

Article

Performance Enhancement in LC Series Resonant Inverters with Current-Controlled Variable-Transformer and Phase Shift for Induction Heating

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Abstract: This article presents an analysis of a converter based on an LC resonant inverter for induction heating applications. It employs a current-controlled variable transformer (VT) in conjunction with phase shift regulation (PS) to operate at a fixed frequency close to the resonance frequency. The converter maintains a small switching angle, enabling substantial load variations without sacrificing zero voltage switching (ZVS) for the transistors. This innovative method enhances the inverter's performance across the entire operating range. Additionally, a new design of the transformer structure with a variable ratio will be analyzed, enabling mathematical modeling. The obtained results demonstrate a performance exceeding 99%. Both the inverter and variable transformer designs were experimentally validated using a 15 kW, 200 kHz converter for induction heating applications with silicon carbide (SiC) MOSFETs.

Keywords: LC resonant inverters; induction heating; current-controlled variable-transformer; PS power control; silicon carbide (SiC) power MOSFETs



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1. Introduction

Induction heating systems use the principle of electromagnetic induction and Joule's law to cause the workpiece to heat up [1]. They are widely used in both domestic and industrial applications, where they are employed in metal melting, heat treatment, welding, and others. Each application involves different workpieces with different heating requirements, where power and frequency stand as the primary factors influencing the heating process [2,3].

In induction equipment, power regulation can be achieved using either current sources or voltage sources based on the power supply structure. Current sources employ parallel resonant oscillators [4], whereas voltage sources are utilized with series or LLC resonant oscillators [5,6]. Currently, the series resonant inverter has become more prevalent than other topologies. One of the reasons is that it utilizes a compact transformer located at the high-frequency inverter output, providing galvanic isolation and enabling impedance adaptation of the resonant circuit through the transformer ratio [7].

Regarding power modulation, pulse frequency modulation (PFM) is commonly employed, offering a wide operating range at the expense of reduced efficiency due to increased frequency and complex filtering caused by a wide noise spectrum [8]. Furthermore, optimization in magnetic components is lost when the operating frequency is increased from resonance. Other techniques such as pulse density modulation (PDM) or phase-shift modulation (PS) provide improvements in efficiency and filtering as they operate at a fixed frequency close to the resonance frequency [9]. However, they lack a wide operating range, causing the loss of ZVS. More sophisticated control techniques, referred to as narrow-frequency-range methods, combine two of these techniques, such as phase shift modulation with frequency modulation (FM&PS). This combination achieves improved efficiency and

performance in response to load changes. However, when a large operating range is required, increasing the frequency is necessary, which degrades optimization [10,11]. The latest publications on maintaining a fixed frequency only focus on LLC-type converters, where in [12], a complex state-plane analysis is used to maintain soft switching, and in [13], a current-controlled variable-inductor is used to enforce the operating range and maintain the ZVS. However, these types of converters have the disadvantage of having to take into account both resonance frequencies when designing the converters, making them complex to control, and they are not suitable for all applications.

To enhance performance, a novel control technique was proposed, combining phase-shift modulation with current control of the transformer ratio (VT&PS). As investigated in this article, the use of a series resonant inverter with an electronically controllable variable transformer ratio allows for the circuit impedance to be adapted dynamically. This enables power control at a fixed frequency through PS modulation while maintaining a fixed low switching angle over a wide operating range, reducing the inverter output current.

This article is divided into the following sections: Section 2 explains the behavior of the series resonant circuit and the influence of the transformer as an impedance adapter. Section 3 analyzes a new model of a controllable variable transformer that can be represented with equations. Section 4 proposes a design procedure for an LC resonant inverter for induction heating with VT&PS, where a mathematical analysis of losses is conducted, and the control principle is discussed. Finally, this will be compared with fixed and variable frequency PS. In Section 5, the design calculations are experimentally verified using a 15 kW induction prototype, comparing it with the two traditional PS control methods.

2. LC Series Resonant Converter Topology

The LC series resonant circuit, as depicted in Figure 1, is composed of a heating inductor (L_s) in series with a capacitor (C_s) to compensate the reactive energy. The equivalent series resistance (R_s) represents the impedance imposed by the piece to be heated. This circuit is connected to the high-frequency inverter through a transformer, providing galvanic isolation and adapting the circuit impedance with the transformer ratio.

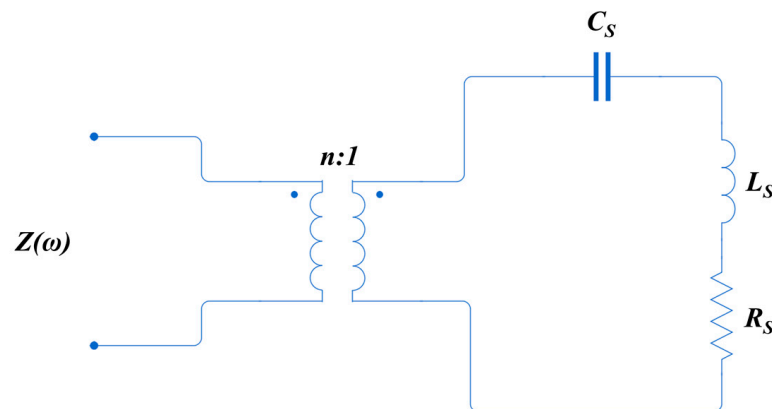


Figure 1. LC series resonant circuit.

The advantages offered by this topology over the parallel LC circuit are the elimination of the input inductance to the inverter, the cost-effectiveness and smaller size of the isolation transformer since it is located at the inverter output rather than in the power supply network, and less complex power regulation [6,9]. The series resonant circuit also presents advantages over LLC, as the design is simpler, using only two reactive elements and simplifying control by eliminating the need to manage two resonance frequencies [6].

The theoretical effect of the transformer as an impedance adapter is expressed as

$$R_p = R_s n^2, \quad (1)$$

where the secondary resistance of the transformer is reflected in the primary (R_p) multiplied by the square of the transformer turns ratio (n). This effect also occurs with the equivalent inductance reflected from the primary (L_p),

$$L_p = L_s n^2, \quad (2)$$

and has the opposite effect on the equivalent capacitance reflected from the primary (C_p)

$$C_p = \frac{C_s}{n^2}. \quad (3)$$

This ensures that changing the transformer turn ratio does not alter the resonance frequency of the circuit,

$$\omega_o = \omega_p = \frac{1}{\sqrt{L_p C_p}} = \frac{1}{\sqrt{L_s n^2 \frac{C_s}{n^2}}} = \frac{1}{\sqrt{L_s C_s}}. \quad (4)$$

and that the quality factor remains unchanged as well

$$Q = Q_p = \frac{L_p \omega_p}{R_p} = \frac{L_s n^2 \omega_o}{R_s n^2} = \frac{L_s \omega_o}{R_s}. \quad (5)$$

However, it has implications regarding the magnitude of impedance, which from the primary side of the transformer, can be expressed as

$$|Z(\omega)| = n^2 \left| R_s + j\omega L_s + \frac{1}{j\omega C_s} \right| = \frac{n^2 R_s}{\cos \alpha}. \quad (6)$$

where α is the argument of the impedance, which represents the phase between voltage and current at resonance, which is given by

$$\alpha = \arg(Z(\omega)) = \arctan \left(\frac{\omega L_p - \frac{1}{\omega C_p}}{n^2 R_p} \right). \quad (7)$$

Therefore, the effect of varying the transformer turns ratio can be observed in Figure 2. The magnitude and phase of the input impedance are plotted against frequency for different turns ratios in an LC series resonant circuit with a Q factor greater than 6, which is typical for induction heating applications [6]. As observed in the figure, for a given frequency, the magnitude of the impedance increases as the transformer ratio increases, exhibiting an opposite trend to the phase. This effect allows for more power to be obtained at the same frequency by increasing the transformer turns ratio. Therefore, this effect can be consistently achieved with the use of a current-controlled variable-transformer. In combination with PS, which varies the effective voltage applied to the resonant circuit, both the power applied to the load and the switching angle can be controlled while maintaining a fixed frequency. This differs from FM&PS control, which achieves a similar behavior in terms of power variation with a fixed switching angle, with the difference being that it needs to modulate the frequency to maintain the phase [10].

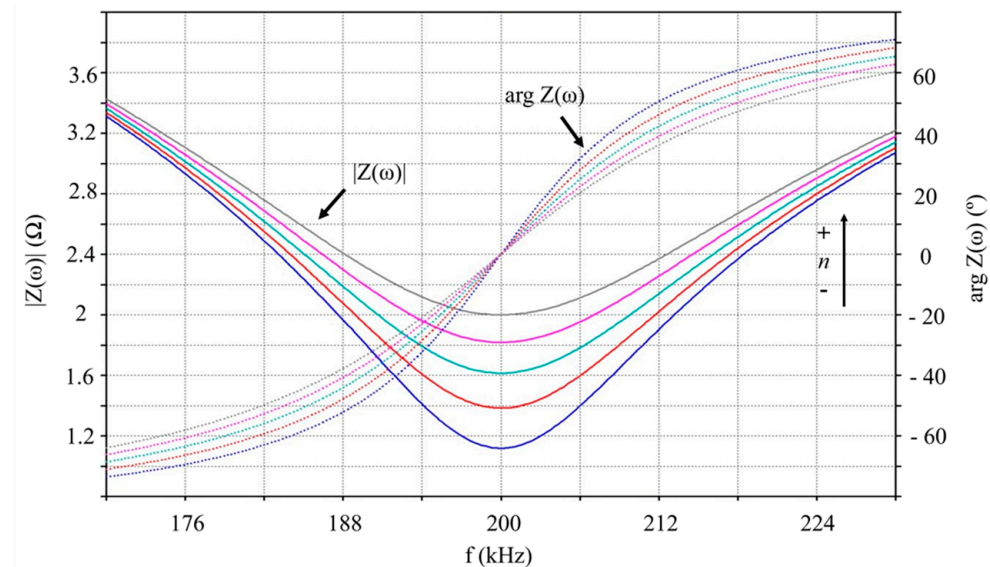


Figure 2. Magnitude of impedance (blue, red, light blue, magenta, and grey continuous curves) and phase of impedance (blue, red, light blue, magenta, and grey dotted curves) are arranged in increasing order of transformer ratio of LC resonant circuit.

3. Analysis of Current-Controlled Variable-Transformer

The analysis of the current-controlled variable-transformer has its origins in previous articles [14,15], and its basic operation is based on the model of the variable inductor already presented in other works [16,17]. Its operational principle is based on distributing the magnetic flux generated by the primary to the secondary by varying the effective permeability of the core using a path that decreases its reluctance when connected to a DC current source [18], thus increasing the influence of the reluctance of the secondary path, which includes an air gap. To achieve this, numerous articles have proposed different magnetic core topologies for various applications. Custom-made non-commercial cores have been constructed, which increase the cost and complicate the design process [19,20]. This issue has been partially addressed by using two “E”-shaped cores [21,22], but even though these are commercial cores, they still require modification due to the position of the air gap. Additionally, they present a complex structure that necessitates an experimental measurement process for transformer design with their relationships.

The structure presented in this article, as shown in Figure 3, addresses these drawbacks by utilizing two commercial “E”-shaped cores with a central air gap, which, symmetrically distributed as described below, allow the inductance model to be easily analyzed. The primary winding is wound between the two outer legs of both cores, which are joined together. The secondary winding is divided between the central legs of the cores, and their windings are connected in parallel. The two lateral arms form the saturation control inductor subjected to DC bias, which is wound in series with opposite polarity to decouple the primary and secondary from DC bias, neutralizing the magnetic flux influence and preventing induced voltages. Therefore, when the core is not subjected to DC bias, the reluctance of the secondary will be lower than that of the core, as it has an air gap, causing the magnetic flux to be closed to a lesser extent by the secondary. However, when the core is subjected to DC bias, the effective permeability of the core decreases by saturating it and thus its reluctance, resulting in the reluctance of the secondary being greater in relation to that of the core, and therefore, magnetic flux is closed through this path.

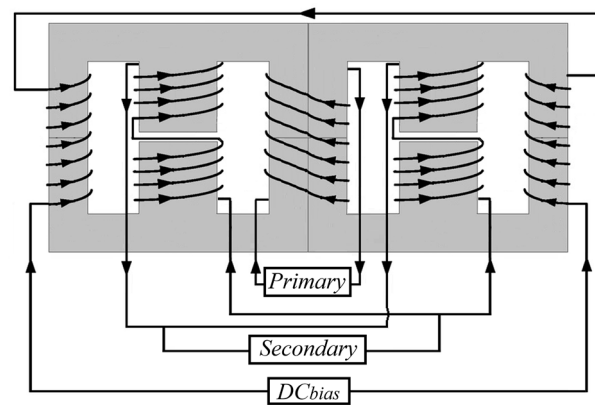


Figure 3. The structure of the current-controlled variable-transformer.

The analysis of the magnetic circuit is based on the relationships that balance the ampere turns and magnetic fluxes of the proposed structure given its symmetry. This allows for the transformer to be analyzed by defining the inductances that compose it [18,23,24]. To maintain symmetry and facilitate design, the inductances reflected from the primary will be analyzed as a single core, and those from the secondary as two in parallel. This leads to the representation of the equivalent inductance circuits shown in Figure 4.

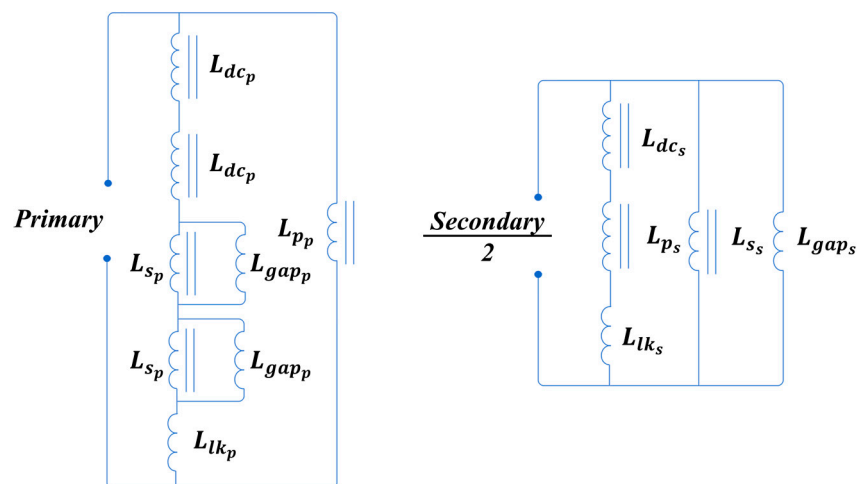


Figure 4. The equivalent inductance circuit of the primary on the left and secondary on the right.

Therefore, the inductance reflected from the primary due to the primary branch and the inductance of the primary branch reflected from the secondary are expressed, respectively, as follows:

$$L_{p_p} = \frac{\mu_0 \mu_i A_p (n_p)^2}{l_{p_p}} \text{ and } L_{p_s} = \frac{\mu_0 \mu_i A_p (n_s)^2}{l_{p_s}}, \quad (8)$$

where μ_0 is the permeability of air, μ_i is the initial permeability of the core, A_p is the effective area of the primary branch, n_p and n_s are the number of turns of the primary and secondary, l_{p_p} is the magnetic path length of the primary arm, and l_{p_s} is the magnetic path length of the primary arm to the secondary.

Regarding the inductances of the secondary branches reflected from the primary and secondary, they are, respectively, expressed as

$$L_{s_p} = \frac{\mu_0 \mu_i A_s (n_p)^2}{l_{s_p}} \text{ and } L_{s_s} = \frac{\mu_0 \mu_i A_s (n_s)^2}{l_{s_s}}, \quad (9)$$

where A_s is the effective area of the secondary branch, l_{ss} is the magnetic path length of the secondary arm, and l_{sp} is the magnetic path length of the secondary arm to the primary.

The leakage inductance, which closes through air rather than through the ferrite core, is expressed for both the primary and secondary as

$$L_{lk_p} = \frac{\mu_0 16 A_p (n_p)^2}{l_{p_p}} \text{ and } L_{lk_s} = \frac{\mu_0 16 A_s (n_s)^2}{l_{s_s}}. \quad (10)$$

The inductance due to the saturation branches, both for the primary and secondary, is also expressed as

$$L_{dc_p} = \frac{\mu_0 \tilde{\mu} A_{dc} (n_p)^2}{l_{dc_p}} \text{ and } L_{dc_s} = \frac{\mu_0 \tilde{\mu} A_{dc} (n_s)^2}{l_{dc_s}}, \quad (11)$$

where the relative permeability subjected to DC biases is denoted as $\tilde{\mu}$, A_{dc} is the effective area of the DC bias branch, and l_{dc_p} is the magnetic path length of the same branch.

Finally, the inductances due to the air gap located in the secondary branches are expressed, when viewed from the primary and secondary, as

$$L_{gap_p} = \frac{\mu_0 A_{gap} (n_p)^2}{l_{gap}} \text{ and } L_{dc_s} = \frac{\mu_0 \tilde{\mu} A_{dc} (n_s)^2}{l_{dc_s}}, \quad (12)$$

where A_{gap} and l_{gap} correspond, respectively, to the effective area and magnetic path length of the air gap.

Therefore, as shown in the equivalent circuit of Figure 4, the maximum primary inductance can be expressed as

$$\frac{1}{L_{p_{max}}} = \left(\frac{1}{L_{p_p}} + \frac{1}{L_{lk_p} + 2L_{dc_p} + \frac{2}{L_{sp} + L_{gap_p}}} \right), \quad (13)$$

while the maximum secondary inductance is given by

$$\frac{1}{2L_{s_{max}}} = \left(\frac{1}{L_{p_s} + L_{dc_s} + L_{lk_s}} + \frac{1}{L_{s_s}} + \frac{1}{L_{gap_s}} \right). \quad (14)$$

Since $\tilde{\mu}$ decreases when exposed to a continuous magnetic field intensity (H_{bias}), to which the L_{dc} winding is subjected, the ratio between L_{max} and L_{min} , with the latter being the minimum value of inductance due to the minimum $\tilde{\mu}$ resulting from maximum exposure to H_{bias} , can be obtained using the following expression for the primary as

$$\frac{L_{p_{max}}}{L_{p_{min}}} = \left(\frac{\frac{1}{2\tilde{\mu}_i} + \frac{6l_{gap}}{2l_{gap}(74+6\mu_i)+3l_{ps}}}{\frac{1}{2\tilde{\mu}_i} + \frac{24(l_{gap}+l_{ps})}{2\mu_e l_{gap} l_{ps} (74+4\mu_i+\tilde{\mu})+1}} \right), \quad (15)$$

and for the secondary as

$$\frac{L_{s_{max}}}{L_{s_{min}}} = \left(\frac{\frac{2\mu_e l_{ss}}{\tilde{\mu}+32} + \frac{\mu_e l_{ss}}{\mu_i} + 1}{\frac{2\mu_e l_{ss}}{\mu_i+32} + \frac{\mu_e l_{ss}}{\mu_i} + 1} \right), \quad (16)$$

where μ_e is the effective relative permeability of the core with a gap from the elementary equation of an inductor,

$$L = \frac{\mu_0 \mu_e A_e n^2}{l}. \quad (17)$$

4. VT&PS Resonant Inverter Design

4.1. Design Procedure

In induction heating applications, certain parameters are typically considered, such as the power range, frequency, quality factor, and inductance of the inductor obtained from its geometry. These parameters depend on the thermal treatment, the geometry of the workpiece, and the material [2]. Starting from these parameters with (4), the value of the capacitor is obtained for the frequency at which it is desired to operate [9]. Using the following expression, the switching frequency is adjusted to achieve a small angle (α) between the inverter output voltage and output current.

$$\omega_{sw} = \omega_0 \frac{\tan \alpha + \sqrt{\tan^2 \alpha + 4Q^2}}{2Q}. \quad (18)$$

The RMS value of the output current is obtained using

$$I_o = \frac{P_o}{\frac{2}{\pi} \sqrt{2} V_d \cos \alpha}, \quad (19)$$

where V_d is the DC input voltage. And the equivalent primary resistance of the transformer is obtained using

$$R_p = \frac{P_o}{(I_o)^2}. \quad (20)$$

The minimum transformer ratio (TR_{min}) is obtained from (19), (20), (5), and (1) for the maximum power at the switching frequency. On the other hand, the maximum transformer ratio (TR_{max}) is defined by the minimum power and is obtained by recalculating (19), (20), and (1) for this value. In Table 1, the starting values and those obtained from applying the design for the LC series resonant circuit are recorded.

Table 1. The initial requirements and results of the design.

Magnitude	Symb.	Eq.	Value		Unit
Output power range	P_o		15	4	kW
Quality factor	Q		10		
Frequency	f_o		200		kHz
DC input voltage	V_d		500		V
Inductor	L_s		1		μ H
Capacitor	C_s	(4)	0.63		μ F
Switching angle	α		15		°
Switching frequency	f_{sw}	(18)	203		kHz
RMS output current	I_o	(19)	32	8.5	A
Primary resistance	R_p	(20)	14.7	55.1	Ω
Transformer ratio	TR	(1)	10.7	20	

For magnetic core design and selection, the first step is to apply the effective area ratio for the maximum flux density (B_{max}) to the area of the winding, resulting in the following expression of the area product:

$$A_p = A_e A_w = \frac{V_d I_o}{2 F_{sw} B_{max} J K'} \quad (21)$$

where

A_e is the specific area of the core (m²).

A_w is the core winding area (m²).

J represents the current density (A/m²).

K is the fill factor, which will be assumed to have a value between 0.5 and 0.9.

With the obtained value, a core with dimensions larger than those calculated is selected, and the loss curves in the core at the operating frequency are compared with the value obtained from $B = \mu H$ using the maximum variation of permeability obtained from the curves of relative permeability variation under DC bias, ensuring that the obtained value is less restrictive than that used for the calculation of the area product. Then, the minimum number of turns that the primary of the transformer should have is calculated using the following expression:

$$n_p = \frac{V_d}{4 F_{sw} B_{max} A_e} \quad (22)$$

With the number of turns of the primary being obtained through expressions (13) and (15) the value of the minimum inductance of the transformer's primary is obtained, i.e., when it is subjected to maximum DC bias [25]. This corresponds to the value of the minimum primary inductance in an open circuit $L_{p_{minoc}}$, and it is assumed that 20% of this value will be the minimum primary inductance in short circuit $L_{p_{mincc}}$, resulting in the primary coupling coefficient (K_p) having a typical value of 0.8,

$$K_p = 1 - \frac{L_{p_{mincc}}}{L_{p_{minoc}}} \quad (23)$$

Therefore, the mutual inductance (L_M) is obtained from

$$L_M = \frac{L_{p_{minoc}} K_p}{1 - K_p} \quad (24)$$

And the maximum secondary inductance reflected from the primary of the transformer is expressed as

$$L_{sec_{max}} = \frac{(L_M / K_p)^2}{2 L_{p_{min}}} \quad (25)$$

Regarding the secondary inductance when the primary is short-circuited, it is obtained for the value of the desired maximum transformation ratio:

$$L_{s_{mincc}} = \frac{L_{sec_{max}}}{T R_{max}^2} \quad (26)$$

To obtain the value of the minimum secondary inductance, which corresponds to the open-circuit test value, it is again considered that the secondary coupling factor (K_s) will be the typical value of 0.8, allowing it to be obtained with the following expression:

$$L_{s_{minoc}} = \frac{L_{s_{mincc}}}{1 - K_s} \quad (27)$$

With the value of $L_{s_{minoc}}$, (14) and (16) are used to determine the number of turns that will yield that inductance value.

To obtain the value of the minimum transformation ratio, it is considered that the coupling coefficients of the primary and secondary are minimum. This is because when the core is not exposed to DC bias, the flux closes to a greater extent through the outer branches

of the core. Therefore, the communication coupling coefficient (K) approaches 0.05. These approximations will be experimentally verified later to validate the design.

$$K = \sqrt{K_p K_s}. \quad (28)$$

Therefore, the minimum transformation ratio is obtained using the maximum primary and secondary inductances obtained from (13), (15) and (14), and (16), respectively, as follows:

$$TR_{min} = \frac{1}{K} \sqrt{L_{s_{max}} / L_{p_{max}}}. \quad (29)$$

If the minimum obtained ratio is less than or equal to the desired one, the design can be considered acceptable. The final step involves using the value of $\tilde{\mu}$ employed in Equations (15) and (16) to determine the necessary H_{bias} to reach the minimum inductance values. For this purpose, the curve on the datasheet regarding the relative permeability subjected to DC bias is consulted, optimizing the balance between the bias current (I_{bias}) and the number of turns of the saturation winding (n_{dc}) given by the following expression:

$$I_{bias} 2n_{dc} = H_{bias} 2l_{dc}, \quad (30)$$

where

H_{bias} is the continuous magnetic field intensity.

l_{dc} is the length of the dc control branch from the center to the center of the core (m).

In combination with the transformer ratio control, phase shift control (PS) will be utilized. The phase shift angle factor (φ), ranging from 0 to 180°, varies the output voltage of the inverter applied to the resonant circuit using the following expression:

$$V_{ab}(\varphi) = \frac{2\sqrt{2}}{\pi} V_d \cos \frac{\varphi}{2}. \quad (31)$$

Therefore, the output current of the inverter as a function of the phase shift angle and the transformer ratio is given by

$$I_o(\varphi, TR) = \frac{2\sqrt{2}}{\pi R_s TR^2} V_d \cos \frac{\varphi}{2} \cos\left(\alpha + \frac{\varphi}{2}\right). \quad (32)$$

And the output power as a function of these two variables is expressed as

$$P_o(\varphi, TR) = \left(\frac{2\sqrt{2}V_d}{\pi}\right)^2 \frac{1}{R_s TR^2} \cos^2 \frac{\varphi}{2} \cos^2\left(\alpha + \frac{\varphi}{2}\right) \quad (33)$$

In Figure 5, the entire operating range allowed in the output power of Equation (33) with the combination of the transformer ratio variation and the phase shift is depicted for the design carried out. As observed in the figure, a dashed red line corresponding to the point of maximum efficiency operation for the entire output power range was plotted. This was obtained by optimizing Equation (32) while maintaining a switching angle (α) of 15 degrees.

In Figure 6, the simplified diagram of the VT&PS control scheme is depicted. The control system inputs are the power and the angle between the voltage and output current, and due to the two control variables, a Multiple Inputs Multiple Outputs (MIMO) system is used, which is nonlinear and coupled, acting with Proportional Integral Differential (PID) control and a decoupled diagonalization network [26,27]. The outputs of this system will therefore be the bias current and the phase angle. To optimize the system and operate at the point of maximum efficiency, represented in Figure 5 by the red dashed line, Equation (32) is used to reduce the effective value of the inverter output current with a digital observer

with a dynamic optimizer [28]. Finally, in logical signal conditioning and in the triggering block, the transistor gate signals with dead times are generated, ensuring that the system operates in ZVS. In the event of a load jump due to rapid load extraction or a change in the permeability of the workpiece causing a change in the inductor value, the system acts quickly by changing the transformer ratio, achieving coupling or decoupling between the load and the inverter. This principle also prevents the transformer from saturating due to primary voltage, despite having designed the transformer to avoid operating near saturation by the primary. This could happen if there is an overvoltage of V_d .

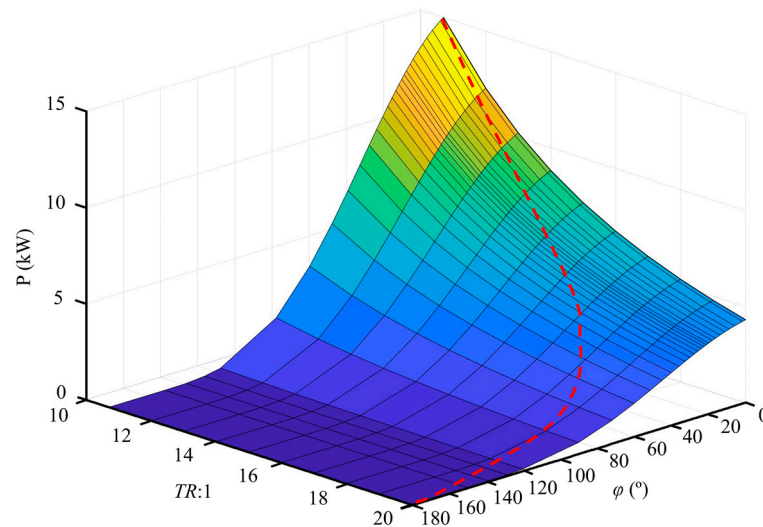


Figure 5. The phase shift and variable transformer as functions of the calculated output power. The red dotted curve corresponds to the optimal control strategy.

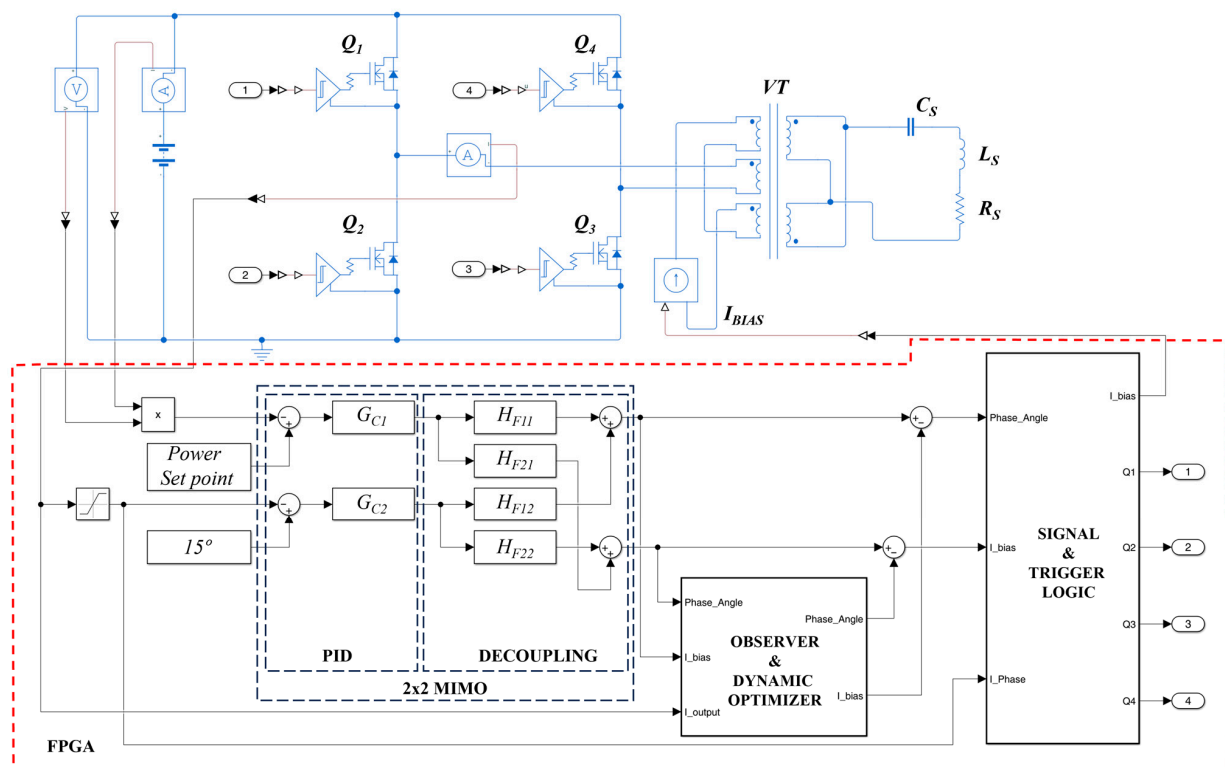


Figure 6. A schematic of the proposed system, detailing the inverter topology, the resonant load LLC with the variable inductor, and the control diagram.

4.2. Losses Analysis

The losses of the inverter and the transformer are mathematically analyzed to verify the improvements introduced by this design over the entire operating range. For this purpose, the optimal phase shift [5,29] switching sequence combined with the variable transformer is utilized.

As for the losses due to the current conducted through the channel of the MOSFETs used in the inverter, they are given by the following expression:

$$P_{cd}(\varphi, TR) = \left(\frac{I_o(\varphi, TR)}{\sqrt{2}} \right)^2 R_{DS_{on}} \quad (34)$$

where $R_{DS_{on}}$ is the ON state MOSFET resistance channel.

For the switching losses, only the OFF losses of the MOSFETs have been considered, as the inverter operates in ZVS throughout the operating range, making the ON switching losses negligible. The energy is obtained from the polynomial equation of the manufacturer's OFF state switching loss curves for the drain voltage equal to V_d and the gate resistance used, resulting in

$$E_{off}(\varphi, TR) = aI_c(\varphi, TR)^2 + bI_c(\varphi, TR) + c \quad (35)$$

where I_c corresponds to the switching current of each transistor and is expressed by the following equation:

$$I_c(\varphi, TR) = \frac{2\sqrt{2}}{\pi R_s TR^2} V_d \cos \frac{\varphi}{2} \cos \left(\alpha + \frac{\varphi}{2} \right) \sin(\alpha). \quad (36)$$

Therefore, the power losses for each transistor are obtained from

$$P_{sw}(\varphi, TR) = E_{off}(\varphi, TR) \frac{\omega_{sw}}{2\pi}. \quad (37)$$

The losses due to the transistor gate depend on the total gate charge Q_G and the gate-to-source voltage V_G , related through the following expression:

$$P_{gate} = Q_G V_G \frac{\omega_{sw}}{2\pi}. \quad (38)$$

Regarding the losses of the transformer, we must consider the losses in the core due to the magnetic flux density, with the influence of DC bias being negligible [18,30]. Therefore, they can be obtained using the following equation:

$$P_{TC} = P_{DT} V_T \quad (39)$$

where P_{DT} is the core loss density of the transformer given by the manufacturer for the operating frequency and magnetic flux density, and V_T is its volume.

For the copper losses P_{TW} of the transformer, it is necessary to determine the effective resistance of the primary winding R_p and the secondary winding R_s , taking into account the conductivity, the total length, and the skin effect. The losses are obtained through

$$P_{TW} = I_o(\varphi, TR)^2 (R_p + R_s). \quad (40)$$

Finally, the total losses are obtained from

$$P_{tot}(\varphi, TR) = 4P_{cd}(\varphi, TR) + 4P_{sw}(\varphi, TR) + 4P_{gate} + P_{TC} + P_{TW}. \quad (41)$$

Meanwhile, the efficiency is given by

$$\eta(\varphi, TR) = \frac{P_o(\varphi, TR)}{P_o(\varphi, TR) + P_{tot}(\varphi, TR)}. \quad (42)$$

4.3. Comparison With Fixed and Variable Frequency

Compared to the fixed frequency design, it is necessary to increase the switching frequency to achieve a larger angle between the output voltage and current [7]; for a Q of 10, an angle of 35° is estimated. Therefore, using Equation (18), the switching frequency for that angle is obtained.

On the other hand, in the FM&PS regulation, the frequency is varied such that the value of α is approximately equal to $\varphi/2$, thus maintaining a constant angle [29]. Therefore, Equation (18) can be rewritten as

$$\omega_{sw} = \omega_o \frac{\tan \frac{\varphi}{2} + \sqrt{\tan^2 \frac{\varphi}{2} + 4Q^2}}{2Q}. \quad (43)$$

The equivalent resistance in the secondary side will increase with respect to frequency according to Equation (6), resulting in Equations (32) and (33) being rewritten as

$$I_o(\varphi, \omega_{sw}) = \frac{2\sqrt{2}Q}{\pi L_s \omega_{sw} n^2} V_d \cos \frac{\varphi}{2} \cos \left(\alpha + \frac{\varphi}{2} \right). \quad (44)$$

for the current and for the power as

$$P_o(\varphi, \omega_{sw}) = \left(\frac{2\sqrt{2}V_d}{\pi} \right)^2 \frac{Q}{L_s \omega_{sw} n^2} \cos^2 \frac{\varphi}{2} \cos^2 \left(\alpha + \frac{\varphi}{2} \right). \quad (45)$$

Table 2 summarizes the differences in design due to the different types of controls, while Table 3 reflects the differences in losses involved in each of the designs operating at 10 kW. The advantages of using SiC MOSFETs in high-frequency induction applications have already been analyzed in previous work [31]. Therefore, for the loss analysis, the C3M0120100K was used as the transistor for each of the inverter switches, as it not only met the specifications for operation in the safe area but also offered the best performance when calculating the losses. The specifications of this transistor can be found in Table 4. For the two ferrite cores of the transformer, the material selected was N27 with shape reference E 70/33/32 from TDK Electronics.

Table 2. The differences in the operating characteristics of the controls.

Magnitude	Symb.	PS		FM&PS		VT&PS		Unit
Switching frequency	f_{sw}	243		203	278	203		kHz
Switching angle	α	35	15	15		15		°
RMS output current	I_o	38	19	32	9.7	32	8.5	A
Transformer ratio	$TR_{max/min}$	10.7		10.7		10.7	20	

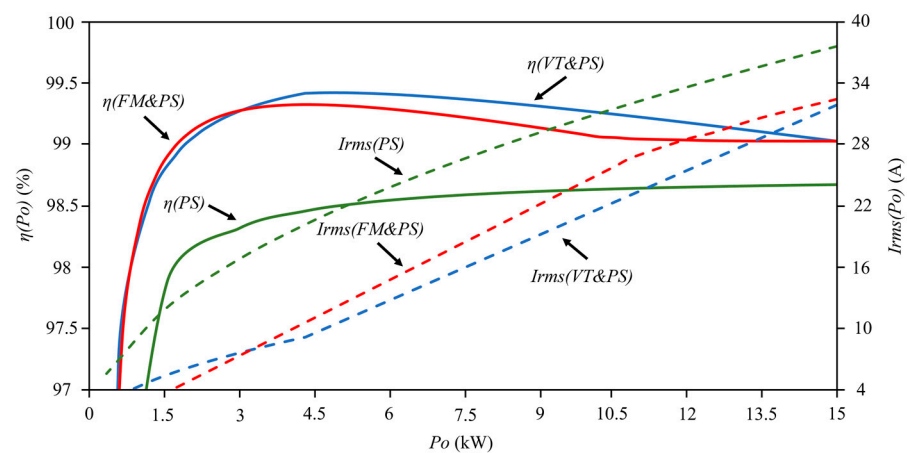
Table 3. Losses for 10 kW operation power.

Magnitude	Symb.	PS	FM&PS	VT&PS	Unit
Conduction losses	P_{cd}	30.92	19.78	16.72	W
Switching losses	P_{sw}	3.83	4.26	3.75	W
Gate losses	P_{gate}	0.115	0.128	0.112	W
Core losses	P_{TC}	10.2	11.22	10.2	W
Wire losses	P_{TW}	9.8	9.28	10.03	W
Total losses	P_{tot}	159.47	117.16	102.57	W

Table 4. Transistor characteristics.

Magnitude	Symb.	Equation	Value	Unit
On resistance	$R_{DS_{on}}$	(34)	120	m Ω
Turn OFF losses	a	(35)	0.0213	$\mu\text{J}/\text{A}^2$
	b		0.436	$\mu\text{J}/\text{A}$
	c		18.991	μJ
Gate charge	Q_G	(38)	22	nC

To observe the improvement in efficiency resulting from this new control compared to conventional ones, the variations in efficiency and output current with respect to the operating power are plotted in Figure 7. As observed when comparing the VT&PS and FM&PS controls, the starting point at maximum power is similar, but the VT&PS control increases efficiency as the current decreases because the transformer increases its ratio, adapting the load to the optimal operating point. The FM&PS control exhibits a similar behavior, but increasing the frequency as the power decreases does not reduce losses as effectively. On the other hand, the PS control maintains a lower efficiency from the beginning because the output current is higher due to having to operate with a larger switching angle.

**Figure 7.** The calculated evolutions of the efficiency and output current of the inverter as functions of the output power for the three control strategies.

5. Experimental Results

To obtain the experimental results of the system, first, the design of the variable transformer was evaluated. Table 5 compares the theoretical values used for the transformer design with the measured values from the short-circuit and open-circuit tests for the primary and secondary of the transformer. These measurements were made to check the

assumptions of the coupling factors both in the primary and secondary for the saturated and unsaturated transformers. On the other hand, Table 6 checks the transformer ratios for different levels of DC bias and how this influences the primary and the secondary inductance. As can be seen in both tables, there are deviations in the values due to component dispersions and possible measurement errors.

Table 5. Theoretical and measured values of transformer inductances.

Control Current	Theoretical Inductance (μH)				Measured Inductance (μH)			
	Open Circuit		Short Circuit		Open Circuit		Short Circuit	
I_{dc} (A)	L_{poc}	L_{soc}	L_{pcc}	L_{scc}	L_{poc}	L_{soc}	L_{pcc}	L_{scc}
0	1198	85.5	1018	84.2	1222	90.5	1100	89.6
1	244.4	38.18	48.9	7.64	254	40.5	54.1	7.99

Table 6. Theoretical and measured values of transformer ratios.

Control Current	Theoretical			Measured		
	L_p (μH)	L_s (μH)	TR	L_p (μH)	L_s (μH)	TR
I_{dc} (A)						
0	1198	85.5	5.6	1222	90.51	7.02
0.25	820.9	65.2	9.5	767.2	60.854	10.7
0.5	425.6	44.9	13.2	435.5	41.281	12.9
0.75	296.52	39.5	16.3	300.2	42.613	16.1
1	244.4	38.18	20	254.16	40.49	20.8

To carry out the measurements that verify the accuracy of the entire design, a 15 kW inverter module was used, which integrates the FPGA to implement the control, sensors, drivers, transistors, and a controllable current source based on the Howland current pump with an OPA548 as a high-current operational amplifier. The inverter is cooled by a water cooler, and the power supply comes from a controllable source. Table 7 shows the main part numbers of the components used in the inverter.

Table 7. The main components of the inverter.

Component	Manufacturer	Reference
Transistor	Wolfspeed, Durham, NC, USA, EE. UU.	C3M0120100K
Gate driver	Infineon, Neubiberg, Germany	1ED3120MC12H
Output current sensor	Allegro, Manchester UK	ACS37003KMCATR-085B5
DC current sensor	Infineon, Neubiberg, Germany	TLE4971
Hight current op-amp	Texas Instruments, Dallas, TX, USA, EE.UU.	OPA548
FPGA	AMD, Santa Clara, CA, USA, EE. UU.	Zynq 7000

The test bench used is shown in Figure 8, and each number corresponds to the following:

- (1) Induction heating inverter with four SiC C3M0120100K;
- (2) DSO with 300 MHz bandwidth;
- (3) Differential voltage probe and Rogowski current probe;

- (4) Hall effect probe;
- (5) Variable transformer;
- (6) Capacitor (C_s);
- (7) Inductor (L_s);
- (8) Workpiece.

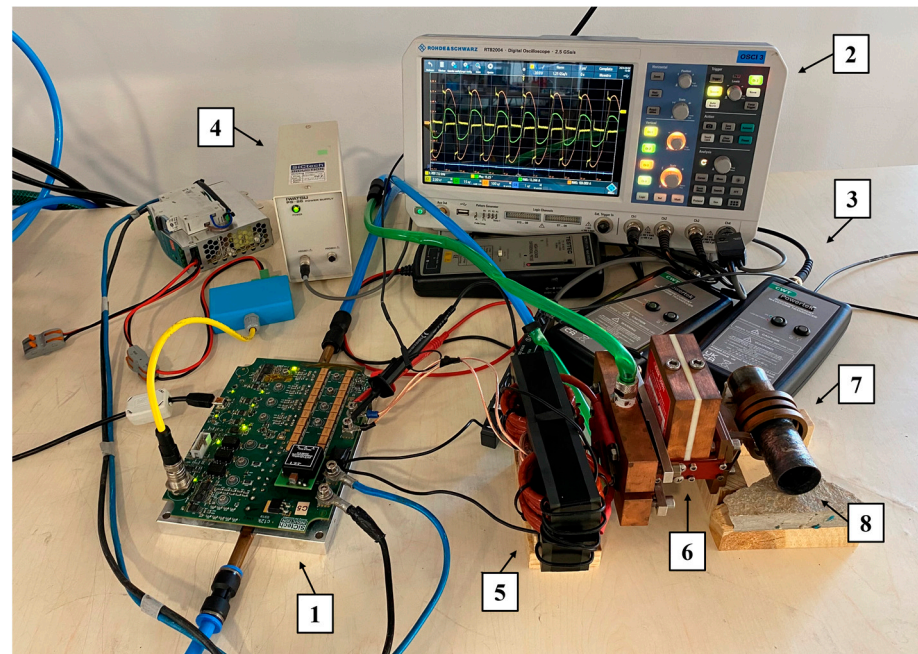


Figure 8. The inverter, variable transformer, and LC resonant load test bench. The numbered labels describe the components on the bottom.

In Table 8, the measured values of the main quantities involved in the design are compared with the theoretical values at two power levels. As can be seen, there is no considerable variation; however, there are deviations due to the parasitic elements of the system that were not considered in the design.

Table 8. Comparison of theoretical and measured results.

Magnitude	Symb.	10 kW		15 kW		Unit
		Theor.	Meas.	Theor.	Meas.	
Phase angle	α	15	15.04	15	14.9	$^{\circ}$
Transformer ratio	TR	12.5	12.51	10.7	10.82	
Output current	I_{RMS}	22.7	22.85	31.9	32.25	A

Figure 9 shows the oscillograms captured by the DSO for the inverter operating at 10 kW of power with three different control techniques and their responses to changes in the load by quickly removing the workpiece from the inductor. For all three control methods, the results obtained from the designs in Table 2 were used. As observed in the left column, the fixed-frequency PS control requires operation at a higher frequency, resulting in a 33° switching angle at that power level. On the other hand, the other two controls, having two variables, can maintain a 15° angle. In the case of FM&PS, this is due to the increased frequency to 229.8 kHz, and in the case of VT&PS, it is due to impedance adaptation of the load by increasing the transformer ratio to 12.5. Additionally, this load adaptation allows for a lower output current from the inverter compared to the other two controls, which is 22.85 A RMS. Furthermore, it is observed that with a more adapted load due to the higher

ratio, the width of the phase shift angle is smaller, reducing losses from OFF switching of the transistors as they have to switch off at a lower current level.

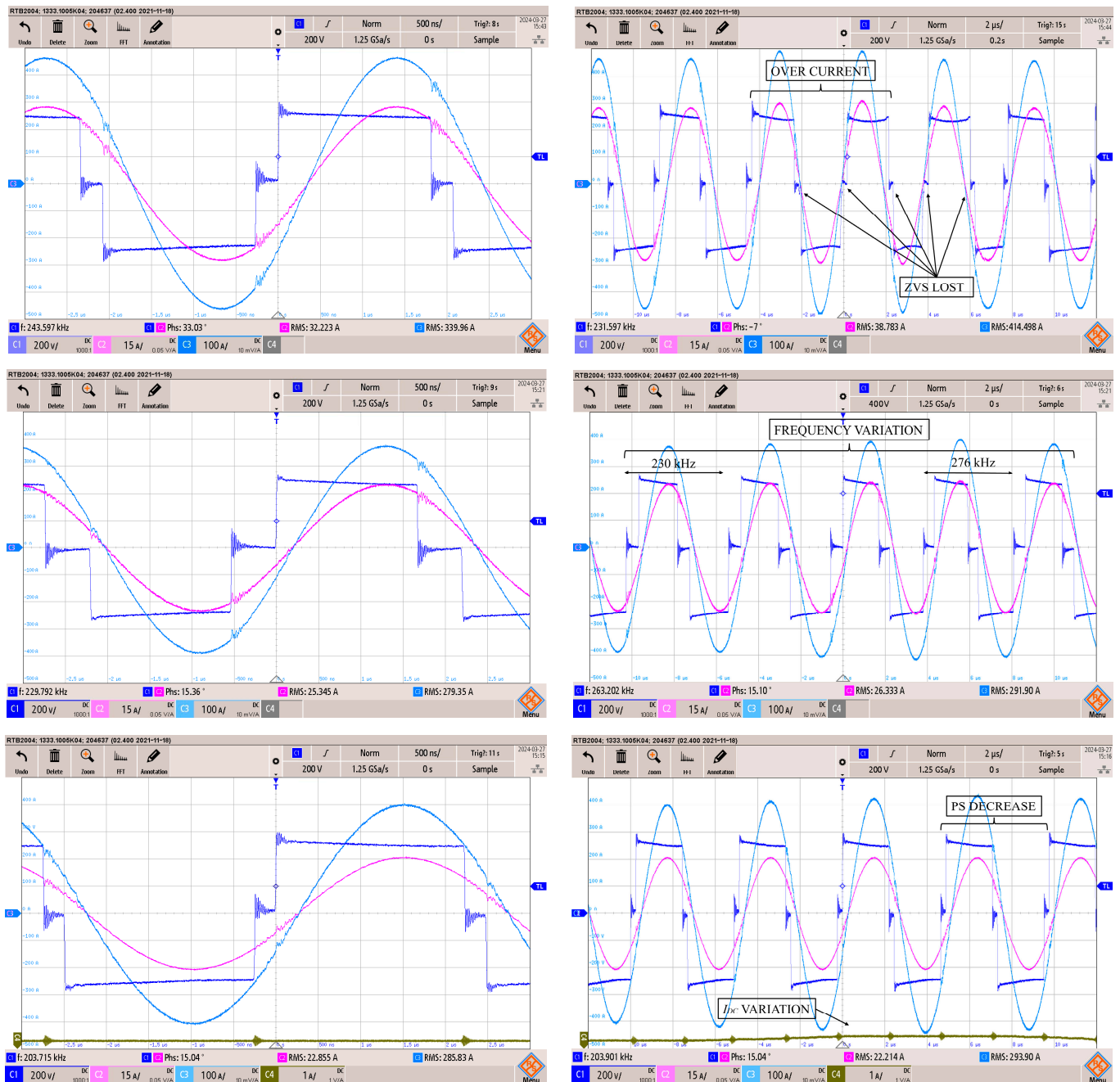


Figure 9. Oscillograms of experimental tests for the LC inverter with different control strategies: PS in the first row, FM&PS in the second row, and VT&PS in the third row. The left column represents operation at 10 kW, and the right column shows the response to a load change at that power level. C1 (dark blue) represents the output voltage (200 V/div), C2 (magenta) shows the output current (15 A/div), C3 (light blue) shows the output current at the secondary of the transformer (100 A/div), and C4 (brown) displays the DC bias current (1 A/div) of the variable transformer in the third row. The time base is 500 ns/div in the left column and 2 μs/div in the right column.

On the other hand, in response to load changes, the PS control, lacking direct regulation with the phase between the output voltage and current, loses ZVS in five commutations, and an overcurrent in the inverter is observed. Both consequences lead to increased losses

in the inverter. In the case of FM&PS control, the angle is maintained, but at the expense of changing the frequency from 230 to 267 kHz, which complicates filtering and reduces the optimization of the reactive elements of the resonant circuit. Lastly, VT&PS control is able to maintain the angle by adapting the circuit impedance with an increase in the transformer ratio from 12.5 to 14.3, allowing for a fixed frequency with ZVS.

Finally, to conclude the comparison among the three types of control and validate the design discussed in the article, the inverter efficiency was measured across the entire power range by calculating the difference between the input power measured directly at the DC voltage supply and the output power measured by instantaneously multiplying the voltage and output current of the inverter bridge using the DSO [31]. Starting from maximum power, as shown in Figure 10, both FM&PS and VT&PS controls achieve the same efficiency because the transformer ratio and frequency are identical. On the other hand, the PS control begins with lower efficiency due to its higher phase shift angle between the voltage and current and the higher frequency. However, as the power decreases, VT&PS control gradually increases its efficiency due to impedance matching by increasing the transformer ratio, unlike FM&PS control, which achieves lower efficiency due to increased output current and frequency. This demonstrates a superior performance of over 99% across the entire operating range, which is higher with VT&PS control compared to FM&PS. As observed in the figure, the measurements closely align with the theoretically calculated values despite the dispersion in real components and the losses not accounted for in the theoretical calculation.

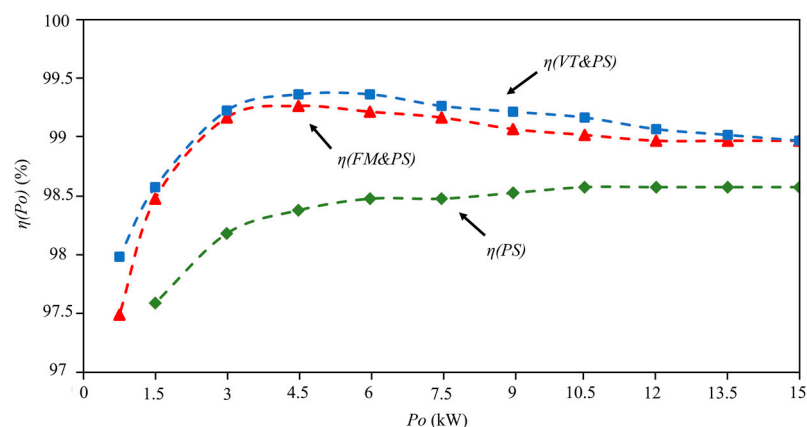


Figure 10. The results of the experimental measurements of the inverter efficiency across the entire power range.

6. Conclusions

This work introduces a novel control technique (VT&PS) which, through the use of a current controlled variable transformer and phase shift, allows for fixed frequency operation close to the resonant frequency in a series LC resonant. This allows for higher efficiency over the entire operating range, a low switching angle, and operation over a wide range of loads while maintaining ZVS.

A new variable transformer design, integrated with a series LC resonant design for induction heating applications, is analyzed and mathematically modeled. The optimal operating point at which the inverter output current is reduced and the implementation of the inverter control scheme are proposed.

The design is corroborated by measuring the values of the transformer inductances and using it together with a 15 kW induction heating inverter with SiC Mosfets and the proposed control. The efficiency and load change behaviors are measured for the proposed control and the two conventional alternatives.

This demonstrates that the inverter is capable of operating at a fixed frequency, close to the resonant frequency, with a fixed angle that reduces the output current, achieving an

efficiency of over 99%, and that it is able to operate in the face of the load changes effects common in induction heating applications.

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