Abstract—In this paper, the method of the realization of a MHz level switching frequency DC-DC converter for high power-density is presented. For high power-density, Gallium Nitride field effect transistor (GaN-FET) and current-mode resonant DC-DC converter are adopted. In addition, the proposed pulse width modulation (PWM) control method which is suitable for the isolated current-mode resonant DC-DC converter operated at MHz level switching frequency, and the novel primary-side zero voltage switching (ZVS) turn on method for the proposed PWM control are presented.

Some experiments have been done with 5MHz isolated DC-DC converter which has GaN-FET, and the total volume of the circuit is 16.14cm³. With the proposed PWM control method, input voltage range is 36-44V, and maximum load current range is 8A at $V_i = 44$V. The primary-side ZVS turn on is confirmed, and the maximum power-efficiency is 89.4%.

Keywords—High Switching Frequency, PWM Control, Current-Mode Resonant DC-DC Converter, GaN-FET

I. INTRODUCTION

Recently, high power-efficiency and high power-density DC-DC converters have been required in information and communication technology (ICT) equipment. Corresponding to the requirement of high power-density, the increase in the switching frequency of these converters has been considered to be one of key technologies. In particular, MHz level switching frequency contributes significantly to downsizing.

However, by adopting high switching frequency, power loss such as switching loss and gate driving loss increases. To solve this problem, the current-mode resonant DC-DC converter is effective because this converter can reduce switching loss. Also, GaN-FETs are suitable for high switching frequency operation as semiconductor switches because of low gate driving loss. Therefore, we have developed a 5MHz current-mode resonant DC-DC converter with GaN-FETs [1].

Current-mode resonant DC-DC converters are usually controlled by pulse frequency modulation (PFM). However, PFM control is hard to design to control the output voltage at MHz level switching frequency operation. The details are going to be described in the section III.

To solve the issue, this paper presents a novel PWM control method for the current-mode resonant DC-DC converter for MHz level switching frequency. This converter topology is same as the conventional current-mode resonant DC-DC converter with synchronous rectification. The feature of the converter is controlling the output voltage without any additional components. By using transformer’s leakage inductance and secondary-side synchronously rectifying switches, the novel control method for boost conversion is realized.

In the previous researches, some PWM-controlled current-mode resonant DC-DC converters have been presented [2-4]. For example, the method of additional auxiliary circuits for regulating output voltage [2-3], and the method of controlling the duty ratio of primary-side switches [4] have been proposed. However, these methods need some additional components for regulating output voltage.
On the other hand, the advantages of the proposed method are controlling secondary-side switches and no additional components.

To accomplish the primary-side ZVS turn on, small magnetizing inductance $L_m$ have been used in current-mode resonant DC-DC converter generally. However, small $L_m$ leads to increase of primary-side current which is cause of reducing power-efficiency. Therefore, to accomplish the ZVS operation in the proposed method, phase-shift between primary and secondary-side switches which control resonant current is adopted. As a result, this proposed converter maintains primary-side ZVS turn on.

The targets of the study are to obtain the high performance which is the small volume, 36-75V or 42-53V of input voltage range, 10A of maximum load current range, the realization of primary-side ZVS turn on, and high power-efficiency.

In the section II, the approach of the realization of DC-DC converter operated at MHz level switching frequency is described. In the section III, the issue of the conventional PFM-controlled current-mode resonant DC-DC converter in MHz level operation is revealed. In the section IV, the proposed PWM-controlled current-mode resonant DC-DC converter is explained. In the section V, the experimental results are revealed.

II. THE APPROACH OF THE REALIZATION OF DC-DC CONVERTER OPERATED AT MHZ LEVEL SWITCHING FREQUENCY

For miniaturization of the DC-DC converter, the increase in the switching frequency of these converters is considered to be one of key technologies. Therefore, this study challenges 5MHz of switching frequency at 120W of output power. Studies that satisfy both the frequency and the output power have not been challenged so far, as shown in Fig. 1. To suppress increasing core power loss with high switching frequency, NiZn ferrite core is used in the study because this core material is suitable for high frequency operation [5]. In addition, to decrease parasitic inductance, multilayer printed circuit board (PCB) and planar transformer like [6], [7] is adopted.

However, by adopting high switching frequency, power loss such as switching loss and gate driving loss increases dramatically. The switching loss $P_{sw}$ is expressed by

$$ P_{sw} = C_{oss} V_{DS}^2 f_s $$

where $C_{oss}$ is output capacitance of FET, $V_{DS}$ is drain-to-source voltage of FET, and $f_s$ is switching frequency. The gate driving loss $P_d$ is expressed by

$$ P_d = Q_G V_{GS} f_s $$

where $Q_G$ is gate charge, and $V_{GS}$ is gate-to-source voltage of FET. From eqs. (1) and (2), these power losses are proportional to $f_s$.

For reducing gate driving loss, GaN-FET is adopted in the study because GaN-FET has low $Q_G$ and be driven at low gate-source voltage. From these reasons, GaN-FET can realize low $P_d$. Also, another researches have been proved utilizing GaN-FET for high switching frequency DC-DC converter is very practical [8-10]. Some literatures [11-13] prove that GaN devices are more effective than silicon devices.

However, because GaN-FETs have low gate-to-source threshold voltage $V_{th}$, low maximum gate-to-source voltage $V_{GSS}$, and high source-to-drain voltage $V_{SD}$, it is difficult to drive these FETs. Therefore, suitable drive circuit for GaN-FET is needed. Some literatures show that driver LM5113 is suitable for driving GaN-FETs [14], [15].

Furthermore, current-mode resonant DC-DC converter topology featuring primary-side ZVS turn on operation is adopted because switching loss can be suppressed. The circuit topology is shown in Fig. 2.

III. THE ISSUE OF THE CONVENTIONAL PFM-CONTROLLED CURRENT-MODE RESONANT DC-DC CONVERTER IN MHZ LEVEL OPERATION

To reduce the switching loss, the current-mode resonant DC-DC converter is widely used. Also, other researches have been proved using current-mode resonant DC-DC converter for high power-efficiency is very practical [16-19]. Generally, the current-mode resonant DC-DC converter is controlled by PFM control which varies switching frequency. PFM control has been valid for the control of the resonant DC-DC converters in kHz level switching operation so far.

To show the static characteristics, the definitions of the converter are

$$ M = \frac{2nV_o}{V_i}, $$

$$ \psi = \frac{L_n}{L_r}, $$

$$ F = \frac{f_s}{f_o}, $$
From literatures, voltage conversion ratio $M$ of PFM-controlled current-mode resonant DC-DC converter is written by

$$M = \frac{1}{\left(1 + \frac{1}{\kappa} \left(1 - \frac{1}{F^2}\right)\right)^2 + Q^2 \left(F - \frac{1}{F}\right)^2}$$ (11)

From eq. (11), the static characteristics of PFM-controlled current-mode resonant DC-DC converter can be leaded. Two examples of the static characteristics are shown in Fig. 3. One is $V_i = 48V$, $n = 2.2$, $L_m = 200nH$, $L_r = 100nH$, $C_r = 10nF$, $Z_0 = 3.16$, $f_0 = 5.04MHz$ and $\kappa = 2$. The other is $V_i = 48V$, $n = 2.2$, $L_m = 200nH$, $L_r = 10nH$, $C_r = 80nF$, $Z_0 = 0.354$, $f_0 = 5.63MHz$ and $\kappa = 20$.

It is assumed that input voltage range is from 42 to 53V at $R_L = 1.22$ and $V_{i} = 12V$. In case of $L_r = 100nH$, for realization of the range, $F$ is changed from 0.81 to 0.99 ($f_0$ is changed from 4.06 to 4.99MHz). In contrast, in case of $L_r = 10nH$, for realization of the range, $F$ is changed from 0.38 to 0.92 ($f_0$ is changed from 2.16 to 5.17MHz).

The difference between two parameters can be confirmed. If switching frequency changes widely for controlling output voltage, the large noise filter will be needed. Therefore, the DC-DC converter is prevented from downsizing by large noise filter. If switching frequency changes narrow for controlling output voltage, the large resonant inductance is needed for controlling.

As a result, in MHz level operation for the miniaturization of the DC-DC converter, it is shown that PFM control is hard to be designed.

IV. THE PROPOSED PWM-CONTROLLED CURRENT-MODE RESONANT DC-DC CONVERTER

The proposed current-mode resonant DC-DC converter can be controlled at fixed switching frequency. In addition, for controlling output voltage, this method need not any additional components.

A. The Circuit Topology

The circuit topology is based on a half-bridge type current-mode resonant DC-DC converter as shown in Fig. 2. The primary-side is the half-bridge topology. $Q_1$ and $Q_2$ are driven in 50% duty ratio, alternatively. $C_{oss}$ and $C_{oss2}$ are parasitic capacitance of primary-side switches.
The circuit can be separated into 5 states in the proposed operation with the switch combination as shown in TABLE I. The operational waveforms are shown in Fig. 4. The equivalent circuits for each state of a half switching term are shown in Fig. 5. In this figure, the switches drawn with weak colors represent OFF, and red line represents current flow. The definitions of the duty ratio \( D \) is followed as

\[
D = \frac{T_\text{on}}{T_s}, \quad (12)
\]

and

\[
D = \frac{1}{2} + D_1 + D_2 + D_3 \quad (13)
\]

\[
D_1 + D_2 + D_3 + D_4 + D_5 = 1/2 \quad (14)
\]

where \( T_s \) is the switching period, and \( T_{on} \) is the on-term of \( Q_3 \) and \( Q_4 \). \( D_1-D_5 \) are the duty ratio of each state.

The definitions of the resonant are followed;

\[
\omega_r = \sqrt{\frac{L_r + L_m}{C_r}}, \quad (16)
\]

\[
Z_r = \sqrt{\frac{L_r + L_m}{C_r}} \quad (17)
\]

and

\[
\omega_{\text{res}} = \sqrt{\frac{L_r C_{\text{res}}}{}} \quad (18)
\]

The definitions of the initial value of the variable are followed;

\[
v_{c1}(0) = V_i/2 - V_c \quad (19)
\]

and

\[
i_{i1}(0) = I_r \quad (20)
\]

The description for each state is described below.

State 1 \((0 < t < D_1 T_s)\):

In this state, \( t_1 \) is defined as \( t_1 = t \). The primary-side switch \( Q_1 \) is turned ON. Also, the secondary-side switch.

TABLE I. CIRCUIT OPERATION STATES

<table>
<thead>
<tr>
<th>State</th>
<th>FET</th>
<th>Diode</th>
</tr>
</thead>
<tbody>
<tr>
<td>Q_1</td>
<td>ON</td>
<td>OFF</td>
</tr>
<tr>
<td>Q_2</td>
<td>ON</td>
<td>OFF</td>
</tr>
<tr>
<td>Q_3</td>
<td>ON</td>
<td>OFF</td>
</tr>
<tr>
<td>Q_4</td>
<td>ON</td>
<td>OFF</td>
</tr>
<tr>
<td>D_1</td>
<td>OFF</td>
<td>OFF</td>
</tr>
<tr>
<td>D_2</td>
<td>OFF</td>
<td>OFF</td>
</tr>
</tbody>
</table>

Fig. 4. The operational waveforms of the proposed PWM control.

Fig. 5. The equivalent circuits for each state of the proposed PWM method.
es both Qi and Qo are turned ON. Qi and Qo are overlapped as indicated with the light gray area in Fig. 4. The resonant inductance \( L \) is magnetized by \( i_i \) for boosting output voltage. From the figure, \( v_c(t) \) and \( i_r(t) \) are become

\[
v_{c1}(t_1) = V_i - \left[ (V_i/2 + V_c) \cos(\omega_0 t_1) + Z_0 I_c \sin(\omega_0 t_1) \right]
\]

and

\[
i_{r1}(t_1) = \frac{-1}{Z_0}(V_i/2 + V_c) \sin(\omega_0 t_1) + I_c \cos(\omega_0 t_1).
\]

(21)

The final values of the state 1 are

\[
V_{c2} = v_{c1}(D_2T_S)
\]

and

\[
I_{r2} = i_{r1}(D_2T_S).
\]

(22)

State 2 \((D_1T_c < t < (D_1+D_2)T_c)\):

In this state, \( t_i \) is defined as \( t_i = t\). After Qi is turned OFF, the direction of the voltage applied to Di is inverted, and diode of Di becomes ON. The inductance current which is magnetized in state 1 flow through diode Di and switch Qi, to the load. From the figure, \( v_c(t) \) and \( i_r(t) \) are become

\[
v_{c2}(t_2) = V_i - nV_o
\]

\[
+ (V_c - V_i + nV_o) \cos(\omega_0 t_2) + Z_0 I_{r2} \sin(\omega_0 t_2)
\]

and

\[
i_{r2}(t_2) = 1/Z_0(V_c - V_i + nV_o) \sin(\omega_0 t_2) + I_{r2} \cos(\omega_0 t_2).
\]

(25)

The final values of the state 2 are

\[
V_{c3} = v_{c2}(D_2T_S)
\]

and

\[
I_{r3} = i_{r2}(D_2T_S).
\]

(26)

State 3 \((D_1+D_2)T_c < t < (D_1+D_2+D_3)T_c)\):

In this state, \( t_i \) is defined as \( t_i = t\). After Qi is turned OFF, the direction of the voltage applied to Di is inverted, and diode of Di becomes ON. The inductance current which is magnetized in state 1 flow through diode Di and switch Qi, to the load. From the figure, \( v_c(t) \) and \( i_r(t) \) are become

\[
v_{c3}(t_3) = V_i + (V_c - V_i) \cos(\omega_0 t_3) - Z_1 I_{r3} \sin(\omega_0 t_3)
\]

and

\[
i_{r3}(t_3) = 1/Z_1(V_c - V_i) \sin(\omega_0 t_3) + I_{r3} \cos(\omega_0 t_3).
\]

(27)

The final values of the state 3 are

\[
V_{c4} = v_{c3}(D_2T_S)
\]

and

\[
I_{r4} = i_{r3}(D_2T_S).
\]

(28)

State 4 \((D_1+D_2+D_3)T_c < t < (D_1+D_2+D_3+D_4)T_c)\):

In this state, \( t_i \) is defined as \( t_i = t\). After Qi is turned OFF, the direction of the voltage applied to Di is inverted, and diode of Di becomes ON. The inductance current which is magnetized in state 1 flow through diode Di and switch Qi, to the load. From the figure, \( v_c(t) \) and \( i_r(t) \) are become

\[
v_{c4}(t_4) = V_i + (V_c - V_i) \cos(\omega_0 t_4) - Z_0 I_{r4} \sin(\omega_0 t_4)
\]

and

\[
i_{r4}(t_4) = 1/Z_0(V_c - V_i) \sin(\omega_0 t_4) + I_{r4} \cos(\omega_0 t_4).
\]

(30)

The final values of the state 4 are

\[
V_{c5} = v_{c4}(D_4T_S)
\]

and

\[
I_{r5} = i_{r4}(D_4T_S).
\]

(31)

C. The Method for Achieving Primary-side ZVS Turn On in the Proposed PWM Control

To accomplish the primary-side ZVS turn on, small magnetizing inductance \( L_m \) have been used generally. However, small \( L_m \) leads to increase of primary-side current which is cause of reducing power-efficiency. Therefore, to accomplish the ZVS operation in the proposed converter, phase-shift between primary and secondary side switches which control resonant current is adopted. As shown in Fig. 4, the proposed operation has phase-
shift between primary and secondary-side switches. \(\Delta t_{ps}\) is the time length of the phase-shift. Without the phase-shift, \(\Delta t_{ps} = 0\text{ns}\), the previous state of the dead-time becomes discontinuous current state. With this situation, initial current cannot be charged enough for ZVS because of secondary parasitic capacitance. Therefore, even with the long dead-time term, ZVS cannot be achieved. With the phase-shift, the problem of the initial current can be solved. From eqs. (34) and (36), the initial current of the dead-time, \(I_{s5}\) is

\[
I_{s5} = \frac{1}{2} Z_0 (V_{c4} - V_i) \sinh(\alpha_b D T_3 S) + I_{r4} \cos(\alpha_b D T_3 S).
\]

(40)

The term of the phase-shift becomes

\[
\Delta t_{ps} = D_3 T_4 + D_4 T_5.
\]

(41)

In eq. (41), as defined as the \(D_3 T_4\) is fixed, the term of state 4 becomes larger with the increase of \(\Delta t_{ps}\). \(I_{s5}\) is related to \(\Delta t_{ps}\), almost linearly. With the enough amount of \(I_{s5}\), the amount of electrical charge \(q_{s5}\) which flows resonant inductance in dead-time becomes larger. For an example, in the conditions of \(C_{oss} = 900\text{pF}\), \(V_i = 36V\), \(R_L = 1.5\text{ohm}\), \(D_3 T_4 = 5\text{ns}\), the relation of \(\Delta t_{ps}\) and \(q_{s5}\) are shown in Fig. 6. The electrical charge \(q_{s5}\) can be calculated with

\[
q_{s5} = \int_{0}^{D_3 T_4} i_s(t) \, dt
\]

(42)

For achieving ZVS, \(q_{s5}\) has to be larger than the amount of the electrical charge of the parasitic capacitance of the switch as

\[
|q_{s5}| > q_{oss}
\]

(43)

where \(q_{oss} = C_{oss} V_i\)

(44)

From the figure, it can be seen that the phase-shift \(\Delta t_{ps}\) between primary and secondary-side switches are valid for ZVS operation for the proposed PWM control.

D. The Advantage of the Proposed Method at MHz Level Operation

Similar to section III, two examples of the static characteristics of the proposed PWM controlled current-mode resonant DC-DC converter are shown in Fig. 7.

It is assumed that input voltage range is from 42 to 53V at \(R_L = 1.2\Omega\) and \(V_i = 12V\). In case of \(L_r = 100\text{nH}\), for realization of the range, \(D\) is changed from 0.5 to 0.61. In contrast, in case of \(L_r = 10\text{nH}\), for realization of the range, \(D\) is changed from 0.5 to 0.55. By comparing between two parameters, the case of \(L_r = 10\text{nH}\) can control in narrow duty ratio than the other one. Therefore, the proposed PWM control method seems to be suitable for the miniaturization of the DC-DC converter because the output voltage can be controlled with small resonant inductance.

V. EXPERIMENTAL RESULTS

In this section, it can be confirmed the difference between the experimental results of the proposed method and the targets of the study.

![Fig 6. The effect of the phase-shift \(\Delta t_{ps}\) vs \(q_{s5}\).](image)

![Fig 7. The static characteristics of resonant DC-DC converter with the proposed PWM control method.](image)

### Table II. Experimental Parameters

<table>
<thead>
<tr>
<th>Specifications</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output reference voltage: (V_o)</td>
<td>12V</td>
</tr>
<tr>
<td>Transformer ratio (n:1)</td>
<td>2:1</td>
</tr>
<tr>
<td>Switching frequency: (f_s)</td>
<td>5MHz</td>
</tr>
<tr>
<td>Resonant frequency: (f_r)</td>
<td>4.98MHz</td>
</tr>
<tr>
<td>Transformer leakage inductance: (L_r)</td>
<td>33nH</td>
</tr>
<tr>
<td>Transformer magnetizing inductance: (L_m)</td>
<td>200nH</td>
</tr>
<tr>
<td>Resonant capacitor: (C_1,C_2)</td>
<td>15.5\muF</td>
</tr>
<tr>
<td>Output capacitor: (C_o)</td>
<td>18.8\muF</td>
</tr>
</tbody>
</table>

### A. The Experimental Conditions

As a prototype digital controller, field programmable gate array (FPGA) Cyclone IV is used, which generates individual gate signal for each switch. The on-term of the gate signals are manually changed with software. The resolution of the gate signals is 1nano second.

Some experiments have been carried out with parameters as shown in TABLE II, and 12V of constant out-
put voltage with open loop control. Components used in the experiment are shown in TABLE III.

The main circuit of the proposed PWM-controlled 5MHz DC-DC converter with GaN-FETs is shown in Fig. 8. The total volume of the main circuit of the proposed 5MHz DC-DC converter is 16.14cm³.

B. The Performance of the Proposed Method

The open loop static characteristics of the 5MHz PWM-controlled DC-DC converter are exhibited as shown in Fig. 9. From the figure, it can be seen that voltage transfer ratio $M$ is controlled by duty ratio $D$, and input voltage range can be achieved 36-44V. The difference between the performance of the proposed method and the targets can be confirmed. The reason is that the ideal transformer turn ratio $n$ is 2.2, however, the actual $n$ is 2. Planar transformer cannot be changed easily because the winding is incorporated into the substrate.

The waveforms of $I_o = 0.89A$ and $I_o = 6.3A$ with the phase-shift are shown in Fig. 10. The blue bar of dead-time shows achieving primary-side ZVS turn on and $\Delta t_{ps}$ is effective for primary-side ZVS operation.

The power-efficiency of the DC-DC converter at $V_o = 12V$ is shown in Fig. 11. The maximum power-efficiency is 89.4%, and maximum load current is 8A at $V_i = 44V$. Also, at low input voltage 44V or 36V, it can be seen that the power-efficiency is relatively low, and maximum load current is low. The reason is that large duty ratio is needed at low input voltage and large output current. The large duty ratio leads to large peak current which is cause of power dissipation. ZVS operation has been confirmed in the range of the experimental conditions.

The temperature distribution of the DC-DC converter as shown in Fig. 12, has been taken of the breadboard at $V_i = 44V$ and $I_o = 8A$. From the results, the temperature of the secondary-side is still in high level. The reason is that large duty ratio leads to the hard-switching of secondary-side switches, and large conduction loss of diodes is happened. On the other hand, the primary-side temperature is relatively low because of achieving ZVS turn on. Therefore, in the future, the hard-switching of secondary-side switches will be improved.

![Fig. 8. The main circuit of the DC-DC converter.](image)

![Fig. 9. The open loop static characteristics of resonant DC-DC converter with the proposed PWM control method.](image)

![Fig. 10. Experimental waveforms of resonant DC-DC converter with the proposed PWM control method.](image)

<table>
<thead>
<tr>
<th>Name</th>
<th>Manufacture</th>
<th>Part Name/ Material</th>
</tr>
</thead>
<tbody>
<tr>
<td>Primary side GaN-FET</td>
<td>EPC</td>
<td>EPC2001</td>
</tr>
<tr>
<td>Secondary side GaN-FET</td>
<td>EPC</td>
<td>EPC2015</td>
</tr>
<tr>
<td>FET Driver</td>
<td>TEXAS INSTRUMENTS</td>
<td>LM5113</td>
</tr>
<tr>
<td>Diode</td>
<td>DIODES</td>
<td>PDS1040L</td>
</tr>
<tr>
<td>Transformer Core</td>
<td>TDK</td>
<td>NiZn Ferrite Core</td>
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<tr>
<td>Resonant Capacitor</td>
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<tr>
<td>Input Capacitor</td>
<td>TDK</td>
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<td>FET Driver</td>
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<td>DE0-nano</td>
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<td>ISO722</td>
</tr>
</tbody>
</table>
VI. CONCLUSIONS

In this paper, the method of the realization of a MHz level switching frequency DC-DC converter for high power-density is described. Furthermore, the novel PWM control method and achieving ZVS operation method for the current-mode resonant DC-DC converter in MHz level operation has been proposed.

The targets of the study is to obtain the high performance which is the small volume, 36-75V or 42-53V of input voltage range, 10A of maximum load current range, the realization of primary-side ZVS turn on, and high power-efficiency.

Some experiments have been done with 5MHz isolated DC-DC converter which has GaN-FET, and the total volume of the circuit is 16.14cm³. With the proposed PWM control method, input voltage range is 36-44V, and maximum load current range is 8A at 44V. The primary-side ZVS turn on is confirmed, and the maximum power-efficiency is 89.4%.

As the future work, to accomplish the targets of the study, detailed analysis of the proposed method with phase-shift, realization of the wide control range and feedback control by digital controller are under considerations.

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REFERENCES