A Transformer Design for High-Voltage Application Using LLC Resonant Converter

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Abstract: The inductor–inductor–capacitor (LLC) resonant converter is a suitable topology for wide output voltage and load range applications with limited circuit parameters. One of the most significant design boundaries of an LLC resonant converter in high-voltage applications is the parasitic capacitance effect of the main circuit components, particularly the transformer and junction capacitances of the secondary rectifier network. Parasitic capacitance effects are much higher in high-voltage applications than in low-voltage applications. Therefore, the use of an LLC resonant converter is limited to high-voltage applications. This study proposes to reduce the capacitive effects of high-voltage transformers and rectification networks with a multi-winding transformer with an integrated rectifier design and to use it in high-voltage applications with the advantages of the LLC resonant converter. For the proposed prototype, comparative experimental measurements were conducted using a conventional scheme. The measurements validate the reliability of the LLC converter for high-voltage applications, improving the output regulation performance while significantly reducing parasitic capacitances.

Keywords: high-voltage transformers; parasitic capacitance; resonant power conversion; transformer model

1. Introduction

In high-frequency power-converter applications, the weight and volume of the converter decrease by increasing operating frequency. Thus, increasing the power density improves the controllability and efficiency of the system. However, increasing the frequency increases the impact of parasitic capacitive elements of transformers, diode rectifiers, and switching elements, which are the main components of the power converter [1]. Parasitic capacitive elements affect the selection of the resonant topology and the operating boundaries of the power converter. In high-voltage applications, parasitic capacitive element values are more effective than those in conventional low-voltage applications, owing to the high turn ratio of the step-up transformer. According to the chosen application, parasitic elements must be considered as the main elements of the resonant circuit or their effects on the resonant circuit should be reduced and neglected [2,3]. Therefore, the choice of the resonant topology depends on the value of the main elements of the circuit relative to the parasitic elements. In other words, the desired resonant topology can be implemented more efficiently and reliably if the proportional size of parasitic elements can be designed for the proposed converter.

The main obstacle to increasing the frequency is the limit of the switching losses, which can be reduced by applying different soft-switching converters. One of the most popular topologies is the LLC resonant converter, which offers load and line regulation for wide load and voltage ranges [4,5], inherited short-circuit capability, and high efficiency. Moreover, the LLC resonant converter provides zero-voltage switching (ZVS) at the primary switches and zero-current switching (ZCS) at the rectifier diodes on the secondary side with low circulating currents, according to the fundamental harmonic approximation (FHA) and time.
domain analysis [6]. Modeling equivalent circuits with an FHA characterizes the DC gain for the operating frequency, soft-switching limitation, and basic voltage–current waveforms without complex analysis [7]. The FHA method provides limited reliability, whereas the multiple parasitic element values are unbalanced and high compared to the selected main element of the converter [8]. LLC resonant converter applications do not consider the effects of the parasitic capacitive elements. However, the effect of parasitic capacitance elements on the resonant circuit should be determined, evaluated, and designed in compliance with these elements. The parasitic effects of the primary switches have been evaluated [1,6–8]. The parasitic capacitance of rectifier networks was examined in [9]. In practice, a capacitor or auxiliary resonant circuit is added to the LLC resonant converter to reduce the effects of parasitic elements [7,8,10]. An LLC resonant converter for high-voltage applications was evaluated [11] without considering the transformer parasitic capacitive elements. The energy stored in the electric field is defined as the parasitic capacitive elements of the transformer that cause large distortions in the current and voltage waveforms, and ZVS losses under light-load conditions in the LLC resonant converter. The increase in parasitic effects necessitates a more comprehensive analysis that FHA does not predict. Therefore, to use an LLC resonant converter for high-voltage applications, a more complex analysis should be performed, or the parasitic effects should be reduced to negligible values by using FHA.

In many studies, methods for calculating the parasitic capacitance values of two-winding [9,12–15] and multi-winding transformers [16,17] have been defined and evaluated. The capacitive elements cannot be neglected. However, designers can reduce their values and effects to circuits. Various approaches have been adopted to address this problem. In [18], transformer parasitic capacitive elements were evaluated and reduced by the winding arrangement, which is difficult to use in high-voltage applications. In [19], the authors experimentally evaluated the concept of reducing transformer capacitive parasitic elements with a layered integrated rectifier by separating stored energy in the electric field AC and DC components to reduce transformer capacitive parasitic elements. The effects of the rectification network selection, number of layers, and magnetic effects were not discussed.

The objectives of this study were to decrease capacitive elements on high-voltage transformers with a winding arrangement, minimize the arrangement’s influence on winding losses and test a proposed winding arrangement with an LLC resonant converter which is sensitive to capacitive elements. The capacitive effects on the LLC resonant converter are presented in Section 2. The parasitic capacitance effects of the high-frequency transformer, rectifier network, and integration of the rectifier and high-frequency transformers are modeled and analyzed in Section 3. The proposed method for a multi-winding transformer with an integrated rectifier to reduce the effect of parasitic capacitance is presented in Section 4. The verification of the established model, comparative measurements, and application of the LLC resonant converter model with the FHA are presented in Section 5. Section 6 provides a discussion and the conclusions of the study.

2. LLC Resonant Converter

The influence of transformer capacitive parasitic elements on LLC resonant converters are presented in this section. Besides the resonant elements conventionally used in the LLC topology, \( L_m \) is the mutual inductance, \( L_{lk} \) is the leakage inductance, \( C_r \) is the resonant capacitance, and the parasitic effects arising from the circuit components; \( C_{tr} \) and \( C_{j} \) are defined as equivalent transformer capacitance referred to primary side and secondary rectifier network junction capacitance, respectively. \( C_{stray} \) is the sum of \( C_{tr} \) and \( C_j' \), where \( C_j' \) is effects of \( C_j \) referred to primary side. \( C_{in1} \) and \( C_{in2} \) are primary switch parasitic capacitances. In Figure 1, the LLC resonant converter is shown to involve parasitic elements in conventional applications. The leakage inductance of the transformer, \( L_{lk} \), can be used as an element of the resonant circuit as a serial resonant inductance in several applications,
and it can also be used with $L_r$, $L_{dk}$ by adding an external inductance, depending on the transformer design.

![Figure 1. LLC resonant converter with parasitic components.](image)

$R_o$, $R_{eq}$, and $N$ are the load resistance, equivalent load resistance, and the turn ratio, respectively.

$$R_{eq} = 8 R_o / N^2 n^2$$

(1)

To analyze the LLC resonant converter, the series resonant frequency $f_{RES}$ of $L_r$ and $C_r$ in series and the pole frequency $f_p$ of $(L_r + L_m)$ and $C_r$, are given by:

$$f_{RES} = 1/2 \pi \sqrt{L_r C_r}$$

(2)

and

$$f_p = 1/2 \pi \sqrt{(L_r + L_m) C_r}$$

(3)

The normalized frequency is $f_n$ while, resonant converter switching frequency is $f_{SW}$ and the quality factor of the converter $Q$ can be expressed as follows:

$$f_n = \frac{f_{SW}}{f_{RES}}$$

(4)

$$Q = \frac{L_r}{C_r R_{eq}}$$

(5)

According to the FHA analysis, while $V_{out}$ is the output voltage, $V_i$ is the input voltage, $L_n = L_m / L_r$ is the inductance ratio and the transfer ratio $G$ is expressed as [11]:

$$G = \left| \frac{V_{out}}{N V_i/2} \right| \frac{1}{\sqrt{\left( \frac{1}{L_n} + 1 - \frac{1}{L_n f_n} \right)^2 + \frac{4}{\pi^2} Q^2 \left( \frac{1}{f_n} - f_n \right)^2}}$$

(6)

When $f_{SW}$ is higher than $f_{RES}$, $G$ decreases proportionally with increasing frequency. Under all circumstances, it operates in the inductive region and $G \leq 1$. In contrast, if $f_p < f_{SW} < f_{RES}$ it operates in the inductive region, depending on $Q$ and $G \geq 1$. If $f_{SW} = f_{RES}$, it operates under all conditions in the inductive region and always $G \approx 1$. The primary switches operate in ZVS, whereas the converter operates in the inductive region.

FHA analysis is only applicable when the parasitic capacitance effects and values are larger than those of the main resonant elements. In Figure 2, the DC gain curve of the LLC resonant circuit is shown in the design load and frequency range according to the FHA, while the quality factor varies. Because parasitic effects were neglected, the voltage gain decreased in proportion to the frequency increment even under light-load conditions above
the resonant frequency. ZVS on the primary-side switches were achieved specifically under light-load conditions, whereas \( f_n > 1 \).

![Figure 2. LLC resonant converter voltage gain according to FHA.](image)

Although parasitic capacitances have negligible values in many applications, the equivalent stray capacitance, \( C_{\text{stray}} \), reaches high values that cannot be neglected in high-voltage applications, owing to the high turn ratio in the transformer. To observe the \( C_{\text{stray}} \) effect more clearly, \( C_n \) should be defined as the ratio of \( C_{\text{stray}} \) to \( C_r \) [20]:

\[
C_n = \frac{C_{\text{stray}}}{C_r}
\]  

(7)

These parasitic effects narrow down the operating limits of the LLC resonant circuit and disturb the soft-switching conditions of the power elements and thereby affect the operation of the converter [1,9,18]. \( C_{\text{stray}} \) causes a high primary charge, distortion of the current and voltage waveforms on the primary side, and loss of ZVS in the primary-side switches, particularly when \( f_n > 1 \) under light-load conditions. Figure 3 shows the DC gain curve for the design load and frequency range including different value of \( C_{\text{stray}} \) while the quality factor varies. The DC gain, at above the resonance frequency, increases a little in proportion to the switching frequency, under light-load conditions where \( C_n \) is equal to 0.1. The DC gain, at above the resonance frequency, increases excessively in proportion to the switching frequency, under light-load conditions where \( C_n \) is equal to 0.5. The operating frequency limits of the LLC resonant circuit are decreased, depending on the \( C_n \) increase.
Figure 3. LLC resonant converter voltage gain with effect (a) $C_n = 0.1$, (b) $C_n = 0.5$.

Figure 4 shows the effect of $C_n$ change under light-load conditions on the mode of operation of the LLC resonant converter and over the frequency limit $f_{max}$ where soft switching is disturbed under any circumstance. $C_{stray}$ causes a high primary charge, distortion of the current and voltage waveforms on the primary side, and loss of ZVS in the primary-side switches, particularly when $f_n > 1$ under light load conditions. As $C_{stray}$ increases, and the ZVS loss frequency ($f_{max}$) gets close to $f_{RES}$. 
The parasitic capacitive effects of the transformer can be reduced by increasing $C_r$ and $L_r$ depending on the quality factor or decreasing $L_m$ [6]. Increasing $C_r$ and $L_r$ leads to a significant decrease in the switching frequency, raising the switching transmission losses to a high level. Decreasing $L_m$ is possible by adding more air-gap to the transformer and decreases transformer efficiency. Even this situation requires more complex designs [8]. Detailed time domain and efficiency analysis must evaluate the influence of parasitic component on the resonant converter [1,5]. LLC resonant converter primary switches cannot operate under ZVS conditions while $C_n$ increases significantly. The use of an LLC resonant converter in high-voltage applications is possible by reducing the effects and values of the main parasitic element, $C_{stray}$, which is not included in the system as the main component in the converter model. For this reason, the design made with the FHA has been made applicable, especially by investigating the high-frequency transformer model and reducing the interference effects.

3. Modelling and Measuring Stray Capacitance

The parasitic effects of the high-frequency transformer, integration of the rectifier and multi-winding high-frequency transformers are modeled and analyzed in this section.

3.1. High-Frequency Transformer

The modeling and measurement of high-frequency transformers have been evaluated in many studies [12–15]. A model of the transformer, including the parasitic elements, is presented in Figure 5, where $N$ is defined as the turn ratio.

![High-frequency transformer including parasitic elements.](image)
The inductive elements \( L_{m}, L_{ikp}, \) and \( L_{iks} \) are defined as mutual, primary winding leakage, and secondary winding leakage inductances, respectively. Inductive elements are defined as magnetic field effects that can be directly measured at low frequencies using short-circuit and open-circuit tests to minimize the effects of capacitive elements. After the measurements, the primary and secondary leakage inductances can be summed into a single equivalent leakage inductance component, \( L_{ik} \), referred to the primary side.

\[
L_{ik} = L_{ikp} + \frac{1}{N^2} \delta_s
\]  

AC winding losses in transformers increase with high frequency due to proximity effects, including skin effect and proximity effect losses and leakage flux will cause higher leakage losses. The ratio of AC resistance \( (R_{ac}) \) to DC resistance \( (R_{dc}) \) including skin effect of an infinite foil conductor with fixed frequency sinusoidal current is presented in (9), where \( \Delta = t/\delta \), and \( t \) and \( \delta \) are foil thickness and the skin depth, respectively [21].

\[
\frac{R_{ac}}{R_{dc}} = \frac{\Delta \sinh(\Delta) + \sin(\Delta)}{2 \cosh(\Delta) - \cos(\Delta)}.
\]  

Based on Dowell’s assumptions ratio of \( R_{ac} \) to \( R_{dc} \), \( F_R \), which represents the proximity effect of \( p \)th layers for foil conductors and fixed frequency sinusoidal current delivered as in [22–24]:

\[
F_R = \frac{R_{ac,p}}{R_{dc,p}} = \frac{\Delta}{2} \left[ \frac{\sinh(\Delta) + \sin(\Delta)}{\cosh(\Delta) - \cos(\Delta)} + \frac{2}{3} (p^2 - 1) \frac{\sinh(\Delta) - \sin(\Delta)}{\cosh(\Delta) + \cos(\Delta)} \right],
\]  

where \( p \) is the number of layers and can be defined as a ratio magneto motive force (MMF) in the highest layer of MMF distribution. Proximity-effect losses dominate the AC losses with increasing number of layers. Proximity effect depends on winding arrangement including the layer porosity factor and layer distances [25,26]. MMF distribution of the \( y \) axis different winding arrangement is presented in Figure 6 where winding symmetry along the \( y \) axis and each primary and secondary turn ratio are the same. \( t_p \) and \( t_s \) are foil thickness on primary and secondary windings, respectively. \( h \) and \( h_{p-s} \) are insulator thicknesses and increase leakage inductance like porosity factor [26].

![Figure 6. MFF distributions](image-url)
There is a contradiction for \( h, h_{p-s} \) and porosity factor between the leakage inductance and the transformer equivalent capacitance. By increasing insulator thickness or decreasing porosity factor, transformer equivalent capacitance decreases, and leakage inductance increases. Decreasing both leakage inductance and transformer equivalent capacitance is not possible because of the physical parameters. However, it is possible to obtain optimal behaviors for both leakage inductance and stray transformer equivalent capacitance with winding arrangements [26]. \( R_{ac} \) increases by increasing the number of turns in each layer and increasing \( h \) and \( h_{p-s} \). Additionally, \( R_{ac} \) increases by width of layer \( (b_w) \). \( R_{ac} \) of the turns in each layer increases by increasing the distance between turns along the width of the layer \( (b_w) \) and core, by distance between turns along the width of layer \( (b_w) \) and primary winding especially in non-interleaving arrangements [21].

Selecting thickness of the conductor is dominating \( F_R \) for layers. With decreasing of the thickness less than \( \delta \), \( F_R \) is close to 1 so that proximity losses will be minimal. Figure 7 shows that the ratio of AC resistance to DC resistance increases with increasing thickness at a fixed frequency. In DC-DC converter applications transformers are switched to non-sinusoidal current waveforms except for the resonance frequency. The optimum value of \( \Delta \) is [27,28]:

\[
\Delta_{opt} = \frac{1}{\sqrt{\frac{5p^2-1}{15}}} \sqrt{\frac{\omega I_{rms}}{I'_{rms}}},
\]

where \( I_{rms} \) is the rms-value of the converter current, and \( I'_{rms} \) the rms-value of the first derivative of converter current \( \omega \) is the fundamental frequency. Selecting \( F_R \approx 1.3 \) will be the optimum \( R_{ac} \) [27]. The influence of the fringing in the air-gap field will cause \( R_{gap} \) to be in the close proximity to the air gap [25]. \( R_{gap} \) can be decreased by increasing the distance between the windings and air gap out of notch radius [28]. Total winding losses, \( R_{tot} \), can be expressed where \( R_{tot} \) is the sum of primary winding resistance \( (R_p) \) and secondary winding resistance \( (R_s) \).

\[
R_p = R_{dc(p)} + R_{ac(p)} + R_{gap(p)}
\]

\[
R_s = R_{dc(s)} + R_{ac(s)} + R_{gap(s)}
\]

\[
R_{tot} = R_p + R_s
\]

The capacitances are defined as electric field effects of the transformer. The capacitances of the transformers \( C_p, C_s, \) and \( C_{p-s} \) have been defined as the primary winding capacitance, secondary winding capacitance, and primary-secondary coupling capacitance, respectively [13]. Unlike inductive components, capacitive elements cannot be measured directly [29]. The capacitive elements should be measured as much as possible from the magnetic core [9]. However, the parasitic capacitances of the transformer can be shown as a single equivalent winding capacitance, \( C_{tr} \), using the network and step response approach techniques. \( C_{tr} \) can be shown on the primary side that can be expressed with scattered or parasitic electrical coupling between windings while all parameters referred to primary side are indicated [9,13,30] by an apostrophe (’). A simplified equivalent model of the primary side is shown in Figure 8.

\[
C'_p = C_p + (1 - N) \cdot C_{p-s}
\]

\[
C'_s = N^2 C_s + N \cdot (N - 1) \cdot C_{p-s}
\]

\[
C_{tr} \cong C'_p + C'_s
\]
Figure 7. The ratio of AC resistance to DC resistance as a function of $\Delta$ and MMF ratio $p$. 
Two resonance frequencies, $f_{series}$ and $f_{parallel}$, are defined as series and parallel resonance frequencies of the transformer, respectively. $f_{series}$ is determined by $L_{lk}$ and $C_{tr}$ and $f_{parallel}$ are determined by $(L_m + L_{lk})$ and $C_{tr}$ [29].

$$f_{series} = \frac{1}{2\pi\sqrt{L_{lk}C_{tr}}}$$  \hspace{1cm} (18)

$$f_{parallel} = \frac{1}{2\pi\sqrt{(L_m + L_{lk})C_{tr}}}$$  \hspace{1cm} (19)

$C_{p-s}$, which creates a low-impedance path between the primary and secondary windings and gives rise to EMI noise, can be reduced by shielding [29,31] or increasing $h_{p-s}$ thickness. $C_p$ and $C_s$ can be reduced by increasing the spacing between the windings, reducing the parallel surfaces, increasing the spacing between the layers, and reducing the potential differences between the parallel faces [18,32]. Making the windings in layers due to electrical insulation is necessary in high-voltage applications; however, decreasing the turn–turn layer capacitances of windings in the layers reduces the $C_s$ [15,17]. However, these methods increase the leakage inductance because they also affect magnetic coupling even if the $C_s$ effect is reduced [19].

In high-voltage applications, the main component of $C_{tr}$ is $C_s$ because $C_s$ must be multiplied by the square of the high turn ratio $N$. The $C_s$ effect is reduced by decreasing the electric field energy created by reducing the turn–turn capacitances and parallel surfaces or the potential differences between parallel surfaces, as specified in [18]. The electric field effects arising from the high-voltage conversion ratio of the transformer cannot be significantly reduced by adjusting only the winding geometry.

### 3.2. Rectifier Network

The modeling should consider the impact of the secondary rectifier network junction capacitance $C_j$ on the resonant circuit while $C_j$ leakage inductance resonant frequency and converter switching frequency are convergent [1]. Various types of rectifier networks can be implemented, such as center-tapped, full-bridge, voltage doubler, and voltage multipliers. In [9], the junction capacitance referred to primary $C_j'$ is compared for different rectifier network selections. $C_j'$ will be half by selecting voltage-doubler compared to full-bridge for same output voltage. Thus, the turn ratio will be half and the $C_j'$ will be half. A voltage multiplier can be selected as the output current decreases, whereas the output voltage drops and voltage ripple increases proportionally with the increase in the number of voltage-multiplier stages and output current [33]. If the turn ratio is low, $L_m$ is significantly higher than $L_{iks}$ so $C_j$ is parallel to secondary winding capacitance and its effect will be proximate.

$$C_j' = N^2 \cdot C_j$$  \hspace{1cm} (20)

In order to define $C_{stray}$, the effects of $C_{junc}$ must be taken into account. The equivalent stray capacitance referred to primary side $C_{stray}$ can be expressed as the sum of $C_{tr}$ and effect $C_j$.

$$C_{stray} \approx C_{tr} + C_j'$$  \hspace{1cm} (21)

**Figure 8.** Simplified high-frequency transformer model, referred to primary side.
3.3. Multi-Winding Transformer

A multi-winding transformer includes more than one primary and/or secondary winding. Each winding can have a different turn ratio. In this study one primary windings and multi secondary windings transformer are evaluated. The storage energy of the electrical field has two components which are \( C_{ac} \) and \( C_{dc} \). \( C_{dc} \) unlike \( C_{ac} \), is not charged and discharged during one period of the switching frequency \[19\]. Thus, effects of \( C_{dc} \) to resonant converter are limited. \( C_{ac} \) and \( C_{dc} \) can be separated by multi-winding transformers where the output of the rectifier network are serially connected; however, voltage distribution of the output, winding arrangement and effects of multi-winding transformer to resonant converter must be evaluated. Each winding can have a different turn ratio. Winding arrangements such as multi-layer and multi-disk are presented in Figure 9 \[34–36\].

![Winding arrangements](image)

**Figure 9.** Winding arrangements (a) multi-layer (b) multi-disk (c) mixed.

A multi-winding transformer with an integrated rectifier for \( z \) secondary windings is shown in Figure 10. Mutual inductance can be separated by mutual inductance of the core \((L_{core})\) and secondary windings \((L_{ms})\) \[37\]. \( C_{p−S} , C_{S−S} , C_{p−i} , L_{ikS}, L_{ms}, R_{s} \) are indicated by the number of windings.

If the turn ratio is low, \( L_{ms} \) is significantly higher than \( L_{ikS} \), so \( C_{p−S} \) is parallel to secondary winding capacitance and its effect will be proximate.\( C'_{sz}, C'_{pz} \) are defined as the sum of the secondary winding capacitances, and sum of the primary–secondary coupling capacitances, respectively, for the number of \( z \) secondary windings. \( C_{tr} \) can be calculated using \( C'_{pz} \) and \( C'_{sz} \) by converting Equations (15) and (19) \[16,17,38,39\]. Calculation of \( C_{S−S} \) has another difficulty if winding voltage distribution is not equal \[19,38,39\].

\[
N_1 = N_{s1} / N_p, \ldots, N_z = N_{sz} / N_p
\]

\[
C'_{pz} = C_p + \sum_{i=1}^{z} (1 - N_i) \cdot C_{p−si}
\]

\[
C'_{sz} = C_p + \sum_{i=1}^{z} N_i^2 C_{si} + N_i^2 C_{pi} + N_i (N_i-1) \cdot C_{p−si}
\]

\[
C_{stray} \approx C'_{pz} + C'_{sz}
\]

The multi-winding transformer includes unwanted cross-coupling effects, which describe the interaction between secondary windings due to magnetic coupling \[34\]. The cross-coupling effects cause power transfer and voltage regulation problems \[40\]. \( M \) is the coupled inductance stack for secondary windings due to cross-coupling effects \[z8\]. The inductance matrix, \( L \), is a positive defined symmetrical matrix and includes coupled
inductance and leakage inductance of each winding \cite{41,42}. Coupling coefficient, $k$, presents the relationship between each winding to other windings. Matrix inductances can be measured and defined \cite{43,44}.

\[
L = \begin{bmatrix}
L_{11} & M_{1-2} & \cdots & M_{1-z} \\
M_{1-2} & L_{22} & \cdots & M_{2-z} \\
\vdots & \vdots & \ddots & \vdots \\
M_{1-z} & M_{2-z} & \cdots & L_{zz}
\end{bmatrix} \tag{26}
\]

Figure 10. Multi-winding transformer.

Once the inductance matrix is obtained, the relationship of output current ($i$) and voltage ($v$) can be calculated for each winding \cite{41}.

\[
i = \begin{bmatrix} i_1 \\ i_2 \\ \vdots \\ i_z \end{bmatrix}^T
\]
\[
v = \begin{bmatrix} v_1 \\ v_2 \\ \vdots \\ v_z \end{bmatrix}^T
\]
\[
v = L \frac{di}{dt}
\]
\[
k_{i-j} = \frac{M_{i-j}}{\sqrt{L_{ii}L_{jj}}}
\]
Since MMF distribution changes with winding arrangement, $R_S$ includes leakage inductance and coupling inductance effects [45,46]. The coupling coefficient is the maximum value for adjacent windings. By selecting a coupling coefficient $k \leq 0.2$, the effects of cross-coupling are very weak, some serial impedance keeps the same value independent of related winding and only introduces 4% inaccuracy on serial impedance [42,47].

4. Winding Arrangement of High-Voltage Multi-Winding Transformer with Integrated Rectifier

In this section, decreasing stray capacitance of the transformer with the proposed winding arrangement is presented. In high-voltage applications, the transformer includes high values for both leakage inductance and stray capacitance. In high-voltage transformers, the key point to study in the resonant converter topology is optimizing the contradiction between the leakage inductance and the stray capacitance. Both leakage inductance and stray capacitance values can be reduced separately. As leakage inductance decreases, stray capacitance increases and vice versa. The leakage inductance cannot be reduced because of the isolation requirement of high-voltage transformers. However, stray capacitance values can be reduced by the proposed winding arrangement. The main objective of the proposed winding arrangement is decreasing stray capacitance and minimizing that arrangement’s influence on leakage inductance and winding losses by considering cross-coupling effects. Since stray capacitance decreases to negligible values, distortion of the current and voltage waveforms on the primary side decrease and the gain of the converter increases as defined in Section 2. Thus, LLC resonant converters can be suitable to use in high-voltage applications. The proposed winding arrangement is presented in Figure 11. The winding direction of the primary coil is along the $y$ axis and the winding direction of the secondary coil is along the $x$ axis. All secondary windings are designed with PCB with the same geometry and turn ratio and a single layer.

![Figure 11. Proposed winding arrangement.](image)

If all secondary windings are connected in series before the rectifier network, it is multi-layer (sectional winding), which is used in industry in many applications. In multi-layer arrangements stray capacitance is extremely high, which is not suitable for LLC resonant converter applications; leakage inductance and winding losses can be calculated as defined in Section 3.1. If all secondary windings are connected in series after the rectifier network, it is a multi-winding transformer. Serial connection of secondary windings is used in many applications in industry. Since coupling relationships between windings are complex more complex analysis is required. In the proposed arrangement, the cross-coupling
effects between primary and all secondary windings are convergent unlike multi-layer and multi-disk.

DC current is the same for all windings. However, secondary winding voltage can vary, due to impedance of leakage inductance and cross-coupling effects. The output voltage distribution (OVD), for all secondary winding must be convergent. Coupling effects must be experimentally examined separately for each design.

There is a contradiction between each secondary winding turn ratio \( N \) and number of layers \( z \), for \( C_{\text{stray}} \) for the proposed arrangement where the total secondary windings are equal as defined in Section 3. Considering the coupling effect is same for all secondary windings, this contradiction can be examined as follows:

- Since all secondary windings are rectified and all layers are designed in the same geometry, \( C_{S-S} \) in all windings has only \( C_{dc} \) and the influence of the converter is negligible. Distance between secondary windings \( h \) and \( h_n \) cause only an increase of leakage inductance and has no effect on \( C_{S-S} \).
- Since distance between primary windings and all secondary windings
  \( (h_p-s) \)
  are the same (and provides the isolation requirement), secondary windings are rectified and all layers are in the same geometry, \( C_{P-S} \) includes \( C_{dc} \), which is negligible. \( C_{p-s} \) includes \( C_{ac} \) and is the same for all secondary windings.
- Since secondary windings are designed in the same geometry, \( C_s \) is the same for all secondary windings. In this case, from (23) and (24) \( C_{\text{stray}} \) can be expressed as a linear function of the output number \( z \).
- Increasing \( N \) increases \( C_s \) because of the turn-to-turn capacitance. \( C_s \) includes only \( C_{ac} \). The \( C_{\text{stray}} \) increase is multiplied by the square.
- Increasing \( z \), because of separation of \( C_{ac}, C_{dc} \), and \( C_{\text{stray}} \) increases proportionally \( C_{\text{stray}} \), which will be significantly reduced.

The resistance for each secondary winding depends on leakage inductance and cross-coupling effects. By selecting thickness of foil as presented in Section 3, the resistance of each winding will be minimized.

The design criteria of decreasing stray capacitance are to set the number layers with the limits of window area of the magnetic core and a coupling factor \( \leq 0.2 \) must be selected.

5. Experimental Results

Two different experiments were performed. The objective of the first experiment was to validate the effects of the number of windings on \( C_{\text{stray}} \) for a multi-winding transformer with an integrated rectifier and compare it with a multi-layer transformer. The second experiment was performed to operate an LLC resonant converter designed with an FHA for high-voltage applications, and to compare multi-layer high-voltage transformers and multi-winding transformers with integrated rectifiers. The prototypes were designed with the same geometry, using a voltage doubler.

The frequency response was measured using a vector network analyzer Bode100 for all the experiments under the two conditions. The measurements were first performed for the parallel resonance frequency, and then for the open-circuit inductance from the primary side. Then, \( C_{\text{stray}} \) was calculated using Equation (19).

5.1. Comparison \( C_{\text{stray}} \) in High-Voltage Transformer

Two transformers were designed with winding arrangements as Figure 12. The TR1 transformer is called a multi-layer transformer, and all the layers are connected in series before rectification. The TR2 transformer is called a multi-winding transformer with an integrated rectifier, and all adjacent layers are connected in series after the rectification circuit. The connection diagrams and prototypes for TR1 and TR2 are shown in Figure 11. The transformer prototype parameters are listed in Table 1.
Table 1. Transformer prototype parameters.

<table>
<thead>
<tr>
<th></th>
<th>Np</th>
<th>Ns-Layer</th>
<th>Zs-Layer</th>
<th>ts</th>
<th>tp</th>
<th>Air Gap</th>
<th>hn</th>
<th>hs</th>
</tr>
</thead>
<tbody>
<tr>
<td>TR1</td>
<td>24 Turns</td>
<td>24 Turns</td>
<td>20 Layer</td>
<td>0.035 mm</td>
<td>0.1 mm</td>
<td>1 mm</td>
<td>5 mm</td>
<td>1 mm</td>
</tr>
<tr>
<td>TR2</td>
<td>24 Turns</td>
<td>24 Turns</td>
<td>20 Layer</td>
<td>0.035 mm</td>
<td>0.1 mm</td>
<td>1 mm</td>
<td>5 mm</td>
<td>1 mm</td>
</tr>
</tbody>
</table>

All measurements and calculations were performed by increasing the number of layers from one to 20 for TR1 with and without the rectifier connected, and T2, respectively. \( C_{ir} \) and \( C_{stray} \) calculations are shown in Table 2.

Consequently, \( C_{ir} \) and \( C_{stray} \) increase exponentially for TR1, and \( C_{stray} \) increases proportionally with the number of layers \( z \) for TR2. The effect of \( C_{ir} \) significantly increased \( C_{stray} \) for TR1. \( C_{ir} \) and \( C_{stray} \) measurements, calculations and frequency response measurements (for \( z = 20 \)) for TR1 and TR2 are presented in Figure 13. The frequency response measurements and calculated \( C_{ir} \) and \( C_{stray} \) values are presented in Table 3.
Table 2. Capacitance calculations and measurements for TR1 and TR2.

<table>
<thead>
<tr>
<th>Layer</th>
<th>Multi-Layer Transformer (TR1)</th>
<th>Multi-Winding Transformer (TR2)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Measurement $C_{tr}$ (nF)</td>
<td>Measurement $C_{stray}$ (nF)</td>
</tr>
<tr>
<td>1</td>
<td>0.111</td>
<td>0.147</td>
</tr>
<tr>
<td>2</td>
<td>0.278</td>
<td>0.413</td>
</tr>
<tr>
<td>3</td>
<td>0.435</td>
<td>0.757</td>
</tr>
<tr>
<td>4</td>
<td>0.620</td>
<td>1.098</td>
</tr>
<tr>
<td>5</td>
<td>0.864</td>
<td>1.601</td>
</tr>
<tr>
<td>6</td>
<td>1.117</td>
<td>2.160</td>
</tr>
<tr>
<td>7</td>
<td>1.283</td>
<td>2.503</td>
</tr>
<tr>
<td>8</td>
<td>1.591</td>
<td>3.251</td>
</tr>
<tr>
<td>9</td>
<td>1.890</td>
<td>3.947</td>
</tr>
<tr>
<td>10</td>
<td>2.072</td>
<td>4.618</td>
</tr>
<tr>
<td>11</td>
<td>2.415</td>
<td>5.394</td>
</tr>
<tr>
<td>12</td>
<td>2.673</td>
<td>5.962</td>
</tr>
<tr>
<td>13</td>
<td>2.944</td>
<td>6.873</td>
</tr>
<tr>
<td>14</td>
<td>3.299</td>
<td>7.755</td>
</tr>
<tr>
<td>15</td>
<td>3.609</td>
<td>8.770</td>
</tr>
<tr>
<td>16</td>
<td>3.880</td>
<td>9.303</td>
</tr>
<tr>
<td>17</td>
<td>4.384</td>
<td>10.452</td>
</tr>
<tr>
<td>18</td>
<td>4.750</td>
<td>11.467</td>
</tr>
<tr>
<td>19</td>
<td>5.216</td>
<td>12.554</td>
</tr>
<tr>
<td>20</td>
<td>5.612</td>
<td>13.540</td>
</tr>
</tbody>
</table>

Figure 13. Capacitance measurement for TR1 and TR2: (a) The comparison of $C_{tr}$ and $C_{stray}$ by layers (blue—TR1 $C_{stray}$, black—TR1 $C_{tr}$, red—TR2 $C_{stray}$); (b) The comparison of calculated and measured $C_{stray}$ for TR2 (red—calculated, black—measured); (c) The frequency response measurement of TR1 and TR2 for 20 layers: (dashed—TR1 $C_{stray}$, dotted—TR1 $C_{stray}$, solid—TR2).
Table 3. Frequency response measurements for prototypes.

<table>
<thead>
<tr>
<th></th>
<th>TR1</th>
<th>TR1</th>
<th>TR2</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_{paralel}$</td>
<td>120.12 kHz</td>
<td>78.96 kHZ</td>
<td>316.9 kHz</td>
</tr>
<tr>
<td>$C_{tr}$</td>
<td>5.94 nF</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$C_{stray}$</td>
<td>13.7 nF</td>
<td>0.832 nF</td>
<td></td>
</tr>
</tbody>
</table>

The measurements demonstrated that the impact of $C_{stray}$ on high-voltage applications is significantly reduced using the proposed method. Therefore, by using the proposed transformer prototype TR2, the LLC resonant converter is suitable for high-voltage applications.

5.2. System Aspect LLC Resonant Converter Application

LLC resonant circuits were designed using the FHA methodology and time domain analysis where $C_{stray}$ was not considered. An experiment was performed to compare the LLC resonant circuits for TR1 and TR2, and all layers were installed on the transformers. The design parameters of the LLC resonant circuit are listed in Table 4, and the test setup is illustrated in Figure 14.

Table 4. LLC resonant converter parameters.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_s$</td>
<td>360 V</td>
</tr>
<tr>
<td>$V_{out}$</td>
<td>12,000 V</td>
</tr>
<tr>
<td>$P_{out}$</td>
<td>1200 W</td>
</tr>
<tr>
<td>$L_r$</td>
<td>78.4 $\mu$H</td>
</tr>
<tr>
<td>$C_r$</td>
<td>66 nF</td>
</tr>
<tr>
<td>$L_m$</td>
<td>286 $\mu$H</td>
</tr>
<tr>
<td>$f_{res}$</td>
<td>70 kHz</td>
</tr>
<tr>
<td>$f_{SW}$</td>
<td>45–150 kHz</td>
</tr>
</tbody>
</table>

Figure 14 shows the measurement results of the LLC resonant circuit for TR1. The measurements show that a high $C_{stray}$ causes large distortions in the current and voltage waveforms for the entire operating frequency range. Distortions are expected under light-load conditions because of oscillations between $C_{stray}$ and $L_r$. However, a fairly high distortion was observed at the resonant frequency. In addition, after the switching
frequency reached 1.5 times the resonant frequency, the resonant tank exited the inductive region, which is not suitable for the LLC resonant circuit and primary mosfets. Therefore, waveform disturbances were extremely high.

![Figure 15](image1.png)

**Figure 15.** The LLC resonant converter with TR1 (1) $V_{AB}$ as blue, (2) $I_R$ as purple, (3) $V_{OUT}$ as turquoise: (a) full load at the resonant frequency $f_{SW} = 70$ kHz; (b) light load at below resonant frequency $f_{SW} = 45$ kHz; (c) light load at the above resonant frequency $f_{SW} = 150$ kHz (ZVS loss); (d) open circuit at the above resonant frequency $f_{SW} = 120$ kHz ZVS loss.

The LLC resonant circuit measurement at resonant frequency, at below resonant frequency and above resonant frequency results for TR2 are shown in Figures 16–18, respectively. The measurements show that the LLC resonant converter operates by remaining in the inductive region under all load conditions and operating frequency ranges in compliance with the FHA. In addition, the LLC resonant tank progressed in the inductive region after the switching frequency reached twice the resonant frequency.

![Figure 16](image2.png)

**Figure 16.** LLC resonant converter improvement for primary-side switches at the resonant frequency $f_{SW} = 70$ kHz (1) $V_{AB}$ as blue, (2) $I_R$ as purple, (3) $V_{OUT}$ as turquoise: (a) full load; (b) heavy load; (c) light load; (d) open circuit.
Moreover, there is no disturbance of ZVS on the main switches in the secondary open-circuit condition.

![Figure 17](image1.png)

**Figure 17.** LLC resonant converter improvement for primary-side switches at below resonant frequency $f_{SW} = 45$ kHz: (a) light load $Q = 0.2$; (b) open circuit; (c) heavy load $Q = 0.8$.

![Figure 18](image2.png)

**Figure 18.** LLC resonant converter improvement for primary-side switches at above resonant frequency $f_{SW} = 150$ kHz: (1) $V_{AB}$ as blue; (2) $I_R$ as purple; (3) $V_{OUT}$ as turquoise; (a) full load; (b) light load $Q = 0.2$; (c) open circuit.

It is not sufficient for the designed LLC resonant circuit to operate only in the inductive region, within the operating range. However, all main switches must be provided to operate under soft-switching conditions. The measurements in Figure 19 show that the main switches operate under soft-switching conditions in the LLC resonant circuit for the proposed transformer TR2, the multi-output transformer with an integrated rectifier.
Moreover, there is no disturbance of ZVS on the main switches in the secondary open-circuit condition.

The LLC resonant circuit operates only in the inductive region and the main switches operate in the ZVS condition. However, as defined in Section 3, in multi-winding transformers coupled inductance and coupling coefficient \( k \) affect power transfer in secondary windings. \( L \) matrix measurements and coupling coefficient calculations for adjacent windings for proposed winding arrangements are presented in Appendix A. The maximum value of coupling coefficient for primary to secondary windings is 0.03 and between adjacent secondary windings is 0.22. The output voltage distribution (OVD), voltage distribution error (VDE) and secondary winding currents must be evaluated to be certain that power transfer is very close for all secondary windings. The secondary winding waveform is presented in Figure 20. OVD and VDE measurements are presented in Table 5.
Figure 20. TR2 current measurements: (a) $I_{pri}$ as green (4); $I_{sec1}$ as gray (R1); $I_{sec20}$ as purple (3); (b) $I_{sec1}$ as gray (R1); $I_{sec20}$ as gray (R2); $I_{sec10}$ as purple (3); (c) $I_{sec1}$ as gray (R1); $I_{sec20}$ as gray (R2); $I_{sec4}$ as purple (3).

Table 5. OVD, VDE measurements and calculations.

<table>
<thead>
<tr>
<th>Layer</th>
<th>OVD (V)</th>
<th>VDE (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>650.94</td>
<td>4.32%</td>
</tr>
<tr>
<td>2</td>
<td>1293.52</td>
<td>1.53%</td>
</tr>
<tr>
<td>3</td>
<td>1927.36</td>
<td>0.57%</td>
</tr>
<tr>
<td>4</td>
<td>2554.36</td>
<td>0.16%</td>
</tr>
<tr>
<td>5</td>
<td>3174.90</td>
<td>−0.07%</td>
</tr>
<tr>
<td>6</td>
<td>3796.58</td>
<td>−0.03%</td>
</tr>
<tr>
<td>7</td>
<td>4420.16</td>
<td>0.02%</td>
</tr>
<tr>
<td>8</td>
<td>5039.18</td>
<td>−0.08%</td>
</tr>
<tr>
<td>9</td>
<td>5666.56</td>
<td>0.08%</td>
</tr>
<tr>
<td>10</td>
<td>6292.80</td>
<td>0.05%</td>
</tr>
<tr>
<td>11</td>
<td>6897.38</td>
<td>−0.26%</td>
</tr>
<tr>
<td>12</td>
<td>7535.40</td>
<td>0.20%</td>
</tr>
<tr>
<td>13</td>
<td>8129.34</td>
<td>−0.36%</td>
</tr>
<tr>
<td>14</td>
<td>8765.08</td>
<td>0.15%</td>
</tr>
<tr>
<td>15</td>
<td>9356.74</td>
<td>−0.33%</td>
</tr>
<tr>
<td>16</td>
<td>9990.58</td>
<td>0.11%</td>
</tr>
<tr>
<td>17</td>
<td>10,573.50</td>
<td>−0.38%</td>
</tr>
<tr>
<td>18</td>
<td>11,216.46</td>
<td>0.18%</td>
</tr>
<tr>
<td>19</td>
<td>11,800.52</td>
<td>−0.33%</td>
</tr>
<tr>
<td>20</td>
<td>12,456.40</td>
<td>0.27%</td>
</tr>
</tbody>
</table>

There is quite a low phase shift inspected at $I_{sec1}$, $I_{sec10}$, and $I_{sec20}$ but it was in acceptable range. However, maximum disturbance found at $I_{sec4}$ can be described as winding-position-related leakage–flux coupling. VDE was less than 5% and in the acceptable range. The experimental results show that:

- Proposed winding arrangement effects to coupling factor for secondary windings is limited;
• Since coupling factor effects are limited, $C_{\text{stray}}$ increases proportionally with the number of layers $z$ where all secondary windings are designed in the same geometry;
• $C_{\text{stray}}$ on high-voltage applications is significantly reduced using the proposed multi-winding transformer;
• LLC resonant converters are suitable for use in high-voltage applications where $C_{\text{stray}}$ on high-voltage applications is significantly reduced.

6. Conclusions

This paper presents the problem of high-voltage application of an LLC resonant converter because of transformer and rectifier network capacitances, and proposes a model circuit diagram for a transformer that will reduce parasitic capacitances to improve LLC converter reliability for high-voltage applications. The disruptive effect of the stray capacitance is shown in the LLC resonant circuit for high-voltage applications. To overcome these problems, a winding arrangement multi-winding transformer with an integrated rectifier scheme is proposed that discriminates over the electric field on the transformer to reduce parasitic capacitance values regardless of the output voltage, leakage inductance and winding loss. A comparative analysis and experimental studies were performed to validate the winding scheme, winding arrangement, model, and resonance circuit.

The results of this study show that the simplified model of a multi-winding transformer with an integrated rectifier agrees with the measurement results, and the proposed multi-output transformer with an integrated rectifier and LLC resonant converter is reliable for high-voltage applications because it minimizes parasitic capacitances and regulates the output voltage even under any load condition in the operating region.

Author Contributions: Conceptualization, A.B.; methodology, U.O.; validation, U.O.; investigation, U.O.; writing—original draft, U.O.; supervision, A.B. All authors have read and agreed to the published version of the manuscript.

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Data Availability Statement: Not applicable.

Conflicts of Interest: The authors declare no conflict of interest.

Abbreviations
The following abbreviations are used in this research:

- AC: Alternative Current
- DC: Direct Current
- $L_m$: Mutual Inductance
- $L_{m\text{core}}$: Mutual Inductance of Core
- $L_{ms}$: Mutual Inductance of Secondary Windings
- $L_{lk}$: Equivalent Leakage Inductance
- $L_{lkp}$: Primary Winding Leakage Inductance
- $L_{lks}$: Secondary Winding Leakage Inductance
- $L_r$: Serial Resonant Inductance
- $L_n$: Inductance Ratio of $L_m$ to $L_r$
- $C_r$: Resonant Capacitance
- $C_{\text{in1}}, C_{\text{in2}}$: Primary Switch Parasitic Capacitances
- $C_{\text{stray}}$: Stray Capacitance
- $C_{\text{tr}}$: Equivalent Transformer Capacitance
- $C_n$: Ratio of $C_{\text{stray}}$ to $C_r$
- $C_p$: Primary Winding Capacitance
- $C'_{p}$: Equivalent Primary Winding Capacitance Referred to Primary Side
- $C_s$: Secondary Winding Capacitance Referred to Primary Side
- $C'_{s}$: Equivalent Secondary Winding Capacitance Referred to Primary Side
- $C_{p-s}$: Primary-Secondary Coupling Capacitance
- $C_j$: Rectifier Network Junction Capacitance
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$C_j'$ Rectifier Network Junction Capacitance Referred to Primary
$C_{s-s}$ Secondary-Secondary Coupling Capacitance for Each Secondary Windings
$C_{ac}$ AC Capacitance of Storage Energy in Electrical Field For a Transformer
$C_{dc}$ DC Capacitance of Storage Energy in Electrical Field For a Transformer
$f_{RES}$ Series Resonant Frequency of Converter
$f_p$ Pole Resonant Frequency of Converter
$f_{SW}$ Switching Frequency of Converter
$f_n$ Normalized Frequency
$f_{series}$ Series Resonance Frequencies of Transformer,
$f_{parallel}$ Parallel Resonance Frequencies of Transformer,
w Fundamental Frequency
$R_o$ Load Resistance
$R_{eq}$ Equivalent Load Resistance
$R_{ac}$ AC Resistance of A Winding
$R_{dc}$ DC Resistance of A Winding
$R_s$ Secondary Winding Resistance
$R_p$ Primary Winding Resistance
$R_{gap}$ Resistance of Close Proximity Air Gap
$R_{tot}$ Sum of $R_p$ And $R_{ps}$
$f_R$ The Ratio of $R_{ac}$ to $R_{dc}$
$\Delta$ The Ratio of $t$ to $\Delta$
$\Delta_{opt}$ Optimum Value of $\Delta$
$t$ Foil Thickness
$t_p$ Foil Thickness of Primary Windings
$t_s$ Foil Thickness of Secondary Windings
$\delta$ Skin Depth
$p_{th}$ Number of Layers for Foil Conductors
$h$ Insulator Thickness
$h_{p-s}$ Insulator Thickness Between Primary and Secondary
$h_n$ Insulator Thickness Between Secondary Winding To Airgap
$b_w$ Width of Layer for a Winding
$Q$ Quality Factor
$G$ Transfer Ratio
$N$ Turn Ratio
$N_s$ Number of Turns for Each Secondary Windings
$N_p$ Number of Turns for Primary Winding
$I_{rms}$ Rms-Value of The Converter Current
$I'_{rms}$ Rms-Value of The First Derivative Of Converter Current
$L$ Inductance Matrix for Multi-Winding Transformer
$i$ Output Current for Each Secondary Winding
$v$ Output Voltage for Each Secondary Winding
$k_{i-j}$ Coupling Coefficient
$M_{i-j}$ Coupling Inductance for Each Windings
$L_{ii}$ Self Inductance for Each Windings
OVD Output Voltage Distribution
VDE Voltage Distribution Error
Appendix A

Table A1. L Matrix for TR2.

<table>
<thead>
<tr>
<th>k_{p-s1}</th>
<th>k_{p-s11}</th>
<th>k_{s1-s2}</th>
<th>k_{s11-s12}</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.10</td>
<td>0.90</td>
<td>0.030</td>
<td>0.208</td>
</tr>
<tr>
<td>0.025</td>
<td>0.25</td>
<td>0.027</td>
<td>0.220</td>
</tr>
<tr>
<td>0.040</td>
<td>0.40</td>
<td>0.030</td>
<td>0.207</td>
</tr>
<tr>
<td>0.007</td>
<td>0.07</td>
<td>0.040</td>
<td>0.174</td>
</tr>
<tr>
<td>0.028</td>
<td>0.28</td>
<td>0.024</td>
<td>0.141</td>
</tr>
<tr>
<td>0.027</td>
<td>0.27</td>
<td>0.019</td>
<td>0.148</td>
</tr>
<tr>
<td>0.022</td>
<td>0.22</td>
<td>0.023</td>
<td>0.145</td>
</tr>
<tr>
<td>0.024</td>
<td>0.24</td>
<td>0.022</td>
<td>0.120</td>
</tr>
<tr>
<td>0.026</td>
<td>0.26</td>
<td>0.016</td>
<td>0.120</td>
</tr>
<tr>
<td>0.017</td>
<td>0.17</td>
<td>0.015</td>
<td>0.014</td>
</tr>
</tbody>
</table>

Table A2. Cross Coupling Coefficient (k) for Adjacent Layers for TR2.

Cross Coupling Coefficient (k) for Adjacent Layers for TR2

<table>
<thead>
<tr>
<th>k_{p-s1}</th>
<th>k_{p-s11}</th>
<th>k_{s1-s2}</th>
<th>k_{s11-s12}</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.10</td>
<td>0.90</td>
<td>0.030</td>
<td>0.208</td>
</tr>
<tr>
<td>0.025</td>
<td>0.25</td>
<td>0.027</td>
<td>0.220</td>
</tr>
<tr>
<td>0.040</td>
<td>0.40</td>
<td>0.030</td>
<td>0.207</td>
</tr>
<tr>
<td>0.007</td>
<td>0.07</td>
<td>0.040</td>
<td>0.174</td>
</tr>
<tr>
<td>0.028</td>
<td>0.28</td>
<td>0.024</td>
<td>0.141</td>
</tr>
<tr>
<td>0.027</td>
<td>0.27</td>
<td>0.019</td>
<td>0.148</td>
</tr>
<tr>
<td>0.022</td>
<td>0.22</td>
<td>0.023</td>
<td>0.145</td>
</tr>
<tr>
<td>0.024</td>
<td>0.24</td>
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<td>0.120</td>
</tr>
<tr>
<td>0.026</td>
<td>0.26</td>
<td>0.016</td>
<td>0.120</td>
</tr>
<tr>
<td>0.017</td>
<td>0.17</td>
<td>0.015</td>
<td>0.014</td>
</tr>
</tbody>
</table>

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1. Li, F.; Hao, R.; Lei, H.; Zhang, X.; You, X. The Influence of Parasitic Components on LLC Resonant Converter. *Energies* 2019, 12, 4305. [CrossRef]


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