NEW BILATERAL ZERO VOLTAGE SWITCHING AC/AC CONVERTER USING HIGH FREQUENCY PARTIAL-RESONANT LINK

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ABSTRACT

A new zero voltage switching partial-resonant link AC/AC converter which overcomes the shortcomings of conventional resonant ac and dc link AC/AC schemes and consists of only 12 unidirectional switches is proposed. The proposed converter synthesizes output ac waveforms with integrals of high-frequency current pulses like series-resonant converters. But switching operations occur at zero voltage instants instead of at zero current instants. The partial-resonant link normally does not resonate during entire operation interval but does only for the duration of switching transients. In particular, the proposed converter can easily and exactly control individual current pulse amplitude in each switching cycle. Symmetrical configuration of the power scheme also provides fully regenerative operation with buck-boost capability. The advantages of the proposed scheme are shown through analyses and simulation results.

I. INTRODUCTION

In general, switching schemes with high frequency resonant AC/AC converters can be classified according to their resonant ac-link and dc-link modes. The high frequency link AC/AC converter using resonant ac- or dc- link schemes which utilize high speed devices such as fast recovery transistors and GTOs have been reported extensively [1]-[9]. These converters not only have high power density but also reduce or eliminate switching loss since converter switching occurs at zero crossings of the link voltage or current and thus enable the total system to operate at very high frequency. Further, the use of the high frequency resonant link speeds up the system response, reduces acoustic noise and provides high resolution output waveforms.

The ac resonant circuit, however, impresses both polarities of ac voltage and current on the link [1]-[3]. The converter switches should be bidirectional, which are usually realized by two inverse-parallel transistors or thyristors for the series and parallel-resonant circuits. As a result, the resonant ac link AC/AC converter requires a total of 12 bidirectional switches or 24 transistors or thyristors for full double bridge. In addition, since the resonant ac-link converters transfer power via resonant tanks, excessive VA ratings of the LC components and the device are usually required to maintain continuous oscillations.

Recent new approaches, using series- or parallel-resonant dc link schemes are proposed in [4]-[7]. The dc link circuit realizes pulsating dc currents or voltages by adding dc offsets to ac-resonant current or voltage to use unidirectional switches. Thus such a system has simple configuration having only 12 unidirectional devices. However their common several main disadvantages are, first they require either large dc inductor or capacitor, second they require high VAR raings of resonant elements L_r and C_r due to full-cycle resonant operation, and third they have high dissipations in the inductor and capacitor. Besides they require additional clamping circuits and substantially complex controls to prevent excess overcharge of the resonant capacitor or current pulse fluctuations under transient conditions [5],[7].

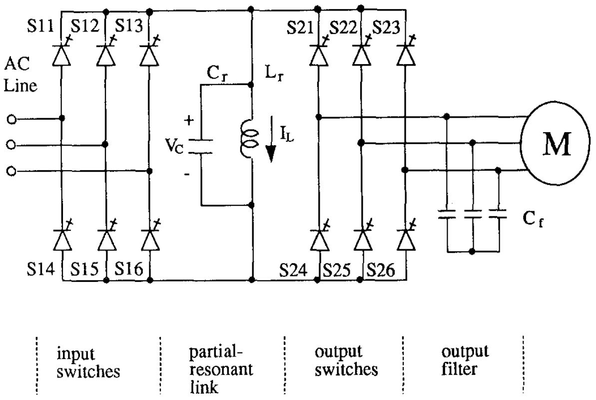


Fig. 1 Proposed zero voltage switching partial-resonant link AC/AC converter.

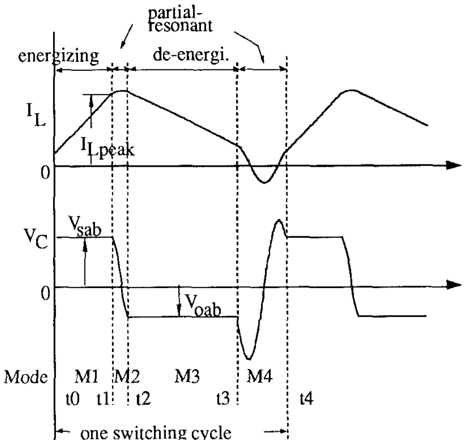


Fig. 2 Typical waveforms illustrating the operation principles of the proposed converter.

A somewhat different type of series-resonant AC/AC converter with reduced number of switches is proposed in [8]. In this case, the converter also requires only 12 unidirectional and natural current commutation switches. However, it also has similar inherent shortcomings in the power transfer and control method [9] such as high VAR ratings of resonant elements, partly hard switching, and high voltage and current stresses.

In this paper, a new zero voltage switching partial-resonant link AC/AC converter which overcomes most of the aforementioned disadvantages and consists of 12 unidirectional switches is proposed. The proposed converter synthesizes output ac waveforms by integrating high-frequency current pulses like series-resonant converters. But switching transients does not occur at zero current instants but at zero voltage instants. The partial-resonant link normally does not resonate during entire operation interval but operates only during the switching transients to ensure zero voltage switching condition. Thus this converter has low VAR ratings of resonant elements L_r and C_r and has low dissipations. In particular, the proposed converter can control individual current pulse amplitude very easily in each switching cycle, which enables improving the output voltage waveform and solving the instability problem usually caused by parasitic resonance between output capacitor C_o and load inductance L_o [6]-[7]. Symmetrical arrangement of power circuit in the proposed converter also provides fully regenerative operation together with buck-boost capability. The advantages of the proposed scheme are shown through analyses and simulation results.

II. PRINCIPLE OF OPERATION

The proposed partial-resonant link AC/AC converter configuration is shown in Fig. 1, which is composed of double full bridge, partial-resonant link, output filter capacitor tank and load. The power circuit requires minimum number of 12 unidirectional devices and has complete symmetricity. The partial-resonant link consists of one small inductor and one capacitor. The value of resonant capacitor is very small compared with that of output filter capacitors.

Each switching cycle can be divided into four modes, that is, energizing, de-energizing and two resonant modes according to the direction of energy flow to the resonant inductor L_r as shown in Fig. 2.

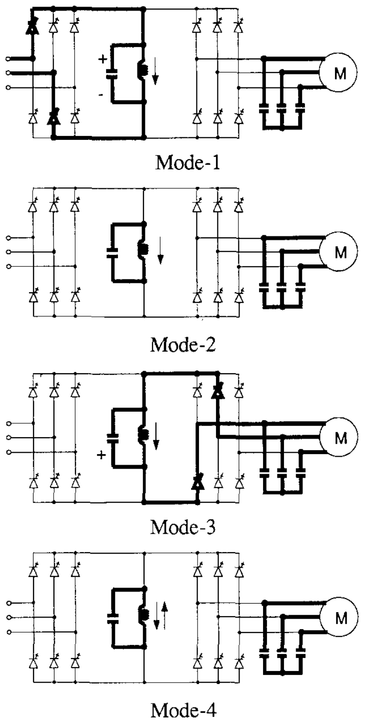
(I). Mode 1 (*energizing*) : Two of switches, S_{11} and S_{15}, are on-state and the others are off-state as shown in Fig. 3(a). L_rC_r tank remains connected to AC source line-line voltage V_{sab} and the inductor current is given by

$$\frac{dI_L}{dt}=\frac{V_{sab}}{L_r} \tag{1}$$

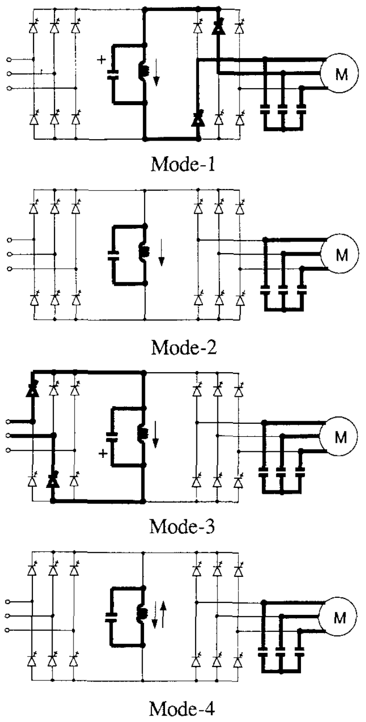
The inductor current increases linearly to a certain value required to compensate output voltage error. When the inductor current reaches the required value I_{Lpeak}, switches S_{11} and S_{15} are turned off under zero voltage switching condition.

(II). Mode 2 (*partial-resonant*) : All the switches remain off-state. The partial-resonant link L_r and C_r resonate partially in order to ensure zero voltage turn-on condition of on-coming switches S_{22} S_{24}. Thus the resonant capacitor voltage V_C decreases via zero down to on-coming output line-to-line voltage V_{oab}. When V_C becomes equal to V_{oab}, switches S_{22} and S_{24} are turned on with zero voltage switching condition.

(III). Mode 3 (*de-energizing*) : L_rC_r tank remains connected to the output line-to-line voltage to which the stored inductor energy is delivered so as to compensate the output voltage error. The inductor current decreases linearly to a predetermined value $I_{\min}$.



(a) powering operation



(b) regenerative operation

Fig. 3 Topological mode diagrams (a) powering operation, (b) regenerative operation.

The current value is needed minimally to change the resonant capacitor voltage to on-coming input line-to-line voltage to which L_rC_r tank is connected for next switching cycle. When the inductor current becomes equal to the minimum current, switches S_{22} and S_{24} are turned off.

(IV). Mode 4 (*partial-resonant*) : All the switches remain off state. The partial-resonant link resonates for approximately one resonant period until the resonant capacitor voltage reaches to the equal value of the next on-coming input line-to-line voltage. When the capacitor voltage becomes equal to the next-connected input voltage, related switches are turned-on under zero voltage switching condition. Thus next switching cycle starts.

Even though the operation modes are illustrated for powering state, symmetrical characteristic of the power circuit provides fully regenerative operation as well as buck-boost capability as shown in Fig. 3(b). During energizing mode, the energy from AC source is transferred to the resonant inductor. The amount of energy $\frac{1}{2}L_rI^2_{Lpeak}$ stored in the inductor is delivered to AC load during de-energizing mode. By repeating this process alternately, we can transfer power from AC source to AC load and can also synthesize three phase output voltages. In particular we can easily control the amount of the stored energy or current pulse amplitude I_{Lpeak} by controlling time-interval $T_{10}(=t_1-t_0)$ of energizing mode as given by

$$I_{Lpeak} = \frac{V_{sab}}{L_r}T_{10}. \tag{2}$$

depending on the output voltage error. The amplitude control of individual current pulse together with high switching frequency operation makes it possible to control an output phase voltage exactly and fastly. Thus output voltage waveform can have high quality without having any oscillation problems.

During two resonant modes, the partial-resonant link L_r and C_r is disconnected from AC source and load, and partially resonates. The resonant inductor current charges or discharges the resonant capacitor until the capacitor voltage equals to the on-coming line-line voltage to be switched on zero voltage condition for the next switches. By this partial-resonant operation, it has so many merits such as low VAR rating of resonant elements, low dissipations, low voltage and current stress, current-pulse amplitude control, etc. compared with other converters [4]-[8].

As mentioned earlier, the proposed converter synthesizes output ac waveforms by integrating high-frequency current pulses like series-resonant converters. Switching transients occur with zero voltage crossings instead of zero current crossings. The partial-resonant link resonates only for the duration of switching transients to ensure zero voltage switching condition, and normally it does not resonate over the entire operation interval. Thus the resonant capacitor is minimally involved in the main power transfer and is used mainly for snubbing or resonant operation for zero voltage switching.

III. OUTPUT VOLTAGE REGULATION METHOD

Fig. 4 shows an overall control block diagram of the proposed converter to regulate three-phase output voltages. It can be operated to regulate three-phase load currents for such an application as induction motor drive. In Fig. 4, the controller compares the reference signals of three line-to-line output voltages with their feedback signals and generates three errors (e_{oab}, e_{obc}, e_{oca}). The line-to-line output voltage having the largest absolute error can be chosen for voltage regulation. Output selection logic also determines the direction of power flow (DPF), if the system operates on either powering or regenerative mode by sensing both polarities of the largest error and the corresponding reference voltage. The direction of power flow can be logically given by

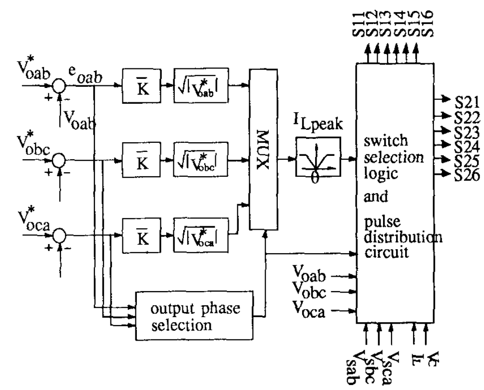


Fig. 4 Overall control block diagram of proposed converter.

$$DPF = S_ES_R + \bar{S}_E\bar{S}_R \tag{3}$$

where S_E is the sign of the largest error and S_R is the reference voltage sign. These processes are performed at the start point of mode 4 of last switching cycle. In addition, at the time point, the controller should compute the I_{Lpeak} required to compensate the output voltage error.

Under no load condition, the amount of energy for compensation can be expressed as

$$\Delta W = \frac{1}{2}C_{eq}(V^{*2}_{oab} - V^2_{oab}) \tag{4}$$

where $C_{eq} = \frac{1}{2}C_f$

if the line-line output voltage having the largest error is V_{oab}. The corresponding peak value of current pulse can be given by

$$I_{Lpeak} = \sqrt{\frac{2\Delta W}{L_r}} \tag{5}$$

The required energy should be stored in the resonant inductor during energizing mode of present switching cycle and then be transferred to the line-line output voltage during de-energizing mode. Thus, the output voltage error could be corrected exactly.

In the actual operation, the output voltage should be regulated to minimize the output voltage error for any load condition within a certain limit range. Thus the controller should be designed to have feedback loop containing an adaptive gain loop. Assuming that $|V^*_{oab} - V_{oab}|$ is sufficiently small and substituting Eq. (4) into Eq. (5) we can modify Eq. (5) as

$$I_{Lpeak} \overset{-}{=} \sqrt{2C_{eq}V_{oab}^*/L_r}\ \sqrt{|V_{oab}^* - V_{oab}|}. \tag{6}$$

The modified equation can be written as follows to be considered in the feedback loop as

$$I_{Lpeak} = K\sqrt{V^{*oab}}\ \sqrt{|V_{oab}^* - V_{oab}|}. \tag{7}$$

This relation can be easily implemented in the controller as shown Fig. 4.

IV. ANALYSIS AND CHARACTERISTICS OF THE PROPOSED CONVERTER

Assuming that the load inductor and output filter capacitor are much greater than the resonant inductor L_r and resonant capacitor C_r the equivalent circuit of the system during each switching cycle can be reduced as shown in Fig. 5. During energizing mode (mode 1), the resonant capacitor voltage is clamped to source side line-line voltage V_s and the inductor current increases linearly as

$$I_L = I_L(t_0) + \frac{V_S}{L_r}(t - t_0) \tag{8}$$

$$V_C = V_O \tag{9}$$

When the inductor current reaches the desired value I_{Lpeak}, S_{11} and S_{15} are turned off. During partial-resonant mode (mode 2), the partial-resonant link $L_r C_r$ tank resonates partially at an isolated state from other circuits. The capacitor voltage and inductor current are given by

$$I_L = I_L(t_1)\cos\omega(t - t_1) + \frac{V(t_1)}{\omega_r L_r}\sin\omega(t - t_1) \tag{10}$$

$$V_C = V_C(t_1)\cos\omega(t - t_1) - \omega_r L_r I_L(t_1)\sin\omega(t - t_1) \tag{11}$$

V_C decreases via zero down to V_O, thus ensuring zero voltage turn on condition of on-coming switches $S_{22}S_{24}$. When V_c becomes equal to V_O, the on-coming switches can conduct on like diodes if the switch gating signals are applied in advance. Thus on-current transient such as spike currents can be eliminated. During de-energizing mode (mode 3) the resonant capacitor voltage is limited to output voltage V_O. The resonant inductor current is de-energized by the output voltage. The capacitor voltage and inductor current are given by

$$I_L = I_L(t_2) - \frac{V_o}{L_r}(t - t_2) \tag{12}$$

$$V_C = V_O \tag{13}$$

When I_c reaches the predetermined minimum current, $S_{22}S_{24}$ are turned-off. During mode 4, all of the switches are off-state like mode 2 and the capacitor voltage and inductor current are given by

$$I_L = I_L(t_3)\cos\omega(t - t_3) + \frac{V(t_3)}{\omega_r L_r}\sin\omega(t - t_3) \tag{14}$$

$$V_C = V_C(t_3)\cos\omega(t - t_3) - \omega_r L_r I_L(t_3)\sin\omega(t - t_3) \tag{15}$$

When V_C reaches to the equal value of the next on-coming input line-line voltage, $S_{11}S_{15}$ can turn on like diode if pre-triggering signals are given and the next switching cycle starts

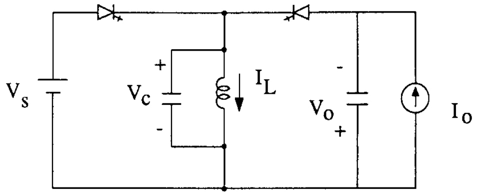
In mode 4 the resonant capacitor voltage has two peak values which become equal if neglecting ESR losses in the $L_r C_r$ tank as shown in Fig 2. The peak voltage is given by

$$V_{Cpeak} = \sqrt{Z_r^2 I_L^2(t_3) + V_C^2(t_3)} \tag{16}$$

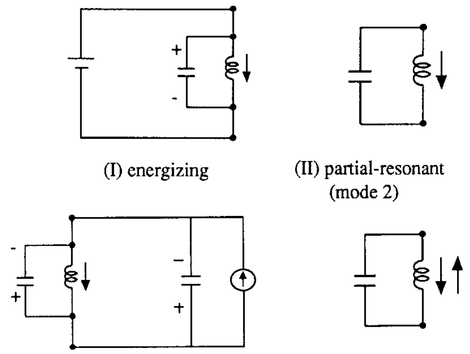
where

$$Z_r = \sqrt{L_r / C_r} \tag{17}$$

and can be plotted as shown in Fig. 6. The resonant peak voltage can be clamped to a certain value by properly controlling $I_L(t_3)$, the predetermined minimum current $I_{\min}$, so that the low switch voltage stress is limited to as low as 1.2 V_{sp} without adding any clamping circuit.



(a) Equivalent circuit.



(I) energizing

(II) partial-resonant (mode 2)

(III) de-energizing

(IV) partial-resonant (mode 4)

(b) Mode diagrams.

Fig. 5 (a) Equivalent circuit, (b) mode diagrams of the proposed converter during one switching cycle T_S ($T_S \ll T_O$ (fundamental output period)).

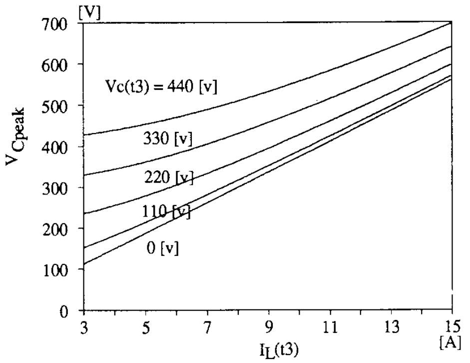


Fig. 6 Peak values of resonant capacitor voltage as a function of $I_L(t_3)$ for several $V_C(t_3)$.

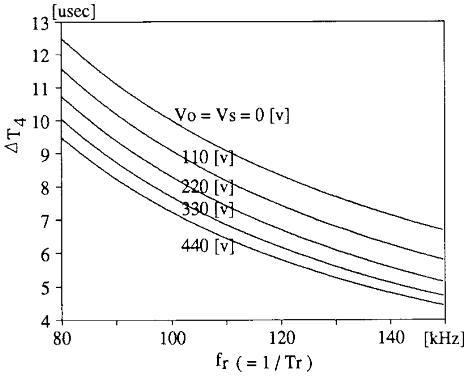


Fig. 7 The duration ΔT_4 of mode 4 with respect of $f_r(=1/T_r)$ for several V_O and V_S at $I_L(t_3) = 12[A]$.

The duration of mode 4 can be expressed as

$$\Delta T_4 = t_4 - t_3$$

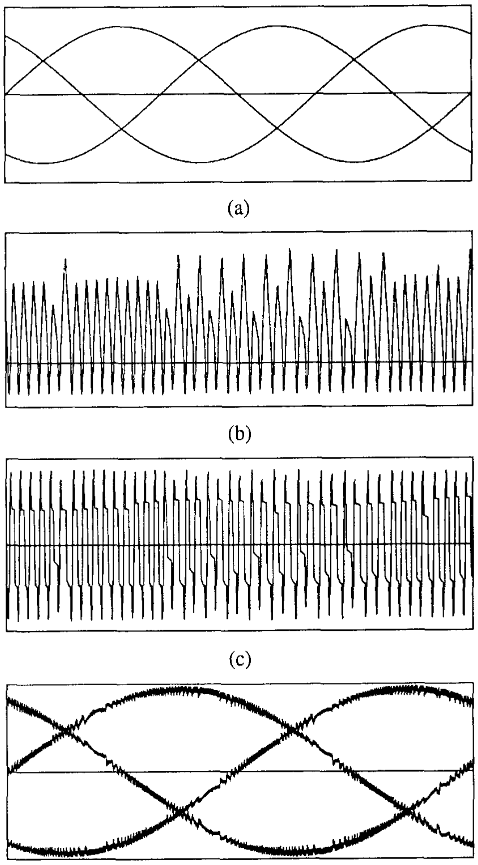
$$= T_r\left(1 - \frac{1}{2\pi}\tan^{-1}\frac{V_O}{\omega_r L_r I_{\min}} - \frac{1}{2\pi}\tan^{-1}\frac{V_S}{\omega_r L_r I_{\min}}\right) \tag{18}$$

where

$$T_r = 2\pi\sqrt{L_r C_r} \tag{19}$$

It can be plotted as shown in Fig. 7 as a function of f_r (= $1/T_r$) for several V_O and V_S. Fig. 8 shows the ratio of $\Delta T_4/T_S$ with respect to V_O/V_S for several output currents I_O. The ratio varies smoothly when V_O/V_S increase from buck region to boost region whereas when the output current increases from 4[A] to 14[A] it reduces so much. Consequently, the ratio of ΔT_4 to T_S decrease so rapidly depending on the increasing output power.

Fig. 9 shows the ratio of I_{Lpeak} to I_O as a function of V_O/V_S for several output currents I_O. The ratio is relatively larger for smaller range of I_O because ΔT_4 has relatively large portion of T_S as shown in Fig. 8. The ratio, however, decreases with increasing I_O and thus the peak values of resonant inductor current are slightly larger than the output current I_O but not too much. Consequently, the VAR rating of the resonant elements and switch current stresses become small.



(a)

(b)

(c)

(d)

Fig. 10 Typical waveforms of proposed converter (a) three input line-line voltages, (b) resonant inductor current I_L, (c) resonant capacitor voltage V_C and (d) three output line-line voltages.

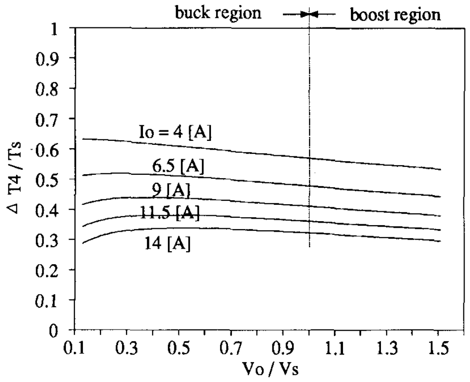


Fig. 8 The ratio $\Delta T_4/T_S$ of the duration of mode 4 to one switching cycle versus V_O/V_S for several output load currents I_O at $V_S = 310[V]$.

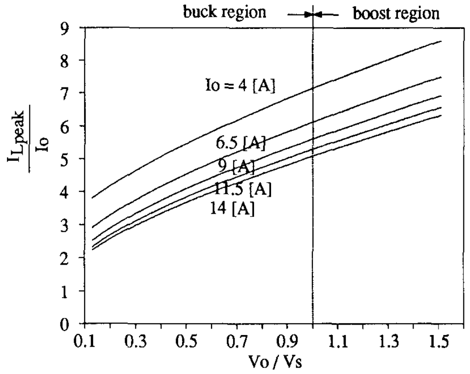


Fig. 9 The ratio of I_{Lpeak} to output load current I_O as a function of V_O/V_S as a function of output load current I_O at $V_S = 310[V]$.

V. SIMULATION RESULTS

In order to verify the operation principle and predicted features, the proposed converter of Fig. 1 is simulated with simple R-L load at 220 [V], 60 [Hz] voltage source. The parameters used here are as follows :

L_r = 60 μH
C_r = 150 nF
C_f = 10 μH
$P_{full\ load}$ = 2 $[kW]$

Switching frequency ranges over 10 [kHz]. Fig. 10 shows typical waveforms of resonant inductor current I_L, resonant capacitor voltage V_C, three output and input line-line voltages. Fig. 11 shows one waveform of three output line-line voltages in the case of abrupt change of output voltage reference from 150 [V] to 300 [V] when the input line-line voltage is 220 [V]. It illustrates good output waveform and buck-boost operation of the proposed converter. Fig. 12 and Fig. 13 also show one waveform of three output line-line voltages when the output frequency changes abruptly from 30 [Hz] to 60 [Hz] and when the output power changes from 1 [kW] to 2 [kW], respectively. Fig. 13 shows a slightly larger deviation from its references as the load power is increased because it is operated without any feedback from three phase load currents. As shown in Fig. 11, 12 and 13 the proposed converter has fast system response and high spectral performance, and doesn't show any instability problem.

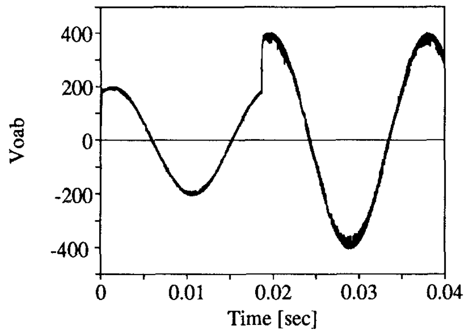


Fig. 11 One of three line-line output voltages in case of abrupt voltage reference from 150 [V] to 300 [V] at $P_0 = 2[kW]$

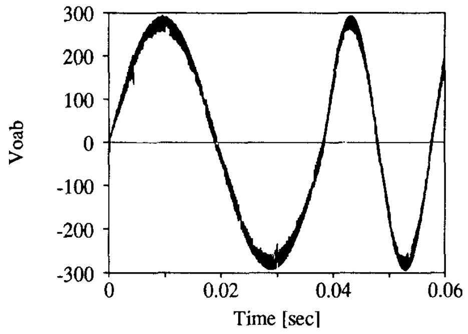


Fig. 12 One of three line-line output voltages in case of abrupt frequency change from 30 [Hz] to 60 [Hz] at $P_0 = 2[kW]$

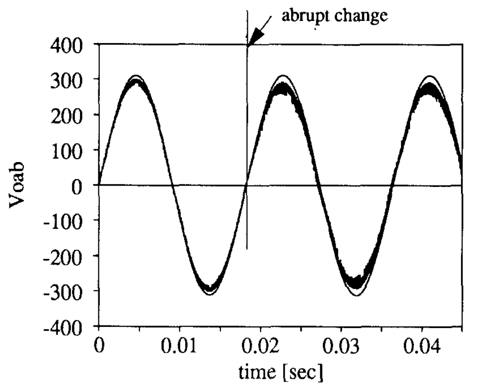


Fig. 13 One of three line-line output voltages in case of abrupt load change from 1 [kW] to 2 [kW] at $f_0 = 60[Hz]$

VI. CONCLUSION

In this paper a new zero voltage switching partial-resonant link AC/AC converter is proposed. The new scheme has very simple structure having only 12 unidirectional devices. Partial-resonant operation happens only for the duration of short switching transients and results in low VAR ratings of the resonant elements, low dissipations in the capacitor and inductor which usually have effective series resistances. Further, by properly controlling the partial-resonant mode (mode 4) the switch voltage stress can be limited to almost supply voltage even without adding any clamping circuits.

The proposed converter can easily control current pulse amplitude, which enables the converter to improve the output waveforms easily without causing any oscillation problems due to output filter capacitor and load inductance. Symmetrical arrangement of the power circuit provides fully regenerative operation and buck-boost capability. The features of the proposed converter can be summarized as follows:

1) simple structure with reduced number of switches,
2) easy current pulse amplitude control,
3) low VAR ratings and low dissipations of resonant elements,
4) low switching loss,
5) low device and component stress
6) buck-boost operation and bilateral power flow capability,
7) fast system response and high resolution output waveforms,
8) no instability problem,
9) simple control and implementation.

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