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Minimizing Coil Power Loss in a Direct AC/AC Converter-based Contactless Electric Vehicle Charger

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Keywords

«AC/AC converter», «Battery charger», «Contactless power supply», «Electric vehicle», «Resonant converter», «Soft switching».

Abstract

This paper explains an optimum load analysis to reduce power loss in a direct AC/AC converter-based contactless electric vehicle charger. The converter has a fewer number of bi-directional switches than a matrix converter and it uses a resonant circuit to utilize zero-current switching. Output power and power loss depend on coupling factor and output load. A proper load value leads to low loss and maximum transmission efficiency. The value was derived using an iterative-based Weierstrass' method due to converter dynamic switching frequency. Other parameters such as link efficiency, link gain and power loss ratio were derived based on steady-state analysis. Simulation results are then presented to validate the theoretical analyses.

Introduction

Inductive-based contactless power transfer (ICPT) has been developed globally to charge an electric vehicle (EV). The main purpose behind its usage is to remove direct wire connections between the charger and the vehicle which leads to reliability, mobility, and safety improvements. Deployment of the charging system can improve EV's charging experience which may lead to an increased popularity and adoption of the vehicle [1]-[3].

In a typical series-series compensated ICPT system connected to a three-phase source, a rectifier and a DC/AC converter are used in a primary side to drive primary coil current. To minimize power loss in primary and secondary coil resistances, a proper optimum load value must be selected. An additional converter at the input side is also added to reduce primary current at a partial load that cannot be achieved by either frequency or dual control method [3]-[5]. This solution adds extra cost to the charging system.

In [6] and [7], the authors proposed a direct AC/AC converter for ICPT that has a lesser number of switches than a matrix converter. It operates based on current injection and free-oscillation strategies which were previously introduced in [8]-[10]. Since the output power is controlled by a primary current amplitude, the system can reduce coil power loss without an additional input-side converter. The switching frequency is based on system resonant frequency to utilize zero-current switching (ZCS), so the value depends on a coils coupling factor as well as a connected load. Due to variable frequency, the optimum load calculation is more complicated than in a series-series compensated ICPT case given in [5]. In addition to the load analysis, link efficiency, link gain and power loss ratio equations will also be

investigated. All of them will be used to formulate current reference to control the output power at low coil power loss.

This paper is organized as follows. A brief description of the proposed direct AC/AC converter topology and its working principle are given in Section Contactless charger system. Theoretical formulations give explanation on how to calculate the optimum load using steady-state analysis approach to achieve low power loss. In Section Simulation results, PLECS software outcomes are presented to verify the theoretical calculations. Finally, conclusions of the paper are given in the last section.

Contactless charger system

Converter topology and its equivalent circuit are given in Fig. 1. In topology figure, the primary side contains four bi-directional switches (S_a, S_b, S_c and S_d). The switches will be used to drive the resonant circuit L_p, C_p and L_s which are controlled by the on-off current controller. The controller receives sensor information marked by dashed lines, which are three-phase input voltage $v_x(t)$, primary side resonant current $i_p(t)$, and connected battery voltage $v_o(t)$. It will change the amplitude of $i_p(t)$ based on battery voltage level provided by the secondary pick-up circuit. The topology can be simplified for modeling purpose by an equivalent circuit model given in Fig. 1b. Parameters R_p and R_s indicate primary and secondary coil resistances respectively. For a particular type of modulation strategy explained in [6], the input side can be approximated by a square wave $v_{in}(t)$. A load R_{eq} is a representation of the secondary circuit rectifier and battery obtained from [11], where the battery is assumed to be purely resistive (R_L).

Fig. 1: A Direct three-phase to single-phase AC/AC converter topology and its equivalent circuit.

Some possible current commutations of the primary side are given in Fig. 2. Case 1 and 2 correspond to injection mode where the primary current $i_p(t)$ flows either from or to one of the input sources which is phase “a” in this case. It can be seen that only either $\{Da2, Sa+\}$ or $\{Da1, Sa-\}$ pair which is connected to the corresponding phase is operational. The mechanism of injection modes for other input phases are also similar. Free-oscillation mode is described in Case 3 and 4, where the current oscillates in the resonant circuit. The switching transitions between injection and free-oscillation are done during zero-crossing of the primary resonant current. Both injection and free-oscillation modes will be used to control the current amplitude which leads to a controlled output power. Converter behavior using one particular modulation strategy is explained in more detail in [6] and [7]. It basically injects the resonant circuit from the input that equals “maximum absolute value of all three-phase inputs”. Every injection is mostly accompanied by free-oscillation. Analyses in this paper are based on such modulation strategy.

Theoretical formulations

To simplify theoretical calculations, all electric components are assumed to be ideal and the load is considered to be purely resistive. The converter is also assumed to be operated in an open-loop configuration. The used modulation strategy produces zero phase-shift between input square voltage and primary reso-

Fig. 2: Possible current commutations of the AC/AC converter.

nant current [6]. Circuit model in Fig. 1b is used for the calculations. All power equations refer to an average power.

Link efficiency

Primary and secondary circuits transmission/link efficiency analysis is used to make the converter operates in low coil power loss region. By looking at Fig. 1b, the link efficiency, or η_{link} is defined as a multiplication between “a ratio of a power transferred to the secondary side to a power put across the primary circuit” and “a ratio of a power dissipated in the load R_{eq} to a total power consumed by the secondary side” [5]. The parameters can be obtained using a steady-state analysis under open loop configuration given in [3] and [5]. Resonant frequency of the system is calculated beforehand. For ICPT that contains only a series capacitor on the transmitter side, the steady-state equations are,

$$\begin{cases} \mathbf{V}_s = \frac{\mathbf{i}_p}{j\omega C_p} + \mathbf{i}_p j\omega L_p + \mathbf{i}_p R_p - \mathbf{i}_s j\omega M \\ 0 = \mathbf{i}_s R_{\text{eq}} + \mathbf{i}_s j\omega L_s + \mathbf{i}_s R_s - \mathbf{i}_p j\omega M, \end{cases} \quad (1)$$

where $\mathbf{V}_s$ is the root-mean-square (RMS) of $v_{\text{in}}(t)$. Their amplitude relationship for a particular modulation method used in this paper is explained in [6], which is $V_s = V_A/\pi$. Variable V_A is an amplitude of averaged $v_{\text{in}}(t)$ during maximum absolute value of three-phase input. The combinations of both equations

in (1) by eliminating either $\mathbf{i}_s$ or $\mathbf{i}_p$ are described in (2) and (3) as,

$$\mathbf{V}_s = \mathbf{i}_p \left[R_p + \frac{\omega^2 M^2 (R_{eq} + R_s)}{\omega^2 L_s^2 + (R_{eq} + R_s)^2} + j \left(\omega L_p - \frac{1}{\omega C_p} - \frac{\omega^3 M^2 L_s}{\omega^2 L_s^2 + (R_{eq} + R_s)^2} \right) \right], \quad (2)$$

$$= \mathbf{i}_s \left[\left(\frac{R_s + R_{eq} + j\omega L_s}{j\omega M} \right) \left(R_p + \frac{1}{j\omega C_p} + j\omega L_p \right) - j\omega M \right]. \quad (3)$$

During resonant ($\omega = \omega_0$), the imaginary part in (2) equals zero, and its frequency can be found by solving the part,

$$\omega_0 L_p - \frac{1}{\omega_0 C_p} = \frac{\omega_0^3 M^2 L_s}{\omega_0^2 L_s^2 + (R_{eq} + R_s)^2}, \quad (4)$$

$$C_p (L_p L_s^2 - M^2 L_s) \omega_0^4 + [L_p C_p (R_{eq} + R_s)^2 - L_s^2] \omega_0^2 - (R_{eq} + R_s)^2 = 0. \quad (5)$$

The roots of biquadratic equation $\equiv$ can be solved by substitution and produces four resonant frequencies,

$$\omega_0 = \pm \sqrt{-A \pm \Omega}, \quad A = \frac{1}{L_p^2} \left[\frac{m^2 (R_{eq} + R_s)^2 - n}{1 - k^2} \right], \quad \Omega = \sqrt{A^2 + \frac{4nm^2 (R_{eq} + R_s)^2}{L_p^4 (1 - k^2)}}, \quad (6)$$

$$m = \frac{L_p}{L_s}, \quad n = \frac{L_p}{C_p}. \quad (7)$$

The practical ω_0 is one that has positive and real value. In a real case, $R_{eq} > 0$, $R_s > 0$ and $0 < k < 1$, therefore $\Omega > A$ is always valid. The practical ω_0 equation under such circumstances is,

$$\omega_0 = \sqrt{-A + \Omega}. \quad (8)$$

During resonant, primary and secondary current expressions from (2) and (3) become,

$$\mathbf{i}_{p(res)} = \mathbf{V}_s / \left[R_p + \frac{\omega^2 M^2 (R_{eq} + R_s)}{\omega^2 L_s^2 + (R_{eq} + R_s)^2} \right], \quad (9)$$

$$\mathbf{i}_{s(res)} = \mathbf{V}_s / \left\{ \frac{R_p L_s}{M} + \frac{\omega_0^2 L_s M (R_s + R_{eq})}{(R_s + R_{eq})^2 + \omega_0^2 L_s^2} + j \left[\frac{\omega_0^3 L_s^2 M}{(R_s + R_{eq})^2 + \omega_0^2 L_s^2} - \frac{R_p (R_s + R_{eq}) - \omega_0^2 M^2}{\omega_0 M} \right] \right\}. \quad (10)$$

The link efficiency is derived during $\omega = \omega_0$,

$$\eta_{link} = \frac{P_{real(p)}}{P_{in}} \frac{P_{out}}{P_{real(s)}} = \frac{|\mathbf{i}_{p(res)}|^2 R_{ref}}{|\mathbf{i}_{p(res)}|^2 (R_p + R_{ref})} \frac{|\mathbf{i}_{s(res)}|^2 R_{eq}}{|\mathbf{i}_{s(res)}|^2 (R_s + R_{eq})}, \quad (11)$$

$$= \frac{\left[\frac{\omega_0^2 M^2 (R_{eq} + R_s)}{\omega_0^2 L_s^2 + (R_{eq} + R_s)^2} \right]}{\left[R_p + \frac{\omega_0^2 M^2 (R_{eq} + R_s)}{\omega_0^2 L_s^2 + (R_{eq} + R_s)^2} \right]} \frac{R_{eq}}{(R_s + R_{eq})} = \frac{\omega_0^2 M^2 R_{eq}}{\omega_0^2 M^2 (R_{eq} + R_s) + R_p [(R_{eq} + R_s)^2 + \omega_0^2 L_s^2]}, \quad (12)$$

$$= \frac{1}{1 + \frac{1}{Q_s \gamma} + \frac{\gamma}{k^2 Q_p} + \frac{2}{k^2 Q_p Q_s} + \frac{1}{k^2 Q_p Q_s^2 \gamma} + \frac{1}{k^2 Q_p \gamma}}, \quad (13)$$

$$k = \frac{M}{\sqrt{L_p L_s}}, \quad Q_p = \frac{\omega_0 L_p}{R_p}, \quad Q_s = \frac{\omega_0 L_s}{R_s}, \quad \gamma = \frac{R_{eq}}{\omega_0 L_s}, \quad (14)$$

where R_{ref} is a real part of reflected secondary side in (2), $P_{real(p)}$ is a power dissipation over R_{ref} , P_{in} is an input power to the primary circuit, $P_{real(s)}$ is a power dissipation across a combined R_s and R_{eq} , and P_{out} is a consumed power on the secondary load R_{eq} . Parameters Q_p as well as Q_s are primary and secondary sides quality factor, k is a coupling factor, M is a mutual inductance, γ is a load matching factor, and ω_0 is a resonant frequency of the primary circuit current [3]. The ω_0 is not constant in this case but depends on k and R_{eq} as given previously in (8).

Parameter R_{eq} that corresponds to the maximum link efficiency is represented as R_{opt} . When $R_{eq} = R_{opt}$, the power transfer efficiency between primary and secondary circuits is at maximum. It is obtained through utilizing optimum value of γ or γ_{opt} . The γ_{opt} itself is achieved by equating $\frac{d\eta_{link}}{d\gamma}$ with zero. For an ICPT with a primary side capacitor, the expression of γ_{opt} is,

$$\gamma_{opt} = \frac{\sqrt{k^2 Q_s^2 + Q_s^2 + 1}}{Q_s}. \quad (15)$$

By assuming $Q_p = Q_s$ and $Q_p \gg 1$, the γ_{opt} becomes,

$$\gamma_{opt} \approx \frac{\sqrt{k^2 Q_s^2 + Q_s^2}}{Q_s} = \sqrt{k^2 + 1}. \quad (16)$$

A combination of equations (7), (8), (16), and the last term in (14) leads to,

$$\frac{R_{eq}}{\omega_0 L_s} = \sqrt{k^2 + 1}, \quad (17)$$

$$0 = \left[L_s \sqrt{k^2 + 1} \right] \left[\sqrt{-A + \Omega} \right] - R_{eq}, \quad (18)$$

$$\begin{aligned} \Psi(R_{eq}) &= R_{eq}^4 + R_s(k^2 + 1)R_{eq}^3 + \left[\frac{R_s^2 m^2 - n(k^2 + 2)}{2m^2} \right] (k^2 + 1)R_{eq}^2 - \frac{nR_s(k^2 + 1)^2}{m^2} R_{eq} - \frac{nR_s^2(k^2 + 1)^2}{2m^2}, \\ &= 0, \end{aligned} \quad (19)$$

where its positive and real root is the R_{opt} . Radicals method of solving quartic equation can be found in [12] and [13]. Since the quartic has many variables, the solution is too complicated to derive.

A different way is to solve all roots simultaneously using Weierstrass' method, an iterative approach that is described in [14]. The calculation involves complex number arithmetic. One disadvantage of the method is longer calculation time that can slow down closed-loop controller performance if it is performed on the fly. In practice, the initial values must be selected based on a trade-off between calculation time and precision. A precalculated lookup table that maps k to R_{opt} can be an alternative to speed up the controller. Root approximation of the quartic equation can be formulated in a following expression,

$$\hat{R}_{eq(i)} = R_{eq(i)} - \frac{\Psi(R_{eq(i)})}{\prod_{j=1, j \neq i}^4 (R_{eq(i)} - \hat{R}_{eq(j)})}, \quad i = 1, 2, 3, 4. \quad (20)$$

Subscript (i) represents a currently calculated root (four roots in total) while (j) indicates other roots except (i) . Parameters $\hat{R}_{eq}$ and R_{eq} constitutes approximative and initial roots respectively. Root approximations are obtained by sequentially calculating every root in (19) within a loop. Approximate result from previous calculation is inserted to the next one. Loop continuation is controlled using $\Delta = \hat{R}_{eq} - R_{eq}$. Through running the loop and selecting Δ to a certain precision value, all roots can be obtained simultaneously [14].

Link gain

Link gain is defined as a ratio of output voltage across secondary load R_{eq} to input voltage over primary circuit [5]. For the ICPT with a series capacitor on the primary under $\omega = \omega_0$, equation derivations are as follows,

$$G = \left| \frac{\mathbf{V}_{out}}{\mathbf{V}_s} \right| = \left| \frac{\sqrt{P_{in} \eta_{link} R_{eq}}}{\mathbf{V}_s} \right| = \frac{\omega_0 M R_{eq} \sqrt{(R_{eq} + R_s)^2 + \omega_0^2 L_s^2}}{\omega_0^2 M^2 (R_{eq} + R_s) + R_p [(R_{eq} + R_s)^2 + \omega_0^2 L_s^2]}, \quad (21)$$

$$= \frac{\eta_{link} \sqrt{(R_{eq} + R_s)^2 + \omega_0^2 L_s^2}}{\omega_0 M}, \quad (22)$$

$$= \frac{\eta_{link}}{k} \sqrt{\frac{1}{m} \left[\left(\gamma + \frac{1}{Q_s} \right)^2 + 1 \right]}. \quad (23)$$

Equation (23) is obtained using descriptions given in equations (7) and (14). The gain at γ_{opt} under assumptions that $m = 1$ and $Q_p = Q_p \gg 1$, is given as,

$$G_{opt} \approx \frac{1}{k} \sqrt{k^2 + 2}. \quad (24)$$

Power loss ratio

Power loss ratio can be defined as a ratio of total power loss in primary and secondary coil resistances to output power in the load [3]. In this ICPT case, the ratio is also calculated during $\omega = \omega_0$. By utilizing (9) and (10), the loss expressions for primary (P_p) and secondary coils (P_s) consecutively are,

$$P_p = |\mathbf{i}_{p(res)}|^2 R_p, \quad P_s = |\mathbf{i}_{s(res)}|^2 R_s. \quad (25)$$

Power loss ratio is then formulated as follows,

$$\lambda = \frac{P_p + P_s}{P_{out}}. \quad (26)$$

Analytical calculations

Contactless charger parameters to be analyzed is given in Table I. They were obtained from calculations based on limitation analysis given in [7]. Variable V_a , V_b and V_c are voltage amplitude from input phase “a”, “b” and “c” respectively. By looking at the parameters, it can also be determined that variable m and n in (7) have values of 1 and 1000 subsequently. Line frequency of all inputs is given under f . The parameters will be used to plot link efficiency, gain, output power, and power loss ratio using MATLAB software. By utilizing equations (13), (23), (11) and (26), the produced plots are given in Fig. 3 and 4.

Table I: ICPT system component values

V_a	V_b	V_c	f	C_p	L_p	L_s	R_p	R_s	k
100 V	100 V	100 V	50 Hz	0.2 μ F	200 μ H	200 μ H	0.3 Ω	0.3 Ω	0.55 – 0.83

From Fig. 3, link efficiency can theoretically be greater than 80% for a certain load range. When the load goes to a M Ω scale, the efficiency as well as output power go to zero. In the link gain case, as the load gets very large, the gain becomes constant. Power loss ratio gets very large as the load becomes much greater than 100 Ω . It can be seen that the higher the k value, the higher η_{link} becomes, but at the same time, the output power gets lower.

The value R_{opt} was acquired by applying Weierstrass' method using MATLAB. Initial value of each root and Δ were set randomly to a complex number of $1 + j$. Iteration will stop when $\Delta \leq 0.1$, which was also chosen randomly. The results in each iteration step for $k = 0.55$ and $k = 0.83$ is given in Table II. For both cases, R_{eq} is the real and positive root, which is 38.698Ω for $k = 0.55$ and 47.587Ω for $k = 0.83$. To get R_{opt} , the values must be converted to R_L through $R_L = R_{\text{eq}}\pi^2/8$. Thus for $k = 0.55$, R_{opt} is 47.742Ω , while for $k = 0.83$, it is 58.708Ω .

Fig. 3: Link efficiency and gain plots with respect to R_L using parameters given in Table I. The x-axis is limited to 100Ω for clarity purpose.

Fig. 4: Output power and power loss ratio plots with respect to R_L using parameters given in Table I. The x-axis is limited to 100Ω for clarity purpose.

For the obtained R_{opt} values, the link efficiency, link gain, output power and power loss ratio are given in Table III. The two G_{opt} were obtained from their corresponding gain plots. It can be seen that the coupling factor k affects all those parameters. In practice, since R_{opt} is affected by k , the value must be re-calculated if component values are changing due to environmental effect such as temperature or repositioning of primary and secondary coils.

Simulation results

PLECS software was used to further validate the theoretical calculations, the results for given component values on Table I are shown in Fig. 5 and 6. The input voltage and currents plots are shown only for $k = 0.55$ since the waveforms are also the same for $k = 0.83$, while their amplitudes are different. Parameter $v_a(t)$, $v_b(t)$ and $v_c(t)$ indicate the three-phase input voltage while $v_{\text{in}}(t)$ marks a voltage over the primary resonant circuit. From Fig. 5, during positive $v_a(t)$, S_{a+} and S_{d+} are operational to produce

Table II: Weierstrass' method calculation results for two k values

$k = 0.55$				
Step	$R_{eq(1)}$	$R_{eq(2)}$	$R_{eq(3)}$	$R_{eq(4)}$
0	$1 + j$	$1 + j$	$1 + j$	$1 + j$
1	$-255.412 - 32.052j$	$-1.233 + 3.116j$	$-6.383 - 0.949j$	$3.900 + 0.781j$
2	$-2.421 - 2.168j$	$47.643 - 62.020j$	$-11.188 - 16.604j$	$1.825 + 0.689j$
3	$-0.879 - 0.640j$	$30.410 + 26.814j$	$-26.297 + 5.972j$	$-0.057 + 1.378j$
4	$-0.315 - 0.758j$	$41.688 - 6.094j$	$-37.753 - 0.414j$	$0.013 + 0.557j$
5	$-0.310 - 0.356j$	$38.655 + 0.078j$	$-38.746 + 0.006j$	$-0.098 + 0.239j$
6	$-0.206 - 0.176j$	$38.698 - 0j$	$-38.750 + 0j$	$-0.158 + 0.152j$
7	$-0.171 - 0.148j$	$38.698 - 0j$	$-38.750 + 0j$	$-0.170 + 0.149j$

$k = 0.83$				
Step	$R_{eq(1)}$	$R_{eq(2)}$	$R_{eq(3)}$	$R_{eq(4)}$
0	$1 + j$	$1 + j$	$1 + j$	$1 + j$
1	$-399.385 - 50.048j$	$-1.178 + 3.119j$	$-6.308 - 0.980j$	$3.563 + 0.904j$
2	$-2.165 - 2.307j$	$67.290 - 105.718j$	$-9.041 - 16.067j$	$1.614 + 0.790j$
3	$-0.780 - 0.548j$	$22.854 + 30.376j$	$-30.277 + 20.167j$	$-0.742 + 1.005j$
4	$-0.151 - 0.807j$	$66.733 - 3.423j$	$-44.740 + 3.847j$	$-0.355 + 0.655j$
5	$-0.184 - 0.485j$	$47.042 - 0.662j$	$-47.730 - 0.048j$	$-0.211 + 0.299j$
6	$-0.184 - 0.209j$	$47.586 + 0j$	$-47.716 + 0j$	$-0.192 + 0.164j$
7	$-0.188 - 0.148j$	$47.587 + 0j$	$-47.717 + 0j$	$-0.189 + 0.145j$

Table III: Analytical results for two different coupling factors

$k = 0.55, R_{opt} = 47.742 \Omega$				$k = 0.83, R_{opt} = 58.708 \Omega$			
η_{link}	G_{opt}	P_{out}	λ	η_{link}	G_{opt}	P_{out}	λ
93.696%	2.596	321.859 W	6.728%	96.983%	1.924	143.730 W	3.112%

$v_{in}(t)$, while during negative $v_c(t)$, only Sc- and Sd- that are being controlled. The same conditions are also applied for different input phase cases during maximum absolute value condition. Through built-in Fourier analysis and mean value calculator in PLECS, some circuit characteristics were extracted and presented in Table IV. The value of V_s was calculated from fundamental amplitude of $v_{in}(t)$ within 6.667 ms time range. Parameter η is a fraction between P_{out} and P_{in} .

Table IV: Simulation results for two different coupling factors

$k = 0.55, R_{opt} = 47.742 \Omega$		$k = 0.83, R_{opt} = 58.708 \Omega$	
V_s	43.840 Vrms	V_s	43.903 Vrms
I_p	8.079 Arms	I_p	3.617 Arms
P_{in}	349.246 W	P_{in}	155.548 W
P_p	19.583 W	P_p	3.924 W
V_{out}	112.499 Vrms	V_{out}	84.619 Vrms
I_s	2.907 Arms	I_s	1.778 Arms
P_{out}	327.069 W	P_{out}	150.479 W
P_s	2.536 W	P_s	0.949 W
η	93.650%	η	96.741%
G_{opt}	2.566	G_{opt}	1.927
λ	6.763%	λ	3.238%

By comparing Table III and IV, it is apparent that both results are similar. The output power depends on

Fig. 5: Simulation results of input voltages and primary resonant current using parameters given in Table I for $k = 0.55$ under a time range of 20 ms. The right figures are the zoomed version of the left one in a time range of 4.2 – 5.2 ms.

Fig. 6: Coils power loss and output power plots within a time range of 20 ms using parameters given in Table I for $k = 0.55$ and $k = 0.83$.

k and R_L . When the value of k is low, consumed powers increase and vice versa. The k value depends on the coil distance, therefore components' rating can limit a flexible coil placement. Power loss ratio is also not constant but depends on k . So the device needs a good cooling system to operate at a low coupling factor. Another point to mention is, the converter creates a low V_s for equal input phases' amplitude of 100 V. This behavior significantly reduces the transferred power to the load R_L .

Conclusion

A load selection to minimize coil power loss in a variable frequency direct AC/AC resonant converter has been presented in this paper. The simulation results confirm the validity of analytical calculations. The proper load selection as well as system gain will be used to formulate current reference to control primary current amplitude in a closed loop configuration, which leads to a controlled output power. Since the load calculation is based on an iterative method, a precalculated lookup table that maps k to R_{opt} can be used in the controller. The modulation strategy should also be improved since it produces a low output power. A prototype of the converter is currently being manufactured. The output power will be increased gradually until it reaches 1.7 kW by increasing the input three-phase amplitude. Practical analyses will be reported in future publications.

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