# ZCS-ZVS Mixed Mode PWM Resonant Converter for Snubber Energy Recovery of 3-Level GTO Inverter

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***Abstract*** **-- In this paper, a new type of PWM series resonant converter is proposed. One pole of the proposed converter is switching at zero-current and the other is switching at zero-voltage. By introducing an auxiliary inductor, zero-voltage switching is possible even at no load. The proposed converter is suitable for high-voltage dc/dc converter because the voltage stresses of switching devices are minimized.**

## I. Introduction

Most of the high-voltage high-power inverters use GTO's as their switching devices because GTO is unique self-commutated device that can switch several kA in several kV. In the GTO inverter, both turn-on and turn-off snubbers must be used because of the di/dt and dv/dt problems.[1] The energy of those snubbers which is equal to several percentages of the full rating of the inverter must be transmitted to somewhere during opposite switching transient. In case of low-power inverter, the snubber energy is simply dissipated in the resistor. However, in the high-power inverter, dissipating the snubber energy in the resistor is a bad solution because of the efficiency and cooling problems. Thus, many lossless snubbers have been developed to recover the snubber energy.[2-3] The conventional lossless snubber which is introduced in our 3300V 1MVA GTO 3-level inverter and its rough operation waveforms are shown in Fig.1. The high-voltage dc/dc converter of Fig.1 recovers the collected snubber energy of $V_{cl}$ to source $V_s$. The source voltage is one half of the total dc link voltage and becomes about 2200V and the snubber energy variation is very large in accordance with the inverter switching frequency variation and the inverter load variation. If the $V_{cl}$ which is the input voltage of dc/dc converter increases the voltage stress of GTO also increases but if this decreases the conversion efficiency decreases. Thus there should be some trade-off to determine the snubber side voltage $V_{cl}$. In our inverter, this voltage is chosen about 300V.

In the high-voltage dc/dc converter, the voltage stresses of switching devices and diodes are the most important

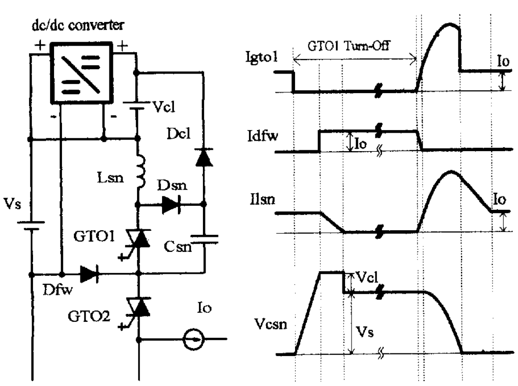



Fig.1 The lossless snubber of 3-level inverter and its rough operation waveforms

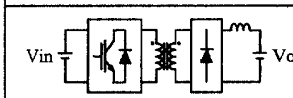
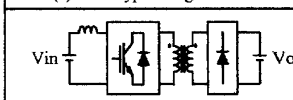
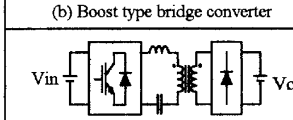
| Bridge types | Voltage stresses | |
|---|---|---|
| | IGBT | Rectifier |
|   (a) Buck type bridge converter | Vin | Much larger than Vo |
|   (b) Boost type bridge converter | Much larger than Vin | Vo |
|   (c) Series resonant converter | Vin | Vo |

Fig.2 Several types of bridge converter and their voltage stresses

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design parameters. Several types of bridge converter and their voltage stresses are shown in Fig.2. The voltage stresses of the devices of (a) and (b) that are connected between the transformer and the filter inductor are very large because of the resonance between transformer leakage inductance and device output capacitance. Thus, we conclude that series resonant converter is suitable for high-voltage dc/dc converter because the voltage stresses of the devices are limited to $V_{in}$ and $V_o$, respectively.

Recently, several types of resonant converters are developed to increase the power density of the converter. However, for almost all of them, either minimum or maximum load is limited for soft switching condition. Thus some technology such as the one used in pseudo-resonant converter must be introduced in order to guarantee soft switching for whole load range.[4]

In this paper, a new series resonant converter (SRC) suitable for high-voltage and wide-load range dc/dc converter is proposed.

## II. The Proposed Converter

The proposed converter for snubber energy recovery can be separated and simply modeled as shown in Fig.3. In this figure, the pulsating snubber current is modeled to average current source $I_{in}$ which varies depending on the switching frequency of the inverter and the inverter load current, This dc/dc converter is somewhat different from

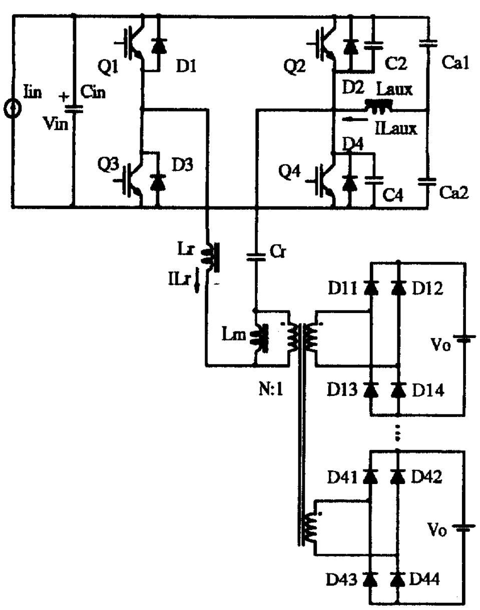



Fig.3 The proposed series resonant converter

conventional dc/dc converter in configuration, that is, the output is the voltage source and the input is the capacitor. Thus the control loop regulates the input capacitor voltage in accordance with input source current by transferring energy into the output voltage source.

Four outputs are connected in series in order to obtain high output voltage. The reason is because the bottle neck in designing high-frequency, high output voltage dc/dc converter is rectifier diodes as the higher breakdown voltage causes the longer reverse recovery time ($T_{rr}$). We must use fast recovery diodes the breakdown voltage of that are not higher than 1000V in order to obtain sufficiently small $T_{rr}$, because the $T_{rr}$ of fast recovery whose breakdown voltage are higher than 1000V are much longer than those of 1000V or lower. Thus, at least four outputs must be connected in series in order to obtain 2800V which is the maximum output voltage.

The conventional control method of SRC is frequency control, however, this is not suitable for wide load range because the variation of switching frequency becomes too large. If the SRC is operated on discontinuous conduction mode (DCM), it can be controlled by constant frequency PWM. The conventional DCM SRC's require reverse blocking diodes in series with the main switches in order to stop resonance at the time when inductor current becomes zero.[5] On the other hand, if the resonant capacitor($C_r$) voltage is less than $NV_O$ at that time, the resonant circuit stops resonance, where N is the transformer turns ratio. By doing so the reverse blocking diodes to the main switches can be eliminated. This condition [7] is given by

$$N^2\left(V_o / I_o\right) \geq \pi Z_r / 2 . \tag{1}$$

The capacitor voltage can be controlled to be smaller than $NV_O$ by charge controlled PWM which is explained in control section.

The resonant frequency is made slightly higher than switching frequency in order to guarantee DCM. The switches $Q_1$ and $Q_3$ are zero-current switched pole and the switches $Q_2$ and $Q_4$ are zero-voltage switched pole.[5-6] In the zero-current switched (ZCS) pole, turn-off switching loss is zero and turn-on switching loss is extremely reduced. In the conventional zero-voltage switching (ZVS) converter with MOSFET, turn-on switching loss is zero and turn-off switching loss is extremely reduced because ZVS capacitor plays the role of turn-off snubber. However, with IGBT, turn-off switching loss is not so much reduced with small ZVS capacitor because of tail current. Thus relatively large ZVS capacitor ($C_{a1}$ and $C_{a2}$) must be selected to reduce turn-off switching loss. The large ZVS capacitors mean that discharging losses at no load become large. Auxiliary inductor ($L_{aux}$), $C_{a1}$ and $C_{a2}$ are introduced in order to guarantee the zero-voltage turn-on of and the

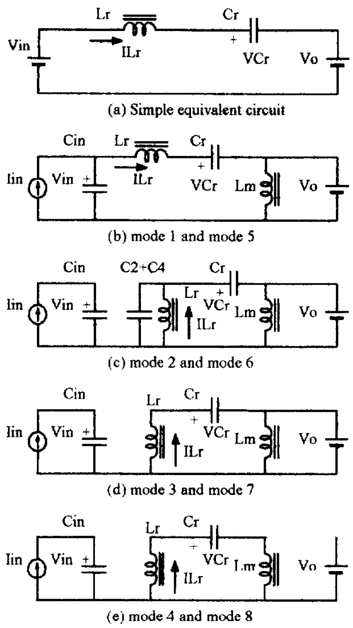



Fig.4 The equivalent circuits of the proposed converter

switches $Q_2$ and $Q_4$ at no load. Conduction loss is increased because of $I_{Laux}$. The increased loss is smaller than 0.2% of output power. $I_{Laux}$ is similar to triangular waveform because the duties of $Q_2$ and $Q_4$ are almost 50% and $C_{a1}$ and $C_{a2}$ are sufficiently large. The energy in $L_a$ is required to be sufficient to turn on opposite diode by itself at the instant of turning off $Q_2$ and $Q_4$. Thus zero-voltage turn-on of $Q_2$ and $Q_4$ is possible at no load.

## III. Operation and Analysis of the Proposed Converter

The equivalent circuits of the proposed converter corresponding to mode variations is shown in Fig.4. Simple equivalent circuit is shown in Fig.4 (a) by relating the input and output with ideal voltage sources and eliminating all the elements except resonant. The differential equations for that circuit are

$$L_r \frac{di_{Lr}(t)}{dt} + v_{Cr}(t) = V_{in} - V_o \tag{2}$$

$$C_r \frac{dv_{Cr}(t)}{dt} - i_{Lr}(t) = 0. \tag{3}$$

The general solution of equations (2) and (3) is

$$i_{Lr}(t) = \cos(\omega_r(t-t_0))i_{Lr}(t_0) - \sin(\omega_r(t-t_0))v_{Cr}(t_0)/Z_r + \sin(\omega_r(t-t_0))(V_{in}-V_o)/Z_r \tag{4}$$

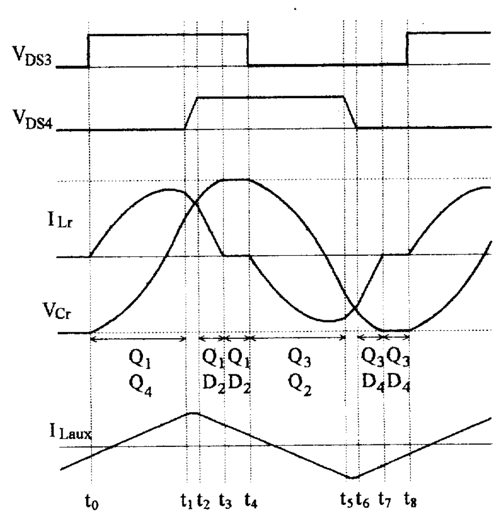



Fig.5 The operation waveforms of the proposed converter

$$v_{Cr}(t) = Z_r \sin(\omega_r(t-t_0))i_{Lr}(t_0) + \cos(w_r(t-t_0))v_{Cr}(t_0) + (1-\cos(w_r(t-t_0)))(V_{in}-V_o)/Z_r \tag{5}$$

where, $Z_r = \sqrt{L_r/C_r}$ is the characteristic impedance and $\omega_r = 1/\sqrt{L_r C_r}$ is the resonant angular frequency.

The waveforms of the proposed converter are shown in Fig.5. The one cycle operation can be divided into symmetrical four modes. The brief explanation of each mode is given as follows:

1) Mode 1 [ $t_0$ - $t_1$ ]

This mode begins at the time of turning on $Q_1$ and ends at turning off $Q_4$. During this mode, $L_r$ resonates with $C_r$ through $Q_1$ and $Q_4$ with the initial values of $i_{Lr}$ and $v_{Cr}$ are zero and $V_{C0}$, respectively. The turn-on switching loss of $Q_1$ is almost zero because the rising time of $i_{Lr}$ is much longer than the switching transient time. During this mode input energy from the snubbers of GTO's transfers to the output and resonant tank. Thus this mode corresponds to effective turn-on time.

2) Mode 2 [ $t_1$ - $t_2$ ]

In this mode, $L_r$, $C_r$ , $C_2$, $C_4$ and auxiliary resonant circuit are activated and $V_{C4}$ increases almost linearly. Thus turn-off switching loss is extremely reduced. This mode is the longest at no load because the interval of mode 1 becomes zero and $i_{Lr}$ is zero. Dead times of $Q_2$ and $Q_4$ are longer than the rising time or falling time of $V_{C4}$ in that case. Mode 2 ends at the time when $V_{C2}$ becomes zero.

3) Mode 3 [ $t_2$ - $t_3$ ]

During this mode, $L_r$ and $C_r$ continue resonance through $D_2$. During this mode resonant tank energy is transferred to the output voltage source $V_o$. $Q_2$ can be turned on at zero voltage during this mode. Thus, turn-on switching loss is zero and reverse recovery problem of $D_2$ does not exist. Mode 3 ends at the time when $I_{Lr}$ becomes zero.

4) Mode 4 [ $t_3$ - $t_4$ ]

In mode 4, the main switches stop operation because $V_{Cr}$ is controlled to be less than $NV_0$. Only small transformer magnetizing current ( $I_{Lm}$ ) flows through $Q_1$. Thus, reverse recovery problem of $D_1$ also does not exist. Mode 4 ends at the time of turning on $Q_3$.

Next four modes are symmetric to mode 1 through mode 4. Thus the current and voltage waveforms are opposite and the other switches of the poles are turned on and turned off.

During the half switching period, the amount of charge that transfers to output is

$$Q_o = \int_0^{T/2} i_{Lr}(t)dt = 2C_r v_{Cr}(0) \qquad (6)$$

$$\int_0^{T/2} i_{Lr}(t)dt = I_o \frac{T}{2} = \frac{V_o T}{2R_o} \qquad (7)$$

where, T is switching period and $R_o$ is effective output

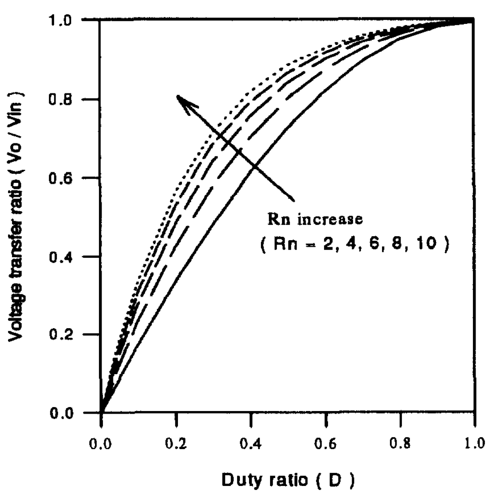



Fig.6 The voltage transfer characteristic curve of the proposed converter

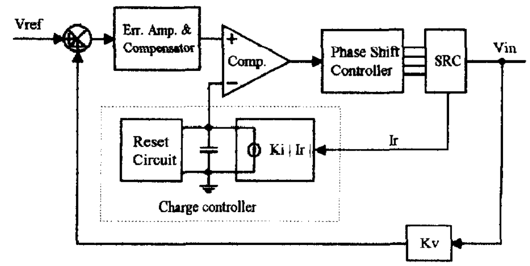



Fig.7 The control block diagram of the phase-shifted charge control

resistance. By solving equation (4) - (7), we can obtain the input to output voltage transfer ratio of the proposed converter. The voltage transfer characteristic curves with respect to $R_N$ and D are shown in Fig.6. In this case $R_N$ is normalized effective output resistance ($R_N = V_o/(I_o Z_r)$) and D is duty ratio ($D = 2(t_1 - t_0)/T$).

## IV. Control of the Proposed Converter

The control block diagram is shown in Fig.7 which shows phase-shifted charge control method.[9-10] In the charge control, the integration of the load current is used as ramp voltage as shown in Fig.7. The integration of the resonant current is, however, proportional to the resonant capacitor voltage in the proposed converter. Thus, in this charge control, the resonant capacitor voltage is automatically controlled and limited.

In the proposed converter, the object of control is, however, not to regulate the output voltage but to regulate the input voltage. Thus, positive error amplifier must be used for negative feedback control as shown in Fig.7 in order to use the commercial peak current mode control IC's such as KA3846 or KA3825.

## V. Experimental Verification

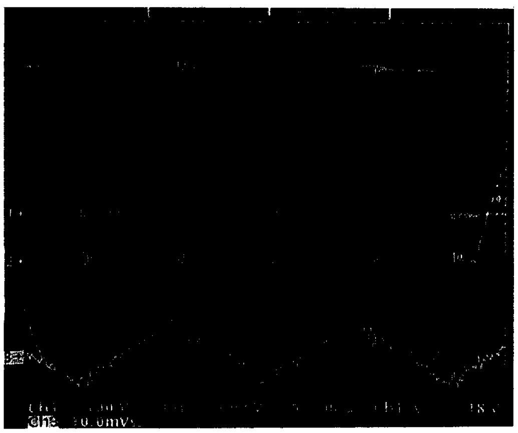
A 10kW prototype of the proposed converter is implemented for experimental verification. Input and output specifications are as follows:

Input voltage : 300V
Input power : 10kW
Output voltage : 1800V - 2800V.

Components used in the experiment are as follows:

$Q_1$ - $Q_4$ & $D_1$ - $D_4$ : CM200DY-12H(POWEREX)
$D_{11}$ - $D_{44}$ : DSEI2x61-10B(IXYS)
Transformer : 2xUU120, PE-1, 12:30x4
$C_2$ - $C_4$ : 40nF
$C_{a1}$ - $C_{a2}$ : 10uF

(a)



(b)

Fig.8 Experimental waveforms
(a) no source current, 5μS/div.
ch.1 : $V_{c4}$, 100v/div.
ch.2 : $I_{Lr}$, 5A/div.
ch.3 : $I_{Laux}$, 20A/div.
(b) half source current, 5μS/div.
ch.1 : $V_{c4}$, 100v/div.
ch.2 : $I_{Lr}$, 20A/div.
ch.3 : $I_{Laux}$, 20A/div.

$C_r$ : 1.8uF,
$L_r$ : 4.7uF (The transformer leakage inductance is included)
$Z_r$ = 1.616Ω
$L_{aux}$ : 50uH

Using commercial peak current mode control IC(KA3846, SAMSUNG) and peripheral circuits, phase-shifted charge controller is implemented.

(c)

Fig.8 Experimental waveforms(continued)
(c) full source current, 5μS/div.
ch.1 : $V_{c4}$, 100v/div.
ch.2 : $I_{Lr}$, 50A/div.
ch.3 : $I_{Laux}$, 20A/div.

The switch voltages and inductor currents are shown in Fig.8. The switching frequency is about 50kH.

## VI. Conclusion

In this paper, a new DCM SRC is proposed. The voltages of all the switching devices and rectifiers are clamped to either input or output side voltage. Thus, the proposed converter is suitable for extremely wide load range and for high voltage dc/dc converter. Soft switching is possible in the whole load range by mixing one zero-voltage switched pole and the other zero-current switched pole having auxiliary series resonant circuit. The phase-shifted charge control enables the proposed converter to eliminate reverse blocking diodes. This control method also makes constant switching frequency operation possible as well.

The operation of the proposed converter is verified by the experiments

## VII. References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