ = \omega^2 M^2 \frac{R_{rec} + r_{rx}}{(R_{rec} + r_{rx})^2 + \Delta X_{rx}^2} + rtx \\ - j\left(\frac{\omega^2 M^2 \Delta X_{rx}}{(r_{rx} + R_{rec})^2 + \Delta X_{rx}^2}\right). \end{aligned} \tag{10}$$

From (10), it is shown that ΔX_{rx} can affect both R_{in} and X_{in}. In another word, by adjusting C_{rx}, R_{in} can be changed. In order to facilitate the analysis, it is assumed that $C_{tx}^{'}$ consists of two parts in series, i.e.,

$$\frac{1}{C_{tx}^{'}} = \frac{1}{C_{tx0}} + \frac{1}{C_{tx1}}, \tag{11}$$

where C_{tx0} is used to form the LC matching network of the PA while C_{tx1} is used to compensate the imaginary part in (10). Therefore, the value of C_{tx1} can be given as

$$C_{tx1} = -\frac{(r_{rx} + R_{rec})^2 + \Delta X_{rx}^2}{\omega^3 M^2 \Delta X_{rx}}. \quad (12)$$

From (9) and (12), it can be seen that after ΔX_{rx} is chosen, C'_{rx} and C_{tx1} will be determined. So far, the rectifier and the coupling coils have been analyzed and the input impedance Z_{in} is calculated. The Class E PA should be designed to realize soft switching condition (ZVS), where (13) and (14) give the optimum parameters of the PA [11].

$$C_S = \frac{0.1836}{\omega R_{in}}, \quad (13)$$

$$C_{tx0} = \frac{1}{\omega^2 L_0 - 1.1525 \omega R_{in}}. \quad (14)$$

With the optimum design of C_S and C_0, (15) can be applied to calculate the DC input resistance R_{dc} of the system [11].

$$R_{dc} = \frac{V_{dc}}{I_{dc}} = 1.7337 R_{in}. \quad (15)$$

Since all system parameters are optimum designed, it is safe to assume that the system can work with high overall efficiency. Suppose the system efficiency is constant (say $\eta_{sys} = 80\%$), the DC-DC voltage ratio can be formulated as

$$K_{sys} = \frac{V_{dc}}{V_o} = \frac{\sqrt{P_{dc} R_{dc}}}{\sqrt{P_o R_L}} = \sqrt{\frac{R_{dc}}{\eta_{sys} R_L}}. \quad (16)$$

From the equations above, the relationship between the variables can be found. Giving the load R_L, the input resistance R_{rec} of the rectifier is determined. R_{dc} is determined by R_{in} and R_{in} is determined by C'_{rx} and M. Therefore, the compensation capacitor of the receiving coil and the mutual inductance of the coils are the only two independent variables that influence the DC-DC voltage ratio. Besides, C_{tx} is adjusted to C'_{rx} to compensate X_{in} brought by ΔX_{rx}. From equations (10), (15) and (16) it is not difficult to find that by adjusting C'_{rx}, K_{sys} can only be reduced. Therefore, we should choose a mutual inductance M which is large enough to achieve desired maximum voltage ratio at very beginning.

C. Analysis of Class E PA

In the previous analysis, the reactance X_{in} is completely compensated by C_{tx1}, keeping Z_{in} seen by the PA pure resistive,which somehow reduces the number of variables affecting K_{sys}. It is known that ZVS can be achieved if the load of the PA is inductive, It can be seen from (16) that if the variation of X_{in} leads to a higher DC input resistance R_{dc} of the PA, K_{sys} will be further increased.

Advanced design system (ADS) from Agilent is used to simulate and investigate the PA performance under varying X_{in}, of which the result is given in Fig. 2. It can be seen from the figure that the PA can work with high efficiency when the input reactance X_{in} is not too high (say $X_{in} < 12\Omega$). In this case, the DC input impedance's variation is relatively large

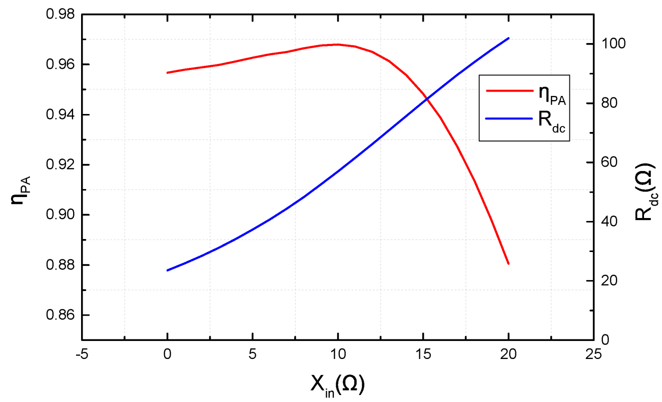


Fig. 2. The PA efficiency and the DC input impedance versus X_{in} ($R_{in} = 10\Omega$).

(about 20~60Ω). Therefore, the maximum system voltage ratio can be increased thanks to a positive X_{in}, which can be generated by increasing C'_{tx}. Actually, this fact provides us a different good solution to increase the DC-DC voltage ratio since the efficiency of a Class E PA with optimum design will decrease significantly if R_{in} is too high. Fig. 3 shows the PA voltage waveforms. It can be seen that ZVS is not realized if X_{in} is too high, which explains the efficiency drop in Fig. 2.

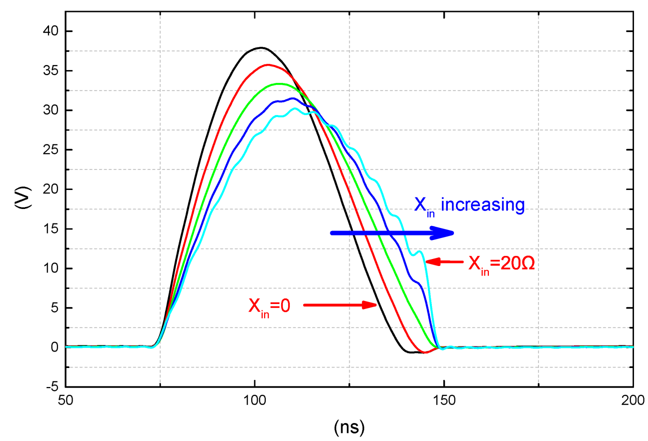


Fig. 3. PA voltage waveforms under various X_{in} ($R_{in} = 10\Omega$).

III. Parameter Design Procedure

This section describes the parameter design procedure of a Class E^2 DC-DC converter with different required voltage ratios. From (10) it can be seen that Z_{in} attains its maximum when ΔX_{rx} is 0. In that case, the voltage ratio K_{sys} reaches its maximum. Thus, in order to design a converter with the highest voltage ratio of $K_{sys,max}$, the mutual inductance must be large enough. Substituting (1) and (15) into (16) yields the minimum mutual inductance:

$$M_{\min} = \frac{1}{\omega}\sqrt{(R_{rec} + r_{rx})\left(\frac{K_{sys,\max}^2 \eta_{sys} R_L}{1.7337} - r_{tx}\right)}. \quad (17)$$

When the system is tuned to be resonant, it will provide the maximum voltage ratio $K_{sys,\max}$. ΔX_{rx} is introduced into the system to adjust the voltage ratio at the range of $K_{sys} \leq K_{sys,max}$, formulated as

$$\Delta X_{rx} = -\sqrt{\left(\frac{\omega^2 M^2}{0.5768 K_{sys}^2 \eta_{sys} R_L - r_{tx}} - R_{rec} - r_{rx}\right)(R_{rec} + r_{rx})}. \tag{18}$$

Fig. 4 illustrates the design procedure of such a Class E^2 converter with required voltage ratio, of which the details are described in the following.

Firstly, choose DC load R_L of the system, obtain the optimum design of C_r and calculate the input resistance R_{rec} and reactance X_{rec} of the rectifier. Note that X_{rec} will be compensated by C_{rx} in the next step together with the reactance of the receiving coil.

Secondly, choose the maximum DC-DC voltage ratio $K_{sys,max}$ of such a Class E^2 converter, i.e., the converter can not work with a voltage ratio higher than $K_{sys,max}$ unless increasing the inductance of the coupling coils or using the voltage ratio expansion method. Calculate M_{min} and design the coupling coils with $M \geq M_{min}$. Then, calculate the initial value of C_{rx}.

Thirdly, choose the actual voltage ratio K_{sys} and calculate the input impedance R_{in} of the coils. If R_{in} is too high compared to ωL_{tx}, then reduce K_{sys} appropriately and recalculate R_{in}. Then, update the capacitance C_{rx} with $C_{rx}^{'}$ and calculate C_{tx1}.

Finally, design C_s and C_{tx0} of the PA. C_{tx0} should be combined with C_{tx1} calculated in the previous step to get the capacitance of $C_{tx}^{'}$. If K_{sys} has been reduced in the third step, increase the voltage ratio by tuning $C_{tx}^{'}$. Fine adjust the system by taking the effects of parasitic parameters into consideration.

IV. Experimental Verification

In this section, a 6.78-MHz example converter is built up to achieve different voltage ratios. MOSFET SUD15N15 and diode DFLS260 are used to build the system. Note that the coupling coils remain unchanged in different experimental setups. Fig. 5 illustrates the experimental setup used to verify the proposed design approach.

Although the DC-DC voltage ratio is adjusted by changing the capacitances, the coupling coils should be designed with a mutual inductance higher than M_{min}, as shown in Fig. 4. Suppose $R_L = 2.5\Omega$ and $K_{sys,max} = 4$, then the minimum mutual inductance can be calculated by (17) as $M_{min} = 235nH$ (with the coil resistances ignored). The primary and secondary coils are designed with different radii so the latter can be embedded in the formal. If the radii of the coils are determined, (19) can be used to choose the turns.

$$M = \frac{\mu_0 \pi N_1 N_2 R_1^2 R_2^2}{2\left(R_1^2 + d^2\right)^{3/2}} \geq M_{\min}, \tag{19}$$

where μ_0 is the permeability of vacuum, N_1 and N_2 are the turns while R_1 and R_2 are the radii of the coils, and d is the distance between the coils, which is equal to 0 since the coils are concentric. Here $R_1 = 14mm$, $R_2 = 9mm$, $N_1 = 5.5$ and $N_2 = 4.5$ are chosen, with the mutual inductance calculated

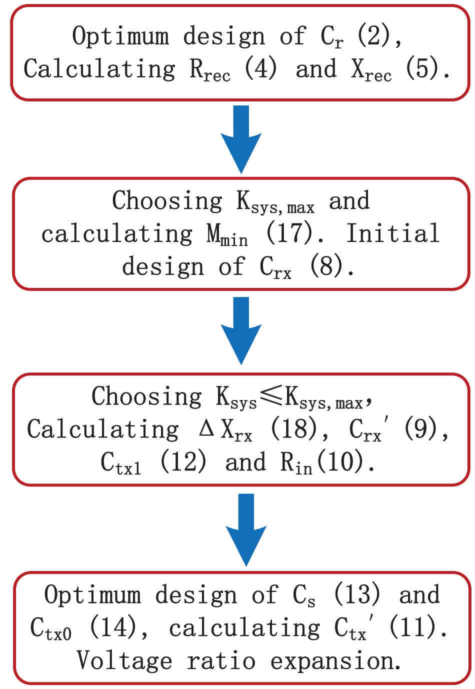


Fig. 4. Design procedure.

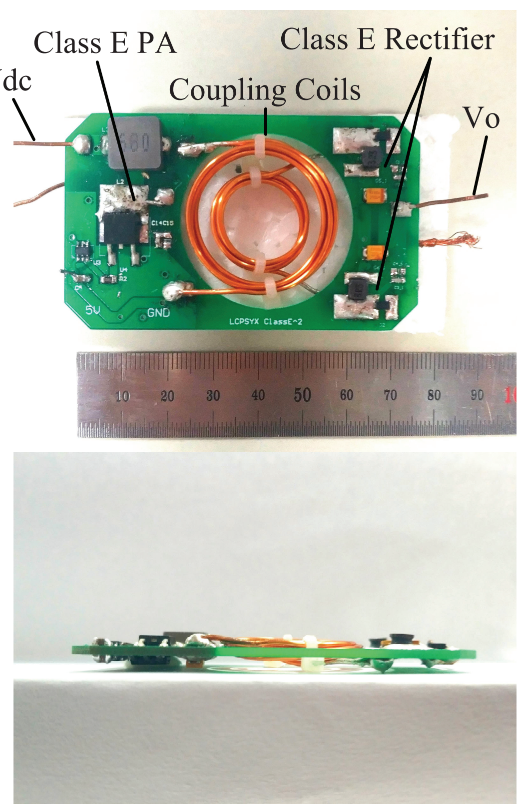


Fig. 5. The experimental 6.78-MHz DC-DC converter.

by (19) as $M = 283nH > 235nH = M_{min}$.

High frequency structure simulator (HFSS) is used to simulate the coils. Fig. 6 shows the simulated coil models, where Port1 and Port2 are the excitations implemented to the coils. The simulation results are given in Table I.

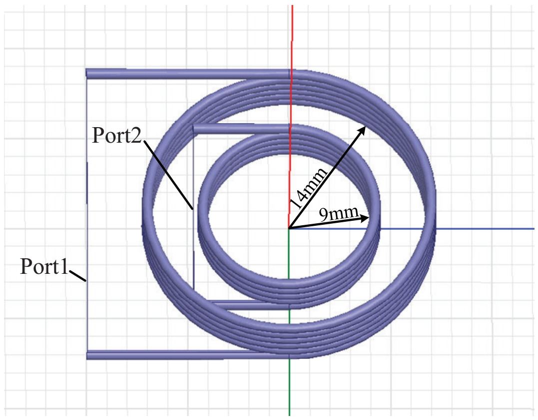


Fig. 6. Embedded coil models simulated in HFSS.

TABLE I
PARAMETERS OF COILS SIMULATED BY HFSS

L_{tx} (nH)	L_{rx} (nH)	M (nH)	$r_{tx}(\Omega)$	$r_{rx}(\Omega)$
1292	526	301	0.34	0.12

Table II gives three sets of system parameters calculated by a MATLAB script based on the design procedure. The real values of L_{tx}, L_{rx} and mutual inductance are measured by vector network analyzer (VNA). The designed voltage ratios are 4 and 2 in the first and second setups. Note that the third and forth setup are constructed with the voltage ratio expansion approach illustrated in section II-C. Note that the parasitic capacitor of SUD15N15 is assumed to be constant, 40pF, which is taken into account when calculating the value of C_s.

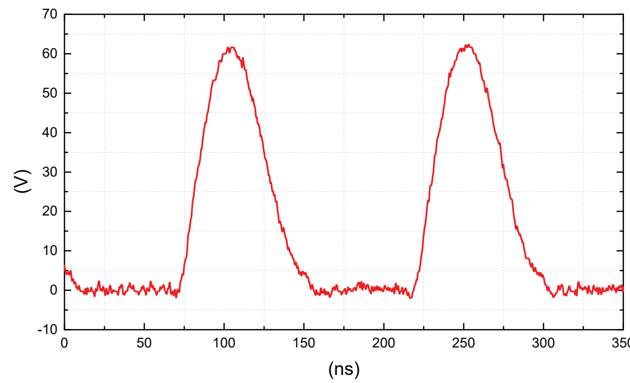
TABLE II
PARAMETERS OF EXPERIMENTAL SYSTEM ($L_{tx} = 1368nH$, $L_{rx} = 557nH, k = 0.33$)

$NO.$	R_L	K_{sys}	C_s	C'_{tx}	C'_{rx}	C_r
	(Ω)		(pF)	(pF)	(pF)	(pF)
I	2.5	4	194	476	1112	1649
II	5	2	417	450	1426	824
III IV	5	-	-	560	-	-

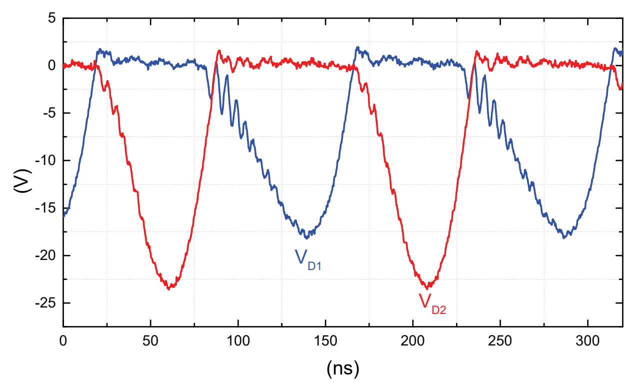
The experimental results are listed in Table III, where K_r is the measured voltage ratio and $\delta_K = |K_r - K_{sys}| / K_{sys}$ is the design error of the voltage ratio. It can be seen from the results that the measured and designed voltage ratio are quite close to each other, which shows that the performance of the real system is in good agreement with the calculation. The experimental results of setup III and setup IV also demonstrate the voltage ratio expansion approach by adjusting C'_{tx}, with the measured voltage ratio increasing from 2.05 to 3.33 and 3.18.

TABLE III
EXPERIMENTAL RESULTS

$NO.$	V_{dc}	V_o	K_r	δ_K	P_o	η_{sys}
	(V)	(V)			(W)	
I	18	4.57	3.93	1.8%	8.35	81.2%
II	18	8.76	2.05	2.5%	15.35	82.4%
III	18	5.41	3.33	-	5.88	80.2%
IV	32	10.07	3.18	-	20.34	82.7%



(a)



(b)

Fig. 7. Voltage waveforms of the converter (setup III). (a) PA MOSFET. (b) Rectifier diodes.

Fig. 7 shows the voltage waveforms of the PA and the rectifier. It can be seen from Fig. 7 (a) that ZVS and ZVDS are both satisfied, which lead to a high PA efficiency. In Fig. 7 (b), the peak voltages of two diodes in are not equal mainly due to the harmonic contents transfered from the primary side. This is because the filter inductor of the PA is eliminated and the inductance of the transmitting coil is not large enough, which results in an insufficient filtering ability. A slight difference can be found between the temperatures of the two diodes (Fig. 8), which has an adverse effect to the performance and life of the rectifier. On the one hand, it is a compromise between system compactness and efficiency. On the other, the proposed voltage ratio expansion method is preferred than increasing the mutual

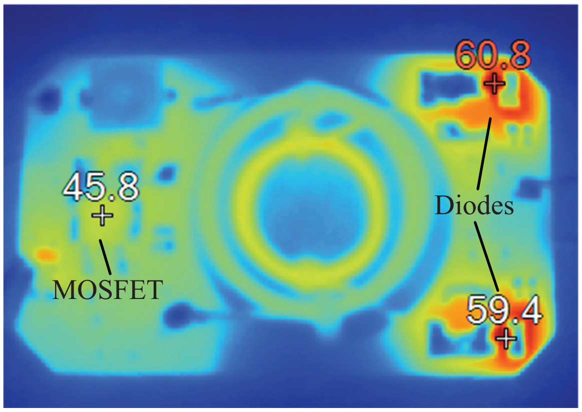


Fig. 8. Thermal image of the converter (setup III, unit: C).

inductance when requiring a higher voltage ratio.

V. Conclusions

In this paper, a compact Class E^2 DC-DC converter via wireless inductive coupling is designed. Calculation procedure is presented to design the parameters, which makes the converter work with high efficiency and different voltage ratios. Both simulations and experiments are carried out to validate the proposed converter. In addition, an effective voltage ratio increasing method is derived, which shows good performance during experimental process. Meanwhile, the analytical method in this paper can be extended to other conversion circuits.

VI. Acknowledgement

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