Design Technique of High Power Efficiency LLC DC-DC Converters for Photovoltaic Cells

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Abstract: This paper proposes a design technique of high power efficiency LLC DC-DC Converters for Photovoltaic Cells. The secondary side circuit and transformer fabrication of proposed circuit are optimized for overcoming the disadvantage of limited input voltage range and, realizing high power efficiency over a wide load range of LLC DC-DC converters. The optimized technique is described with theoretically and with simulation results. Some experimental results have been obtained with the prototype circuit designed for the 80 - 400 V input voltage range. The maximum power efficiency is 98%.

Keywords: Power Efficiency Improvement, Photovoltaic Generation System, Half Bridge, LLC Resonance, Eddy Current, Litz Wire

1. Introduction

The introduction of renewable energy has been promoted as a substitute for fossil fuels recently. Solar energy has attracted attention as an energy source capable of reducing carbon dioxide emissions. Installations of photovoltaic generation systems in private residences and commercial buildings and on idle land are becoming increasingly common. When solar cells are used, they are connected to the photovoltaic system via a power conditioner, as shown in Figure 1, which is combined with a DC-DC converter and a DC-AC inverter. When a power conditioner is connected to AC power line, the electrical isolation is necessary for safety. For example, the DC-AC inverter in a later stage can be isolated using a low-frequency transformer; however, this approach is not practical because of the volume and weight of the transformer.

LLC resonant DC-DC converters have been incorporated into power conditioners. These converters are composed of relatively few parts, resulting in a compact, lightweight system with high power efficiency and ultra-low noise. Also, these converters are composed of a high-frequency transformer, which is drastically lighter than a low-frequency transformer.

This converter adopts a half-bridge LLC resonant DC-DC converter that aggressively uses the transformer leakage inductance as a resonant inductance. The magnetic flux corresponding to the leakage inductance crosses the primary and secondary winding wire, which causes eddy current losses in copper wire. A technique [1] involving the use of a magnetic plated wire as the winding...
wire has been proposed as a loss reduction method. However, this approach increases costs. Another method [2] is adding a core to the transformer to avoid leakage flux crossing to the winding wire; however, such a transformer structure is more complicated.

Therefore, in this paper, a practical loss reduction method based on optimization of the strand diameter of the litz wire is proposed. Furthermore, to improve power efficiency and reduce losses due to reverse current, a voltage doubler rectifier circuit [6] is adopted in the secondary side instead of the commonly used center-tap rectifier circuit [3] (pp. 4-11), [4], [5].

2. LLC Resonant Circuits

2.1. Circuit Configurations

Figure 2 shows the equivalent circuit of the half-bridge LLC resonant DC-DC converter. In Figure 2, transformer T is represented in an equivalent circuit. \(L_m\) is the magnetizing inductance, \(L_{r1}\) is the leakage inductance of the primary side, and \(L_{r2}\) is the leakage inductance of the secondary side that was converted to the primary side. These parameters are used as a resonant inductance. \(L_r\) is the self-inductance of the primary side. Here,

\[ L_{r1} = L_{r2} = L_r, \]

Q₁ and Q₂ are MOSFET switches. \(C_c\) and \(C_v\) are current and voltage resonance capacitors, respectively. The voltage doubler rectifier circuit consists of diodes D₁ and D₂ and capacitor \(C_d\) and \(C_o\). The MOSFET switches Q₁ and Q₂ alternately repeat ON and OFF. An interval referred to as the “dead time” occurs during the ON time of Q₁ and Q₂.

In the case of center-tap rectifier circuits, the reverse voltage applied to the rectifying diode is two times the output voltage; the ringing voltage is also applied. In contrast, in the case of the voltage doubler rectifier circuit, the reverse voltage applied to the rectifying diode is equal to the output voltage.
Therefore, the loss due to the reverse current in the voltage doubler rectifier circuit can be reduced compared with that in the center-tap rectifier circuit.

![LLC resonant half-bridge DC-DC converter](image)

**Figure 2.** LLC resonant half-bridge DC-DC converter

<table>
<thead>
<tr>
<th>Location</th>
<th>FHA Analysis</th>
<th>PSIM Simulation</th>
<th>Experiment</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{in}$</td>
<td>270V</td>
<td>270V</td>
<td>270V</td>
</tr>
<tr>
<td>$L_P$</td>
<td>64.7 μH</td>
<td>64.7 μH</td>
<td>64.7 μH</td>
</tr>
<tr>
<td>$N_P$</td>
<td>12T</td>
<td>12T</td>
<td>12T</td>
</tr>
<tr>
<td>$N_S$</td>
<td>17T</td>
<td>17T</td>
<td>17T</td>
</tr>
<tr>
<td>$L_r$</td>
<td>8.86 μH</td>
<td>8.86 μH</td>
<td>8.86 μH</td>
</tr>
<tr>
<td>$k$</td>
<td>0.863</td>
<td>0.863</td>
<td>0.863</td>
</tr>
<tr>
<td>$C_r$</td>
<td>1.56 μH</td>
<td>1.56 μH</td>
<td>1.56 μH</td>
</tr>
<tr>
<td>$C_d$</td>
<td>1.12 μF</td>
<td>1.12 μF</td>
<td>1.12 μF</td>
</tr>
<tr>
<td>$R_L$</td>
<td>67.2 Ω</td>
<td>67.2 Ω</td>
<td>67.2 Ω</td>
</tr>
<tr>
<td>$C_v$</td>
<td>Unimplemented</td>
<td>0.036 μF</td>
<td>0.036 μF</td>
</tr>
<tr>
<td>$R_{DC} (N_P)$</td>
<td>0 mΩ</td>
<td>9.19 mΩ</td>
<td>9.19 mΩ</td>
</tr>
<tr>
<td>$R_{DC} (N_S)$</td>
<td>0 mΩ</td>
<td>31.2 mΩ</td>
<td>31.2 mΩ</td>
</tr>
<tr>
<td>$R_{ON} (Q_1,Q_2)$</td>
<td>0 mΩ</td>
<td>30 mΩ</td>
<td>About 30 mΩ</td>
</tr>
<tr>
<td>$V_{F of Diode} (Q_1,Q_2)$</td>
<td>0 V</td>
<td>1 V</td>
<td>About 1 V</td>
</tr>
<tr>
<td>$V_{F} (D_1,D_2)$</td>
<td>0 V</td>
<td>1 V</td>
<td>About 1 V</td>
</tr>
</tbody>
</table>

1 The coupling coefficient of the transformer

2.2. Circuit Analysis

Using first harmonic approximation (FHA) method to analyze the relationship between switching frequency and output voltage $V_o$. The load resistance $R_L$ was 67.2Ω because the rated output
power was 2,150 W when the rated output voltage was 380 V. The primary and secondary winding DC resistance was set as 0 mΩ, and the voltage resonance capacitor was assumed to be unimplemented. The other constants are listed in Table 1. All the elements were placed on the primary side for simplification. Figure 3 shows equivalent circuits for the analysis by FHA. In Figure 3, \( R_{ec} \) is the equivalent AC resistance of the secondary circuits including the load resistance. FHA was performed by approximation analysis using the first term obtained by expanding the square-wave voltage applied to the resonant circuit to Fourier series as follows.

The fundamental harmonic element \( e_{\text{in1}} \) of the square wave \( e_{\text{in}} \) in Figure 2 is represented by the following equation:

\[
e_{\text{in1}} = \frac{2V_{\text{in}}}{\pi} \sin(\omega t),
\]

(2)

Similarly, the fundamental harmonic element \( e_{\text{ot}} \) that replaced square wave \( e_{\text{o}} \) to the primary side becomes Eq. (3).

\[
e_{\text{ot}} = \frac{2nV_{\text{o}}}{\pi} \sin(\omega t),
\]

(3)

From Eq. (2) and (3), the voltage conversion rate \( M \) is represented as Eq. (4).

\[
M = \frac{nV_{\text{o}}}{V_{\text{in}}}
\]

(4)

The power loss of the equivalent AC resistance is equal to that of the load resistance \( R_{L} \).

\[
\frac{V_{\text{o}}^2}{R_{L}} = \left( \frac{2nV_{\text{o}}}{\pi R_{ac}} \right)^2
\]

(5)

Rearrangement of Eq. (5) gives Eq. (6).

\[
R_{ac} = \frac{2n^2R_{L}}{\pi^2}
\]

(6)

Using the circuit in Figure 3, the voltage conversion ratio \( M \) is obtained from Eq. (7). However, the equation from (8) to (17) shows each constant and the definitions of the variables.

\[
M = \frac{kF^3}{\sqrt{\left[ F(F^2 - \omega^2_{\text{ra}})\right]^2 + \left[ \frac{1}{2} \omega^2_{\text{ra}}(F^2 - 1)(F^2 - \omega^2_{\text{rb}})\right]^2}}
\]

(7)

\[
n = \frac{N_p}{N_s}
\]

(8)

\[
C = \frac{C_d}{n^2}
\]

(9)

\[
C_t = \frac{C_{dr}}{C + C_{fr}}
\]

(10)

\[
\omega_{r1} = \frac{1}{\sqrt{2}} \sqrt{\frac{1}{\frac{1}{\omega_{\text{ra}}}^2 + \frac{1}{C_t L_m}}} \left(1 + \sqrt{1 - \frac{4C_t^2}{CC_{fr}}(1 - k^2)}\right)
\]

(11)

\[
\omega_{r2} = \frac{1}{\sqrt{2}} \sqrt{\frac{1}{\frac{1}{\omega_{\text{ra}}}^2 + \frac{1}{C_t L_m}}} \left(1 - \sqrt{1 - \frac{4C_t^2}{CC_{fr}}(1 - k^2)}\right)
\]

(12)

\[
\omega_{\text{ra}} = \frac{\omega_{\text{ra}}}{\omega_{r1}}
\]

(13)
\[ \omega_{rb} = \frac{\omega_{r2}}{\omega_{r1}} \]  
(14)

\[ \omega_{r0} = \frac{1}{\sqrt{L_1C_0}} \]  
(15)

\[ F = \frac{\omega}{\omega_{r1}} \]  
(16)

\[ Q = \frac{\omega_{r1}L_m}{R_{ac}} \]  
(17)

The output voltage \( V_o \) is calculated as follows using Eq. (4) and (7):

\[ V_o = \frac{M V_{in}}{n} \]  
(18)

\[ \text{Figure 3. The equivalent circuit of LLC resonant half-bridge DC-DC converter for the FHA analysis} \]

2.3. Circuit Analysis Results

Figure 4 shows the relationship between the switching frequency and output voltage, with the results of FHA analysis, circuit simulation with PSIM, and experimentation.

Analysis Conditions

In PSIM and experiments, the DC resistance of the primary and secondary windings was 9.19 mΩ and 31.2 mΩ, respectively. The constitution of the litz wire in the experiments was 0.03 \( \varphi \) of 3900 on the primary side and 0.03 \( \varphi \) of 1440 on the secondary side. However, the litz wire is so thin that it tends to melt. Therefore, a minimum litz wire strand diameter of 0.03 \( \varphi \) was used here. In Table 1, FHA analysis, PSIM simulation and experimental conditions are summarized.

2.4. Operating Mode

The behavior of LLC resonant circuit is divided into the area I, II, and III, as shown in Figure 4. Figure 5 shows drain current waveform and a drain-source voltage waveform of \( Q_1 \) of the respective areas. The switching state of areas I and II are resonating in zero voltage switching (ZVS) mode. Leakage inductance is resonating with the resonant capacitor \( C_t \) in area I (Eq. (10)). Area II is separated into two terms. In the first term \( t_1 \), the leakage inductance is resonating with the resonant capacitor \( C_t \). In the second term \( t_2 \), the primary self-inductance is resonating with the current resonance capacitor \( C_r \). The input power is not transmitted to the secondary side in the second term \( t_2 \). Figure 5 (b) shows the waveform of the boundary condition between area I and II. The current waveform flowing through the primary and secondary windings becomes sine wave. The switching state
of area III is resonating in the zero-current switching (ZCS) mode. Commonly, the use of area III is avoided because there is in hard switching area.

Figure 4. The relationship between switching frequency and output voltage

(a)  
(b)  
(c)  
(d) The dotted line: Hard Switching
2.5. Circuit Analysis Result Discussion

2.5.1. Comparison of Experimental and Simulation (PSIM) Results

The experiments were conducted up to 400 V by the rated output power. The experimental results were approximately equal to the simulation results.

2.5.2. Comparison of Experimental and FHA analysis Results

LLC resonant circuit is non-linear; however, FHA is an approximate analysis of resonant angular frequency \( \omega_{\text{r1}} \). Therefore, it can be performed linear analysis of LLC resonant circuit in this frequency region.

3. Efficiency Improvements through the Use of Thin Strand Diameter Litz Wire

3.1. Electromagnetic Analysis

As shown in previously, LLC resonant circuit is composed of the transformer with weak coupling to make leakage inductance as shown in Figure 6. The structure leads the occurrence of the magnetic leakage in/around the transformer. In this section, to analyze the effect of the leakage in the transformer, the simulation results of the leakage magnetic flux distribution in the transformer and the current density in the windings were shown. The simulation was performed with electromagnetic simulator JMAG. The objective of this simulations is to analyze the loss increasing with eddy current.

LLC resonant circuit uses the leakage inductance for resonant operation. The AC loss of windings is considered with the following three factors; (i) Leakage magnetic flux crosses in the windings, inducing eddy currents and thereby resulting in losses. (ii) The loss is occurred due to the skin effect. (iii) The loss is occurred due to the proximity effect between the windings. The maximum switching frequency of the DC-DC converter is 120 kHz, and the depth of the epidermis is 0.2 mm. Thus, skin effects were ignored because the wire diameter of the litz wire used here is less than 0.1 mm. Losses due to the proximity effect are ignored because the frequency is less than 100 kHz and the copper wire diameter is 0.1φ [7], [8] (pp. 368-375), [9] (pp. 2229-2235), [10] (pp. 738-745). The simulation was done with two-dimensional simulation. The simulation area is plane PQRS shown in Figure 6. The result is considered as equivalent for each plane rotated around center line C. PQ71 and PC95 cores manufactured by TDK were used, and the center gap was 1.5 mm. However, the litz wire windings were too many to analyze. Consequently, the number of litz wire windings was reduced and the larger strand diameter litz wire was used in the magnetism simulation. The configuration of the litz wire according to the analysis results were as follows;

(i) Strand diameter of litz wire: 0.5φ for both primary and secondary sides.
(ii) A number of litz wire winding: 38 lines on the primary side; 15 lines on the secondary side.

Simulation conditions:

- Input Voltage: 270 V
Input Current: 115 A
Operating Frequency: 39.7 kHz

The value of input current 115 A is set as the assuming the peak current value of the transient state of LLC resonant circuit. The configuration of the litz wire is listed in Table 2.

**Figure 6.** The transformer with primary / secondary division bobbin

**Table 2.** Litz wire structure (experiment)

<table>
<thead>
<tr>
<th>Strand diameter of the litz wire</th>
<th>Structure of the litz wire</th>
<th>DC Resistance</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Primary Winding</td>
<td>Secondary Winding</td>
</tr>
<tr>
<td></td>
<td>(Sectional Area)</td>
<td>(Sectional Area)</td>
</tr>
<tr>
<td>0.03 φ</td>
<td>3900 (2.76 mm²)</td>
<td>1440 (1.02 mm²)</td>
</tr>
<tr>
<td>0.06 φ</td>
<td>990 (2.80 mm²)</td>
<td>360 (1.02 mm²)</td>
</tr>
<tr>
<td>0.1 φ</td>
<td>351 (2.76 mm²)</td>
<td>130 (1.02 mm²)</td>
</tr>
</tbody>
</table>
Note: The arrow in the figure shows the orientation of the magnetic flux.

**Figure 7.** Magnetic flux density distribution and direction in the transformer

3.2. *Analysis Results by Electromagnetic Simulator JMAG*

Figure 7 shows the analysis results by electromagnetic simulator JMAG, whereas Figure 8 and 9 show the magnetic flux density distribution and the current density distribution in winding wire, respectively.

**Figure 8.** Magnetic flux density distribution in winding wire

**Figure 9.** Current density distribution in primary winding wire
3.3. Consideration of the Analysis Results by Electromagnetic Simulator JMAG

Figure 9 shows the current density distribution in the primary winding wire. Thus, the same winding is in the section of A and B in Figure 9; the results show that the current density in Sec. B is apparently higher than that in Sec. A. Figure 8 shows the magnetic flux density distribution in the primary winding. According to Figure 7 and 8, the magnetic flux density of B is higher than A. These results suggest that eddy currents occur in the winding wire because of leakage flux. The same consideration can be applied to the secondary winding.

Figure 10. Analysis model of eddy current loss (a) Analysis model of eddy current loss; (b) Cylindrical- coordinate model

3.4. Consideration of loss reduction by eddy current

3.4.1. Theoretical Analysis of Eddy Current Loss

For leakage flux crossing into the winding wire, theoretical analysis [11] using the simple model in Figure 10 indicates that eddy current losses occur in the winding wire. By these results, we analyzed the factors that influence these losses and examined techniques for reducing them.

The model is shown in Figure 10 was used for the analysis. This model consists of a single copper wire with a circular section, a length L in the z-axis direction, and radius $a$; also, the external magnetic field $H_e$ by leakage magnetic flux (x-axis direction). The direction of eddy current is:

- $\frac{\partial B}{\partial t} > 0$: Blue arrow
- $\frac{\partial B}{\partial t} < 0$: Red arrow

The model is shown in Figure 10 was used for the analysis. This model consists of a single copper wire with a circular section, a length L in the z-axis direction, and radius $a$; also, the external magnetic field $H_e$ by leakage magnetic flux (x-axis direction). The direction of eddy current is:

- $\frac{\partial B}{\partial t} > 0$: Blue arrow
- $\frac{\partial B}{\partial t} < 0$: Red arrow
field $H_0$ crosses in the x-axis direction. The theoretical analysis of the eddy current loss was analyzed as follows.

For a copper wire with conductivity $\sigma$, permeability $\mu$, and time factor $e^{j\omega t}$, the magnetic potential $A_z$ is shown in Eq. (19).

$$\frac{\partial^2 A_z}{\partial r^2} + \frac{1}{r} \frac{\partial A_z}{\partial r} + \frac{1}{r^2} \frac{\partial^2 A_z}{\partial \theta^2} = j\omega \mu \sigma A_z,$$

(19)

The solution of Eq. (19) is shown in Eq. (20) and (21):

$$A_z = C J_1(\lambda r) \sin \theta, \quad (r<a)$$

(20)

$$A_z = (Dr + E/r) \sin \theta, \quad (r>a)$$

(21)

$$\lambda^2 = -j\omega \mu \sigma,$$

(22)

In these equations, $C$, $D$, and $E$ are calculated under the following boundary conditions:

(i) In the case of $r = a$, the tangent direction of magnetic field $H_\theta$ and the normal direction of magnetic density $B_r$ are continuity.

(ii) In the case of $r \to \infty$, Eq. (21) should coincide with external magnetic field $H_0 (= \mu H_0 \sin \theta)$.

$$C = \frac{2\mu H_0}{\lambda J_0(\lambda a)}$$

(23)

$$D = \mu H_0,$$

(24)

$$E = \frac{a^2 \mu H_0 J_2(\lambda a)}{J_0(\lambda a)},$$

(25)

In the case of copper wire length $L$, when external magnetic field $H_0$ is crossed, the power flow $P_p$ from the copper wire surface to the internal region is represented by Eq. (26).

$$P_p = \frac{1}{2} a L \int_0^{2\pi} E_z H_0^* |_{r=a} d\theta,$$

(26)

where (*) expresses a conjugate complex number.

The loss of the copper wire is calculated using Eq. (27).

$$P_L = R_e(P_p) = \frac{2 \pi L}{\sigma} |H_0|^2 R_e \left[ \lambda a \left( \frac{J_1(\lambda a) J'_1(\lambda a)}{J_0(\lambda a)} \right) \right],$$

(27)

Eq. (28) can also be obtained from an approximate calculation of the copper wire loss per unit volume by Eq. (27).

$$P_{LI} = \frac{P_L}{\pi a^2 L} = \frac{\pi}{2} (\pi a^2) f^2 \sigma \mu^2 |H_0|^2,$$

(28)

where $f$ is switching frequency.

3.4.2. Consideration of Reducing Eddy Current Losses

The results of the theoretical analysis of eddy current loss based on Eq. (28) indicate that eddy current losses in winding wire are proportional to the cross-sectional area of the litz wire and the square of the switching frequency. If the switching frequency becomes low, it might increase the
magnetic flux density. Therefore, as the reduction method of eddy current loss, the effect of decreasing the strand diameter of the litz wire was examined.

3.4.3. Temperature Variation with Variation of the Strand Diameter of the Litz Wire

Figure 11 shows the measurement result of the primary winding temperature when the strand diameter of the litz wire was varied. The DC resistance of the winding wire is the same for each experiment. The configuration of the litz wire at the time of the experiments is listed in Table 2.

The experimental input voltage of the DC-DC converter is set to 400 V as maximum input voltage and 270 V as rated voltage, respectively. Also, output power is 2150 W. Figure 12 shows the experimental setup. In case that input voltage is 270 V, output power is 2150 W, and the strand diameter of the litz wire is 0.03 φ, the oscillation frequency is 43.5 kHz. Moreover, in case of input voltage 400 V, it is 74.1 kHz. From Figure 11, in the conditions of the strand diameter of the litz wire is thin, and the input voltage is low, the temperature-increase of the primary winding wire becomes low. In the case of an input voltage of 400 V, the temperature-increase in the primary winding for a litz wire strand diameter of 0.1φ was 99 deg. The result was two times higher than that of a strand diameter of 0.03 φ which was 48 deg.

Figure 11. Comparison of experimental and calculated results for temperature and the strand of the litz wire

Figure 12. Experimental setup
3.4.4. Consideration of Relation Between Temperature-Increase and the Strand Diameter of the Litz Wire

To confirm the validity of Eq. (28) for eddy current loss, the experiment has been done. The experimental results are shown in Figure 11. In this figure, the relation between temperature-increase and the strand diameter of the litz wire of primary winding are shown. The temperature is assumed to be proportional to the loss only to simplify the analysis.

The error associated with this equation is too large to enable an actual theoretical analysis because of Eq. (28) is based on a simple model. Therefore, other experimental values were presumed using the two data points marked as “X” (X1, X2) in Figure 11. According to the results, which are shown in Figure 11, at points where the temperature-increase of the primary winding because eddy currents are relatively small. From the result, Eq. (28) coincides with the experimental results.

The calculation values were obtained with the following method. The calculation process is summarized in Figure 13.

1. \( \Delta T_1 \) and \( \Delta T_2 \) are defined as follows:
   \( \Delta T_1 \): \( \Delta T_1 \) is the temperature-increase induced by eddy currents to the primary side in experimental result of point X1.
   \( \Delta T_2 \): \( \Delta T_2 \) is the temperature-increase of the primary side winding caused by Joule heating and tilt heating occurred by secondary winding and iron loss of core except for eddy currents.
2. When the input voltage is high, the switching frequency becomes high; thus, the iron loss increase.
   The temperature increase induced by iron loss is neglected for purposes of calculation. If this temperature-increase is high, the temperature-increase would be high when the input voltage varies from 270 to 400 V, irrespective of the strand diameter of the litz wire. The difference between the calculation and experimental results is small in the cases of 0.03 and 0.06 \( \varphi \).
3. According to Eq. (28), the eddy current loss is proportional to the cross-sectional area of the strand of the litz wire and to the square of the frequency. \( \Delta T_1 \) and \( \Delta T_2 \) are calculated as the temperature-increase induced by eddy current loss using measuring points X1 and X2 for 0.03 \( \varphi\) and 270 V.
4. Obtaining \( \Delta T_1 \) and \( \Delta T_2 \)
At point $X_2$, the strand diameter of the litz wire is twice that at point $X_1$. Therefore, for the eddy current loss to increase fourfold according to Eq. (28), the temperature must also increase fourfold. $\Delta T_2$ is assumed to remain constant concerning points $X_1$ and $X_2$. $\Delta T_1$ and $\Delta T_2$ are then determined from the following simultaneous equations:

At the point $X_1$: DC input $270V$, litz-wire strand diameter $0.03 \, \varphi$

$$\Delta T_1 + \Delta T_2 = 41 \, \text{[deg]},$$

(29)

At the point $X_2$: DC input $270V$, litz-wire strand diameter $0.06 \, \varphi$

$$4\Delta T_1 + \Delta T_2 = 52 \, \text{[deg]},$$

(30)

Solving these equations gives

$$\Delta T_1 = 3.7 \, \text{deg}, \quad \Delta T_2 = 37.3 \, \text{[deg]},$$

(31)

5. Using Eqs. (28) and (31), the other experimental points were estimated. Because of LLC resonant DC-DC converter is controlled by PFM, the switching frequency increases with the input voltage. When the input voltage is becoming higher, the switching frequency is also becoming higher. According to Eq. (28), for the loss by eddy current to be proportional to the square of the frequency. Therefore, $\Delta T_1$ must also increase proportionally.

As the reason for the error of the analysis and experimental results of temperature-increase is occurred because of the definitions that, in the analysis, the magnetic flux affects all part of the winding. Although, in the experiment, the magnetic flux affects a part of the winding.

3.4.5. The result of eddy current countermeasure

Eddy current losses in winding wire are proportional to the square of the resonant frequency and the cross-sectional area of the strand of the litz wire. A decrease in frequency might induce an increase in the magnetic flux density. Therefore, using a smaller diameter wire will reduce the eddy current loss. However, an excessively thin litz wire melts during soldering; thus, a strand diameter $0.03 \, \varphi$ of the litz wire is suitable. Moreover, a temperature-increase of the litz wire with a strand diameter of $0.02 \, \varphi$ is predicted as shown in Figure 11. In the case of a strand diameter of $0.02 \, \varphi$, it is possible to reduce the loss by 4 deg. compared with that in the case of a strand diameter of $0.03 \, \varphi$ at a DC input voltage of $400 \, V$. However, a litz wire with a strand diameter of $0.02 \, \varphi$ is difficult for the manufacturing. If a technique for manufacturing such a wire could be established in the future, it would provide an effective countermeasure to the thin strand diameter of the litz wire.

4. Power Efficiency

The experimental results related to DC-DC converter efficiency are shown in Figure 14 of DC input voltage $270 \, V$ and Figure 15 of DC input voltage $400 \, V$. A maximum efficiency of 98% can be achieved with a strand diameter of $0.03 \, \varphi$, an input voltage of $270 \, V$, an output voltage of $380 \, V$, and an output power of $1.2 \, kW$. 
Table 3 shows the results of a comparison of the efficiency with those previously reported in other works [12] (pp. 1492-1503), [13] (pp. 1212-1222), [14] (pp. 4854-4861), [15] (pp. 1746-1756), [16] (pp. 2982-2991), [17] (pp. 1243-1252), [18] (pp. 6041-6049). As can be seen from the table 3, the efficiency of the DC-DC converter proposed in this study is higher than those of previously reported converters.

Table 3. Comparison of other works

|-------------------|-------------------|--------------------|------------------|------------------------|

Figure 14. Litz wire diameter vs. Efficiency (Input voltage 270 V)

Figure 15. Litz wire diameter vs. Efficiency (Input voltage 400 V)
5. Conclusion

From the above discussion, the following conclusions can be obtained.

- Analysis of LLC resonant DC-DC converter by the FHA method is more practical at higher frequencies than at \( \omega r_1 \).
- The eddy current occurred by the magnetic leakage flux causes power loss in the winding wire.
- The eddy current loss in the winding wire is proportional to the square of the frequency and the cross-sectional area of the strand of the litz wire. Thus, the use of a thinner litz wire can reduce eddy current losses. However, an excessively thin litz wire tends to melt. The litz wire strand diameter of 0.03 \( \phi \) is considered optimum.
- Maximum efficiency of 98 % was achieved with a litz wire strand diameter of 0.03 \( \phi \), an input voltage of 270 V, an output voltage 380 V, and an output power of 1.2 kW.

References

5. Yutaka Suehiro1, Kazuhiro Yamaguchi1, Youichi Ito, Current-Resonant DC-DC-converter Designed to Achieve 95% of Total Efficiency in Power Conditioner for Fuel Cell. INTELEC 2009, Incheon, Korea, October 2009.


