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Analysis of a Three-Level LLC Series Resonant Converter for High- and Wide-Input-Voltage Applications

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ABSTRACT –
In this paper, the analysis of a three-level LLC series resonant converter (TL LLC SRC) for high- and wide input-voltage applications is presented. It consists of two half-bridge LLC SRCs in series, sharing a resonant inductor and a transformer. Its main advantages are that the voltage across each switch is clamped at half of the input voltage and that voltage balance is achieved and simple driving signals . Thus it is suitable for high-input-voltage applications. Based on the results of these analyses, a design example is provided and its validity is confirmed by an experiment involving a prototype converter with an input of 600V and an output of 48 V/20 A.
Keywords — High and wide input voltage, LLC series resonant converter (SRC), three-level (TL).

I. INTRODUCTION

In this paper, a new TL LLC SRC for high- and wide-input voltage applications is introduced, and its analysis results under wide input voltage are presented. In high-input-voltage applications such as three-phase systems, fuel cell systems, photovoltaic systems, and ship electric power distribution systems, the three-level converter (TLC)is advantageous because the voltage across each main switches half of the input voltage. However, the TLC is associated with high levels of switching loss due to its high input voltage, a condition that results in low efficiency. For this reason, it is necessary to apply a soft-switching technique when using a TLC. All main switches and rectifier diodes in the converters are softly switched without additional circuits under all load conditions. While this paper deals with a TL LLC SRC, the analysis results can be applied to other TL LLC SRCs for wide-input-voltage applications**.**

II.THREE-LEVEL LLC SERIES RESONANT CONVERTER

Fig. 1 shows the circuit configuration of the TL LLC SRC. The TL LLC SRC discussed in this paper consists of two half-bridge (HB) LLC SRCs in series that share a resonant inductor and a transformer. The upper switches of each HB LLC SRC are driven with a constant duty ratio (D = 0.5) simultaneously. On the other hand, the lower switches of each HB LLC SRC are driven complementarily to the upper switches. These operations allow the voltage across each switch to be clamped at half of the input voltage and that voltage balance to be achieved.

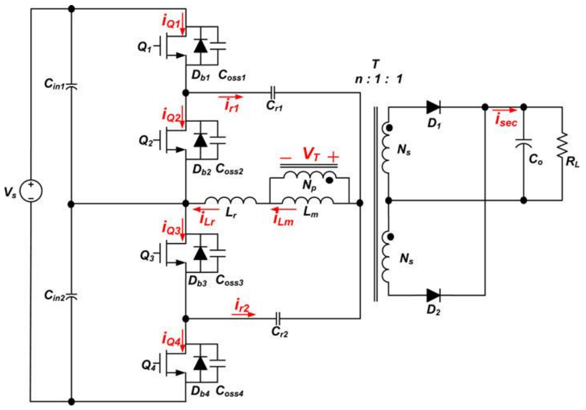

Fig.1. Circuit Diagram of Three-Level LLC Series Resonant Converter

The converter also has several advantages as follows. First, its driving signals are simple.

Thus, it can be readily implemented using only a general-purpose controller without additional circuits. Only two gate-driving ICs or gate-driving transformers are required. Because no additional dead time between the driving signals is required, the efficiency can be easily optimized. Moreover, clamping diodes are not required. All main switches and rectifier diodes are softly switched under all line and load conditions. High efficiency can be achieved over a wide-input-voltage. The voltages of resonant capacitors are a quarter of the input voltage, allowing the use of a low voltage rating capacitor. Due to these advantages, it can be said that the converter is suitable for high- and wide input-voltage applications.

III. STEADY-STATE OPERATION UNDER A HIGH INPUT VOLTAGE

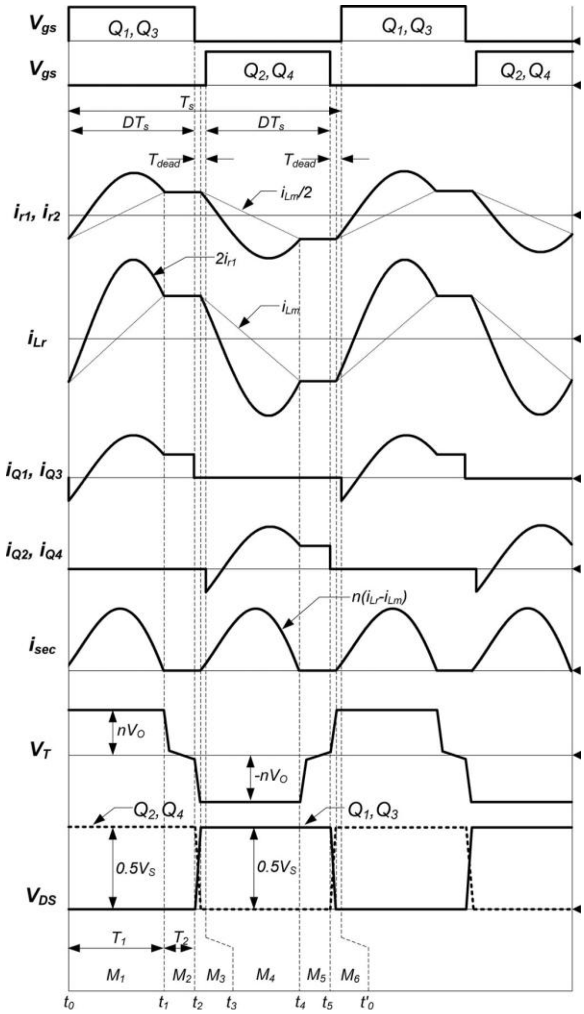

Fig.2 Key operating waveforms of the converter in a steady state.

Referring to the figure 3, the switches Q_1 and Q_3 are driven with a constant duty ratio (D = 0.5) simultaneously. On the other hand, the switches Q_2 and Q_4 are driven complementarily to switches Q_1 and Q_3 .In order to illustrate the operation of the converter; several assumptions are made as follows.

1) The output capacitor C_O is large enough to be considered as a voltage source.
2) Two input capacitors in the series C_{in1} and C_{in2} are large enough to be considered as two voltage sources of $V_S/2$.
3) The main switches are all MOSFETs with parasitic diodes of D_{b1} , D_{b2} , D_{b3} , and D_{b4} .
4) The output capacitances of all MOSFETs have the same capacitance of C_{oss} .
5) The two resonant capacitors C_{r1} and C_{r2} have the same capacitance of C_r .
6) The main transformer T has a turn ratio of $n = N_P/N_S$

Mode 1 [t_0–t_1]:

Mode 1 begins when the switches of $Q_1$and Q_3 are turned ON with ZVS at t_0 . The current of L_r $i_{Lr}(t)$ increases with a sinusoidal shape by the resonance of C_{r1} , C_{r2} ,and L_r and is divided into $i_{r1}(t)$ and $i_{r2}(t)$. The rectifier diodeD_1 is also conducting. The primary voltage of transformer $V_T(t)$ clamps at nV_O , where n is the turn ratio and V_O is the output. The magnetizing current $i_{Lm}(t)$ is linearly increased from a negative to a positive value by the reflected voltage nV_O. During this mode, the difference between $i_{Lr}(t)$ and $i_{Lm}(t)$ is transferred to the output stage.

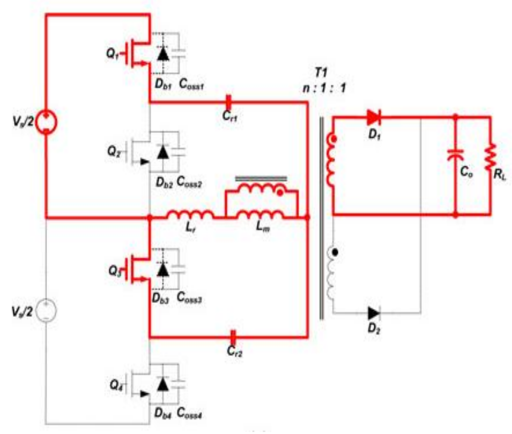

Fig. 3 Operating circuits during one switching period at Mode 1.

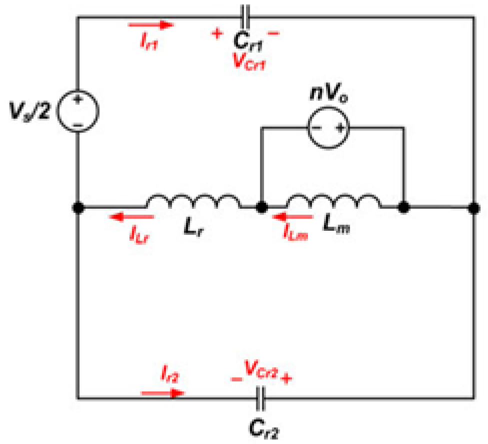

Fig. 4 Equivalent circuit describing mode 1

The voltages and currents during this mode are expressed as follows:

$$(V_{Cr1}t) = \frac{VS}{2 - nVO - Vk1\cos\left[\omega O\,(t - t0) + \phi k1\right]} \quad 1$$

$$V_{Cr2}\,(t) = \frac{VS}{2 - VCr1\,(t)} \quad 2$$

$$ir1\,(t) = ir2\,(t)$$

$$= \frac{Vk1\sin\left[\omega O\,(t - t0) + \phi k1\right]}{2zO} \quad 3$$

$$i_{Lr}\,(t) = 2i_{r1}\,(t)$$

$$= \frac{Vk1\sin\left[\omega O\,(t - t0) + \phi k1\right]}{zO} \quad 4$$

$$iLm(t) = iLr\,(tO) + \frac{nVO\,(t - t0)}{Lm} \quad 5$$

$$iQ1\,(t) = iQ3\,(t) = \frac{iLr\,(t)}{2} \quad 6$$

$$V_{DS\,Q2}(t) = V_{DS\,Q4}(t) = \frac{V_S}{2} \quad 7$$

$$i_{sec}(t) = n(i_{Lr}\,(t) - i_{Lm}(t)) \quad 8$$

Where

$V_{k1} = \sqrt{(z_O\, i_{Lr}\,(t_0))^2 + (V_S/2 - V_{Cr1}\,(t_0)\ nV_O)^2}$,

$\phi_{k1} = \tan^{-1}\frac{(z_O\, i_{Lr}\,(t_0)(2 - V_{Cr1}\,(t_0) - nV_O))}{V_S}$,

$\omega_O = 2\pi f_O = \frac{1}{\sqrt{2L_rC_r}}$, $z_O = \frac{L_r}{2C_r}$,

Mode 2 [t1 –t2]:

Mode 2 begins when $i_{Lr}\,(t)$ goes back to the same level as $i_{Lm}(t)$. At this moment, D_1 is turned OFF with zero current switching (ZCS) and the secondary side of the transformer is separated from the primary side. Then, the magnetizing inductance Lm participates in the resonance. During this mode, the power to the load is only supplied by the output capacitor C_O . The equivalent circuit describing this mode can be obtained by removing nV_O from the circuit model related to mode 1.

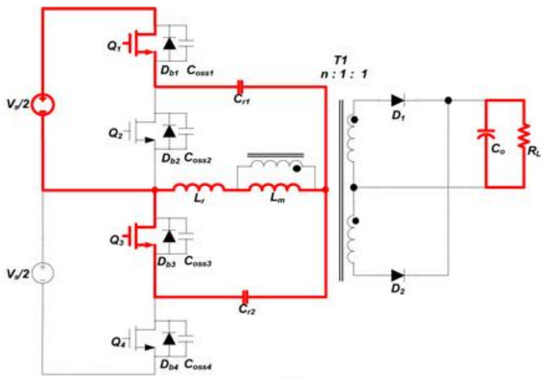

Fig. 5 Operating circuit during one switching period at Mode 2.

The voltages and currents during this mode are expressed as follows:

$$VCr1\,(t) = \frac{VS}{2 - Vk2\cos\left[\omega O\,(t - t1) + \phi k2\right]} \quad 9$$

$$VCr2\,(t) = \frac{VS}{2 - VCr1\,(t)} \quad 10$$

$$ir1\,(t) = ir2\,(t) = \frac{Vk2\sin\left[\omega' O\,(t - t1) + \phi k2\right]}{2z'O} \quad 11$$

$$i_{Lr}\,(t) = \frac{V_{k2}\sin\left[\omega'_O\,(t - t1) + \phi_{k2}\right]}{2z'_O} \quad 12$$

$$iLm(t) = iLr\,(t) \quad 13$$

$$V_{DS\,Q2}(t) = V_{DS\,Q4}(t) = \frac{V_S}{2} \quad 14$$

Where

$V_{k2} = \sqrt{(z'_O i_{Lr}(t_1))^2 + (V_{Cr2}\,(t_1))^2}$

$\phi_{k2} = \tan^{-1}(z_O i_{Lr}(t_1)$

$\omega''_O = 2\pi f'_O = \frac{1}{\sqrt{2(L_r + L_m)C_r}}$,

$z'_O = \sqrt{(L_r + L_m)/2C_r}$

Mode 3 [t2 –t3]:

Mode 3begins when the switches of Q_1 and Q_3 are turned OFF. Then, the voltages across Q_1 and Q_3 are linearly increased from zero and the voltages across Q_2 and Q_4 are linearly decreased from half of the input voltage by $i_{Lm}(t_2)$.In addition, $V_T\,(t)$ is decreased to $-nV_O$. If the voltages across Q_1 and Q_3 become half of the input voltage, $V_T\,(t)$ becomes $-nV_O$ and D_2 begins conducting. At the same time, the parasitic diodes of Q_2 and Q_4 , namely D_{b2} and D_{b4} , are forward biased and the resonance of C_{r1} , C_{r2} , and L_r occurs in again. At the end of this mode, the switches of Q_2 and Q_4 are turned ON with ZVS.

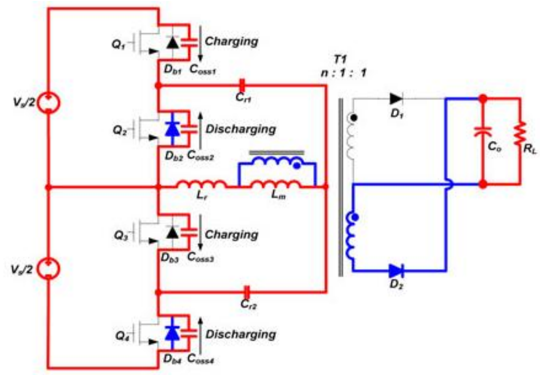

Fig. 6 Operating circuit during one switching period at Mode 2.

The voltages and currents can be expressed as follows:

$$V_{Cr1}(t) = V_{Cr1}(t_2) + \frac{iLm(t2)(t-t2)}{2cr} \quad 15$$

$$VCr2\,(t) = \frac{VS}{2 - VCr1\,(t)} \quad 16$$

$$ir1\,(t) = ir2\,(t) = \frac{iLm(t2)}{2}$$

$$= \frac{-iLm\,(t0)}{2} \quad 17$$

$$VDS\ Q1\,(t) = VDS\ Q3\,(t) = \frac{iLm\,(t2)(t-t2)}{4Coss} \quad 18$$

$$VDS\ Q2\,(t) = VDS\ Q4\,(t) = \frac{VS}{2} - \frac{iLm\,(t2)(t-t2)}{4Coss} \quad 19$$

The operating principles for modes 4–6, are also similar to modes1–3. .During the mode 4[t_3–t_4]: The switches of *Q2* and *Q4* are turned ON with ZVS and the difference between *iLr (t)* and *iLm(t)* is transferred to the output stage. During the mode 5[t_4–t_5]: *D*2 is turned OFF with zero current switching (ZCS) and the secondary side of the transformer is separated from the primary side. the power to the load is only supplied by the output capacitor *Co*. During the mode 6 the switches of *Q2* and *Q*4 are turned OFF with ZVS At the same time, the parasitic diodes of Q_1 and Q_3 , namely D_{b1} and D_{b3} , are forward biased and the resonance of C_{r1} , C_{r2} , and L_r occurs in again. Thus continuously the load is provided with dc supply in all six modes of operation of the converter.

IV. STRESS ANALYSIS

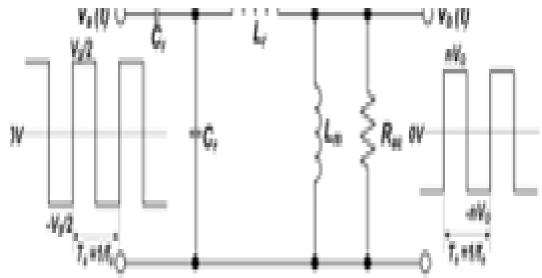

Fig.7. AC equivalent circuit for the discussed converter

From the equivalent circuit for the converter is derived shown in Fig.7, the ac equivalent load resistance R_{ac} and the fundamental components of the input and output voltages of the resonant tank are as follows:

$$Rac = \frac{8n2RL}{\pi 2} \quad 20$$

$$V_{Fa}(t) = \frac{V_{Fa}(t)}{\pi} \quad 21$$

$$V_{Fb}(t) = \frac{4nV_o \sin(2\pi f s t).}{\pi} \quad 22$$

The gain of the resonant tank *k is* expressed as follows:

$$k = Lm/Lr \quad 23$$

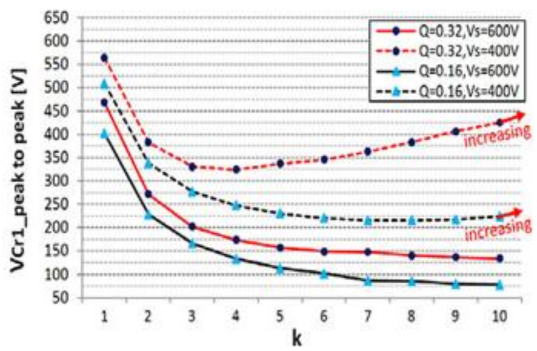

Fig 8 Peak-to-peak voltage stress of the resonant capacitor.

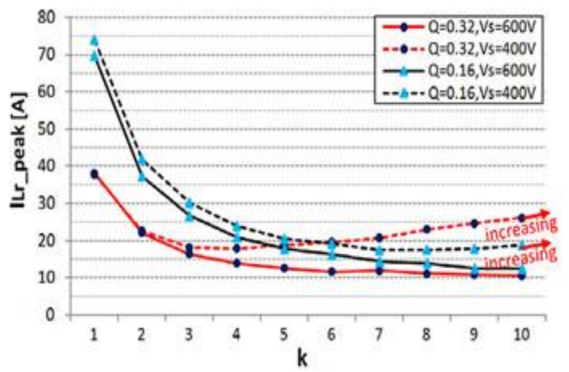

Fig 9 Peak current stress of resonant inductor.

From the fig 8, all stresses of the resonant tank decrease initially and finally become constant with increasing *k* at *Vs* max. This occurs because the energy stored in *Lm* decreases as *k* increases but shows nearly no change at larger *k* values. In addition, the lower value of *k* increases the current stress of the resonant inductor because the energy stored in *Lm* increases due to the lower resonant frequency. All stresses of the resonant tank increase as the input voltage decreases. This arises because the energy stored in *Lm* increases due to the lower switching frequency. For this reason, it is important to consider all stresses at *Vs* min in applications with a wide input voltage range. For stresses of the resonant tank at *Vs* min, whereas they decrease initially, eventually increase as *k* increases. This occurs because the energy stored in *Lm* decreases at first while the energy stored in the resonant capacitor increases due to the increasing value of T_2 as *k* increases. Moreover, a higher *k* value increases the voltage stress of the resonant capacitor but makes the current stress of the resonant inductor lower at relatively low *k* values.

V. SIMULATION RESULTS

Three level LLC Series resonant converter using soft switching technique is analyzed and simulated using MATLAB. Switching frequency of the is converter made close to resonance frequency. Fig 4 and fig 5 shows the output voltage and output current.

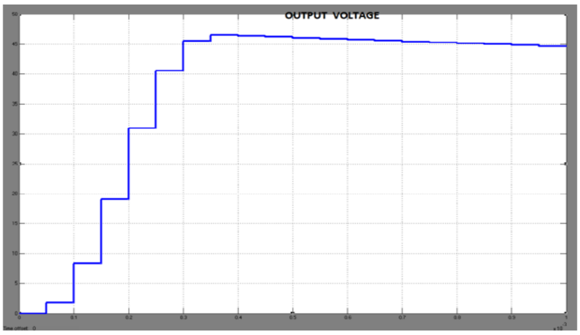

Fig .10 Output Voltage

The output voltage is 48V with the maximum input of 600 volt..

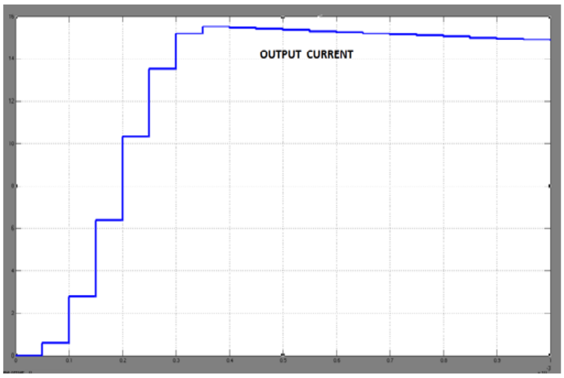

Fig .11 Output Current

The converter is operated with the switching frequency of 50 kHz.

VI CONCLUSION

This paper has analyzed a series resonant inverter Thus so far, there has been little research on the topic of TLLLC SRCs. From the analysis results, it is confirmed in Section V that in the proposed converter, the voltage across each switch is half of the input voltage, while the efficiency is 95.1% under a full load(20 A) and a 600-V input.. These analysis results can be applied to all TL LLCSRCs for wide-input-voltage applications.

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