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Analysis of the two-quadrant converter with rectifier with near sinusoidal input currents

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Abstract: A new topology for a two-quadrant converter is presented. In the AC/DC transfer mode, the converter works as a rectifier with near sinusoidal input currents, whereas in the DC/AC transfer mode, it works as a square-wave pulse switching inverter. The converter is characterised by smaller power losses and reduced electromagnetic interference problems. Possible applications in adjustable speed drive with regenerative braking, wind energy conversion systems and small hydro interconnections with induction generators are also discussed.

Nomenclature

L	inductance
C	capacitance
ω	angular frequency
I	instantaneous current
V	instantaneous voltage
V_{m1}	amplitude of supply phase voltage
V_{m2}	amplitude of converter phase voltage
V_d	DC link voltage
$\underline{x}$	complex vector
P	active power
α	phase angle between v_{U0} and v_{R0}
$I_{(1)}$	amplitude of fundamental supply current
R, S, T	phases of the three-phase system
a	active
ref	reference

1 Introduction

Many applications use three-phase converters for two-quadrant operation in AC power supplies where the objective is to produce sinusoidal current waveforms on the AC side. For example, in motor drives with regenerative braking, the power flow through the utility interface converter reverses during the regenerative braking, whereas the kinetic energy associated with the inertias of the motor and load is recovered and fed back to the utility system [1–3]. In order to reduce the higher current harmonics on the AC side, usually, the three-phase converters for two-quadrant operation use pulse width modulation (PWM) switching. Governments and international organisations have introduced new standards (in the USA, IEEE 519, and in Europe IEC 61000-3), which limit the harmonic content of the currents drawn by rectifiers from the power line [4–6].

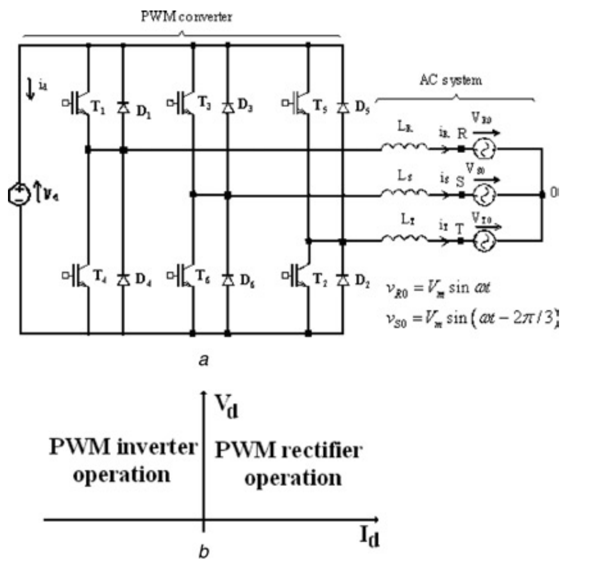

Figure 1 *Converter for two-quadrant based on PWM principle*

a Configuration
b Operation modes

Fig. 1*a* presents the most popular topology used in adjustable speed drives, uninterruptible power supplies (UPS) and, more recently, in PWM rectifiers. This topology has the advantage of using a low-cost three-phase module with a bi-directional energy flow capability. Because of the rapid changes in voltages and currents of a switching converter, a PWM rectifier is a source of electromagnetic interference (EMI). The PWM rectifier, even though it has near sinusoidal input currents, has important drawbacks as compared with the inexpensive three-phase rectifiers with diodes: larger switching losses and high per-unit current rating [1–2].

In [7], a two-quadrant converter with rectifier with near sinusoidal input currents (RNSIC) is introduced. When the power transfer is made from AC to DC, the converter operates as RNSIC. If the transfer is reversed, a PWM inverter connected to the utility grid starts working.

In what follows, a new solution for two-quadrant converter using RNSIC is presented.

2 New converter for two-quadrant operation

The proposed new converter for two-quadrant is equipped with six transistors [e.g. insulated gate bipolar transistor (IGBT)] having square-wave pulse switching (that is not PWM) operation, as shown in Fig. 2*a*. When the energy is transferred from the AC side to DC side, the transistors are off and the converter works as an RNSIC, as shown in Fig. 3 [7–9]. When the energy is transferred from the DC side to the AC side, the transistors are controlled to conduct for θ angles (square-wave pulse switching) and the converter works as inverter, as shown in Fig. 2*b*.

This is a much more convenient solution when compared with that in [7] in which we have presented a static frequency converter having in input a combined two-quadrant converter. Apart from the fact that the combined converter, which included an RNSIC, had the transistors operating according to the PWM principle, it had also three supplementary switches, ($S_1 - S_3$), being so more complex.

2.1 AC/DC operation mode

Fig. 3 shows an AC/DC converter generating reduced higher-order current harmonics in the mains, called for short in what follows RNSIC, and which is a module of the two-quadrant converter in Fig. 2*a*. The capacitors $C_1 - C_6$ have the same value C and they are DC capacitors. Inductors L_R, L_S and L_T have the same value, denoted by L_1, and they are connected on the AC side. L_1 and C fulfil the condition

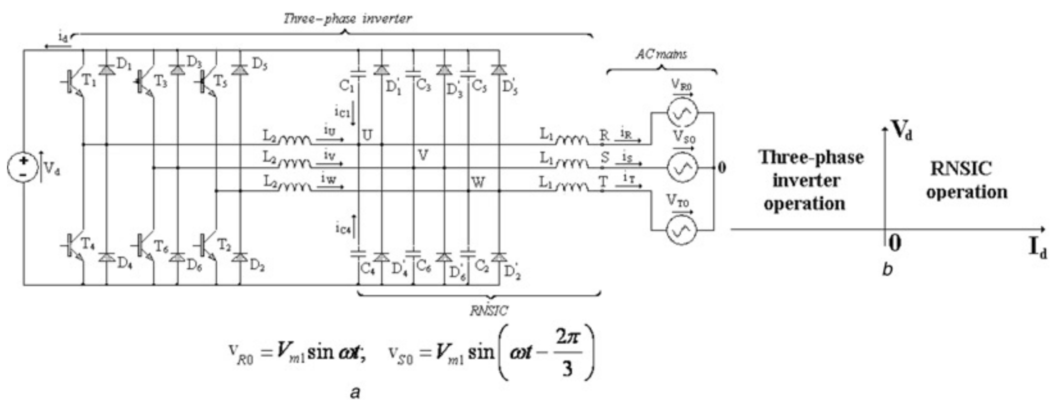

Figure 2 *Converter for two-quadrant with RNSIC*

a Configuration
b Operations

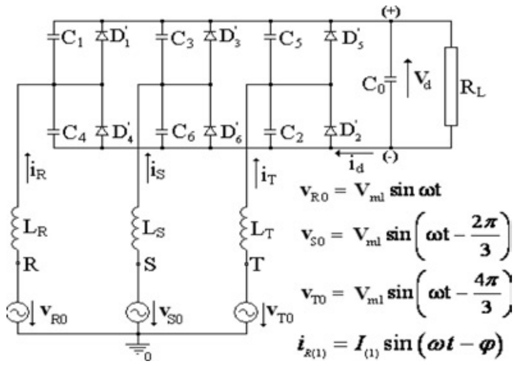

Figure 3 *Configuration of RNSIC converter*

$0.05 \le L_1 C\omega^2 \le 0.10$ in order for the phase currents i_R, i_S, i_T to be practically sinusoidal (ω denotes the mains angular frequency), [7–9].

Considering that the currents i_R, i_S, i_T are practically sinusoidal and have the amplitude $I_{(1)}$ depending on the load resistor R_L, the mean value of the current I_d can be calculated as [7]

$$I_d = \frac{3\,I_{(1)}}{2\,\pi}\left(1 + \cos\,\omega\,t_1\right) \tag{1}$$

where ωt_1 is the angle of turning on for the diodes $D_1' - D_6'$.

There are two extreme cases during RNSIC converter functioning. In the first case, if $R_L = 0$ (and so, $V_d = 0$ and $\omega t_1 = 0$), the capacitors C_1–C_6 are short-circuited and the angle $\varphi = +90^\circ$ is inductive. In this case, the phase currents are sinusoidal and have maximum amplitude, equal to I_{max}. In the second case, if the voltage V_d exceeds the value $\sqrt{3}V_{m1}/(1 - 2L_1C\omega^2)$, the diodes $D_1' - D_6'$ do not conduct any more and the angle $\varphi = -90^\circ$ is capacitive (and so $R_L = \infty$ and $\omega t_1 = \pi$). For this latter case, the phase currents are also sinusoidal and the amplitude has a minimum value I_{min}, referred to as the holding current. The ratio I_{min}/I_{max} has the value

$$\frac{I_{min}}{I_{max}} = \frac{2\,L_1 C\omega^2}{1 - 2\,L_1 C\omega^2} \tag{2}$$

Fig. 4*a* shows the variation of the angle φ, the phase displacement angle between the phase voltage and the fundamental of the phase current, as a function of the mean rectified voltage V_d rated to the reference value $V_{ref} = 3\sqrt{3}V_{m1}/\pi$ specific for the classical three-phase rectifier [1]. The voltage V_d can be set at a certain value by the load current.

Fig. 4*b* shows the variations of the output voltage V_d rated to the reference value V_{ref} and the amplitude of the phase current $I_{(1)}$ rated to the reference value I_{max} as a function of the ratio R_L/R_{Lr} (R_{Lr} denotes the rated load resistor for $\varphi = 0^\circ$).

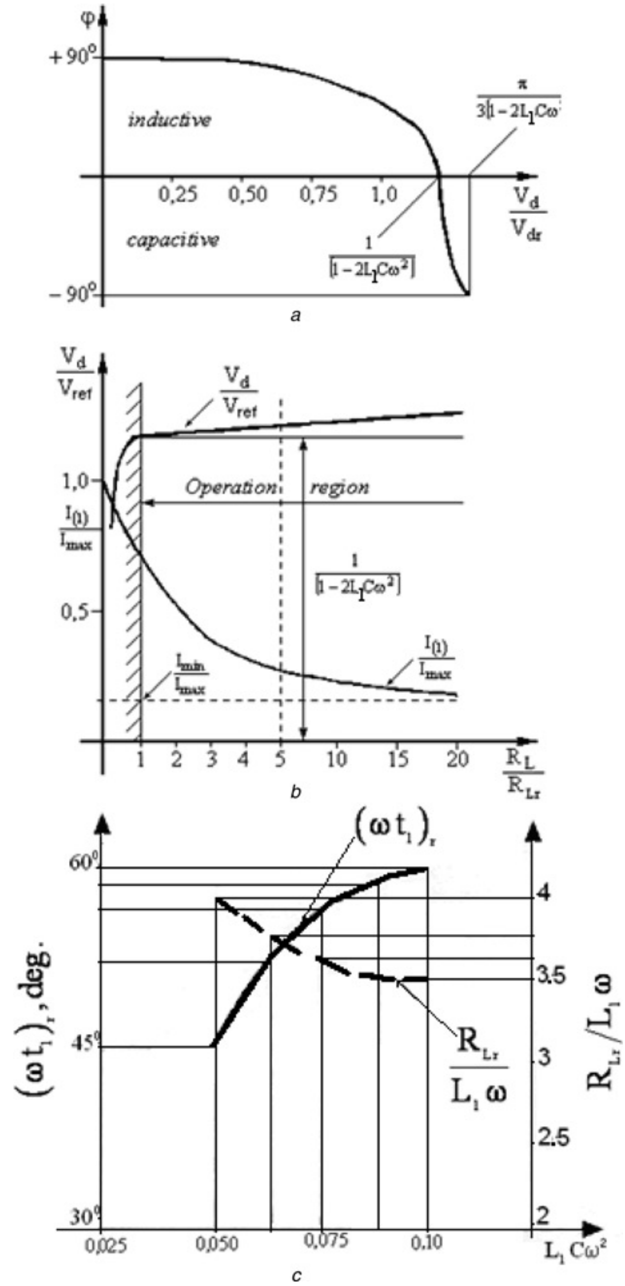

Figure 4 *Characteristics of RNSIC converter*

a Angle φ as a function of ratio V_d/V_{ref}
b Ratio V_d/V_{ref} and $I_{(1)}/I_{max}$ as a function of R_L/R_{Lr}
c Rated angle $(\omega t_1)_r$, and the ratio $R_{Lr}/L_1\,\omega$ as a function of $L_1 C\omega^2$

The rated operation of the RNSIC converter is defined for $\varphi = 0^\circ$ and $R_L/R_{Lr} = 1$. For this case, the variations of the rated angle $(\omega t_1)_r$, the angle corresponding to the beginning of diodes conduction and the ratio $R_{Lr}/L_1\,\omega$ are given in Fig. 4*c*, as a function of the parameter $L_1\,C\,\omega^2$. The interval between 45° and 60° for $(\omega t_1)_r$ ensures a reduced content of higher harmonics for the input currents. One can design this converter using the diagrams in Fig. 4*c* [7].

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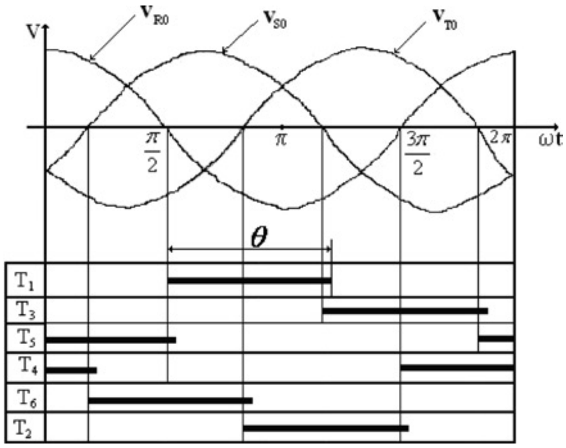

Figure 5 *Control programme of the transistors*

It is important to notice that, considering the load resistor as infinite, the classical three-phase rectifier with diodes and passive filters has a holding (idle) current $I_{\min}$ because of the presence of passive filters at its entry, which determines capacitive load behaviour. This holding current is larger than that resulted at the RNSIC converter depicted in Fig. 3.

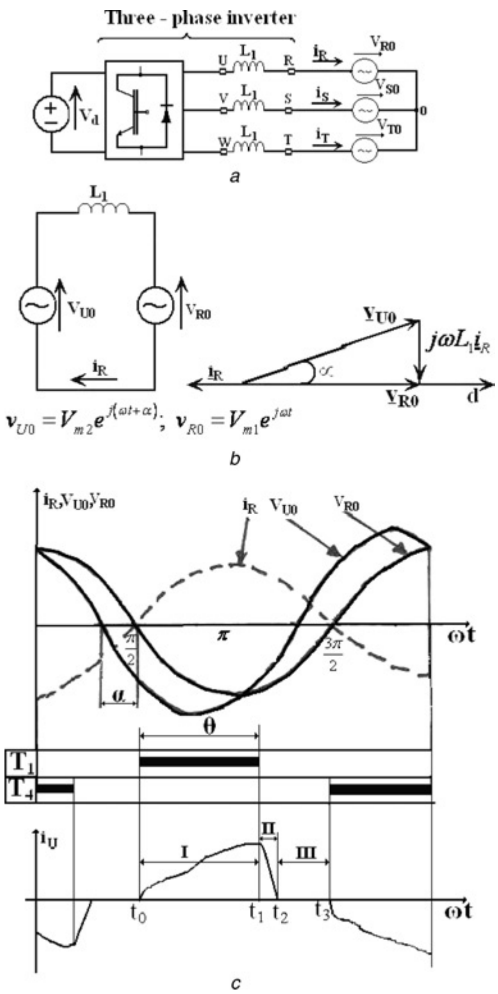

Figure 6 *Inverter operation mode of the two-quadrant converter with an RNSIC*

a Simplified representation
b Phasor diagram at unity power factor
c Waveforms of the currents and voltage

2.2 DC/AC operation mode

In what follows, we describe the operation of the converter shown in Fig. 2*a* as an inverter. The control programme of the transistors is shown in Fig. 5 and the corresponding waveforms in Fig. 6. During the first stage, which starts at t_0, the transistor T_1 begins to conduct and capacitor C_1, charged at initial voltage $V_{\rm in}$, is discharged to final voltage, $V_{\rm end}$, whereas capacitor C_4, initially charged at voltage $(V_{\rm d} - V_{\rm in})$ is charged to $(V_{\rm d} - V_{\rm end})$ by the help of the oscillatory processes in which transistor T_1 and inductor L_2 take part. After the blocking of transistor T_1, made at t_1, the second stage begins, when the energy accumulated in inductor L_2 is rapidly transferred to DC and AC sources through the diode D_4. Finally, in the third stage, which lasts between t_2 and t_3, the current $i_{\rm U}$ is zero, and the current $i_{\rm R}$ has a practically sinusoidal waveform, flowing through capacitors C_1 and C_4. At the end of this stage, capacitor C_4 is charged at voltage $V_{\rm in}$, and C_1 at voltage $(V_{\rm d} - V_{\rm in})$. Inductor L_2 has values two times smaller than L_1. For the case of operation in inverter mode, the voltage $V_{\rm d}$ is considered to be 15–25% greater than for the case of rectifier system operation. Diodes D_1'–D_6' are chosen according to the RNSIC component design specification, whereas diodes D_1–D_6 are rated for much smaller average currents.

According to the phasor diagram in Fig. 6*b*, current $i_{\rm R}$ is given by [2]

$$i_{\rm R} = \frac{v_{U0} - v_{\rm R0}}{j\omega L_1} \tag{3}$$

whereas its active value, $i_{\rm Ra}$, is given by

$$i_{\rm Ra} = \frac{V_{m2}}{\omega L_1}\left[\sin(\omega t + \alpha) - \frac{V_{m1}}{V_{m2}}\sin\omega t\right] \tag{4}$$

The active power transferred to the AC source is given by

$$P = \frac{3}{2\pi}\int_0^{2\pi} i_{\rm Ra} V_{\rm m1} \cos\omega t \,{\rm d}\omega t = \frac{3V_{m1}V_{m2}}{2\omega L_1}\sin\alpha \tag{5}$$

In order to obtain a unitary power factor at the AC source, it results from (4) that

$$\cos\alpha = \frac{V_{\rm m1}}{V_{\rm m2}} \tag{6}$$

It results that the value of the power transmitted to the AC source could be varied by modifying the amplitude V_{m2} (thus, the angle θ and the voltage V_d) and the angle α, as shown in Fig. 5.

The switch of the converter in Fig. 2*a* from the inverter operation mode to the rectifier operation mode and reverse can be rapidly accomplished during a utility grid cycle $T = 2\pi/\omega$.

Obviously, the converter shown in Fig. 2*a* can operate as controlled rectifier with power factor correction. In this case, the energy transfer is made from the AC to the DC side, by adequately controlling the transistors. The phase currents i_R, i_S and i_T are in phase with the voltages v_{R0}, v_{S0} and v_{T0}. These currents are practically sinusoidal and their amplitudes depend on the voltage V_d and on the way the transistors are controlled.

2.3 Possible applications

Possible applications of the converters for two-quadrant operation with RNSIC are their usage in static frequency converters with DC voltage link, designed for supplying variable voltage and frequency to the three-phase induction motor drives, as shown in Fig. 7.

For the time intervals when the induction motor drive is in the motoring regime, the input converter becomes an RNSIC converter. For this case the transistors T_1-T_6 are off. The output switch-mode converter operates as a PWM inverter [10–13]. The energy is transmitted from the power supply to the motor and the voltage on the filtering capacitor C_0 is less than $V_{d0} = \sqrt{3}\,V_{m1}/(1 - 2L_1C\omega^2)$.

During the time interval while the induction machine (IM) is operating in braking mode, the energy received from the motor is transmitted to the power supply. The switch-mode converter operates as a rectifier and the voltage across C_0 is greater than $(15\%-25\%)V_{d0}$. Furthermore, the energy is transmitted into AC mains by means of a three-phase inverter made up of transistors T_1-T_6, three inductors L_2, diodes D_1-D_6 and RNSIC. One must observe also the fact that the total duration of operation as a generator for the

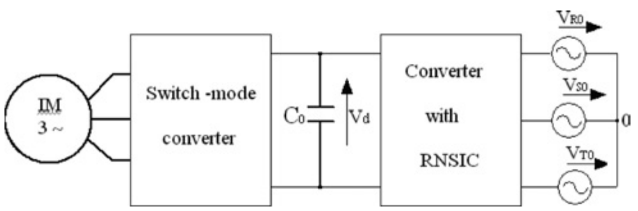

Figure 7 *Static frequency converter for two-quadrant with RNSIC*

asynchronous machine is much smaller when compared with the total motor functioning duration.

Having in view that the wind energy is non-polluting, a significant growth of its importance in electrical power generation is expected in the future. Although remarkable technological progress has been achieved in the last years, currently, the power produced according to this principle is not yet totally competitive from the economic point of view. A wind-energy-conversion system has to draw the maximum power when the speed of wind varies over a wide range [1, 2]. The energy produced by the generator is transmitted in the distribution grid by means of a frequency converter with DC voltage link. Fig. 8 presents a new variable-speed wind-energy-conversion system. The energy produced by the generator is transmitted in the distribution grid by means of a frequency converter made of an RNSIC converter, a capacitor C_0 and a square-wave pulse switching inverter. Capacitors $C_1'-C_6'$ have the capacity equal to C', whereas capacitors $C_1''-C_6''$ have the capacity equal to C''. Inductors L_u, L_v and L_w have the same inductance L and include the leakage inductance of the stator of the induction generator. Denoting by $\omega_{s\,max}$ and $\omega_{s\,min}$ the maximum and minimum values of the stator angular frequency ω_s, one must fulfil the following relations

$$0.05 \le LC'\omega_{s\,max}^2 \le 0.10 \qquad (7)$$

$$0.5 \le L(C' + C'')\omega_{s\,min}^2 \le 0.10 \qquad (8)$$

in order to ensure practically sinusoidal stator currents i_u, i_v and i_w. For the U phase, for example, the voltage is expressed as $v_{U_o} = V_1 \sin \omega_s t$ and the fundamental of the corresponding phase current is $i_{U(1)} = I_{(1)} \sin(\omega_s t - \varphi)$. The angle φ is positive and so the RNSIC converter behaves as a resistive-capacitive load for the induction generator.

In Fig. 9, we present the induction generator self-excitation process based on the existence of a remnant flux in the rotor [14]. In order to ensure the necessary magnetisation current $I_m = I_{(1)} \sin \varphi$, for small values of ω_s, one operates with the switches S_1-S_3 closed, and for large values of ω_s, the switches are open. In this way, the variation domain of the phase voltages v_{U0}, v_{V0} and v_{W0} is reduced. One can observe that the rms value of the current that flows through the switches S_1-S_3 is $<6-8\%$ of $I_{(1)max}$ and so it is not necessary to design them for large rated currents. Of course, for the capacitive part of the RNSIC converter, one must adopt more than two sections for the case of large variations of ω_s. In Figs. 10*a* and 10*b*, the characteristics of the RNSIC converter for this application are depicted. For small

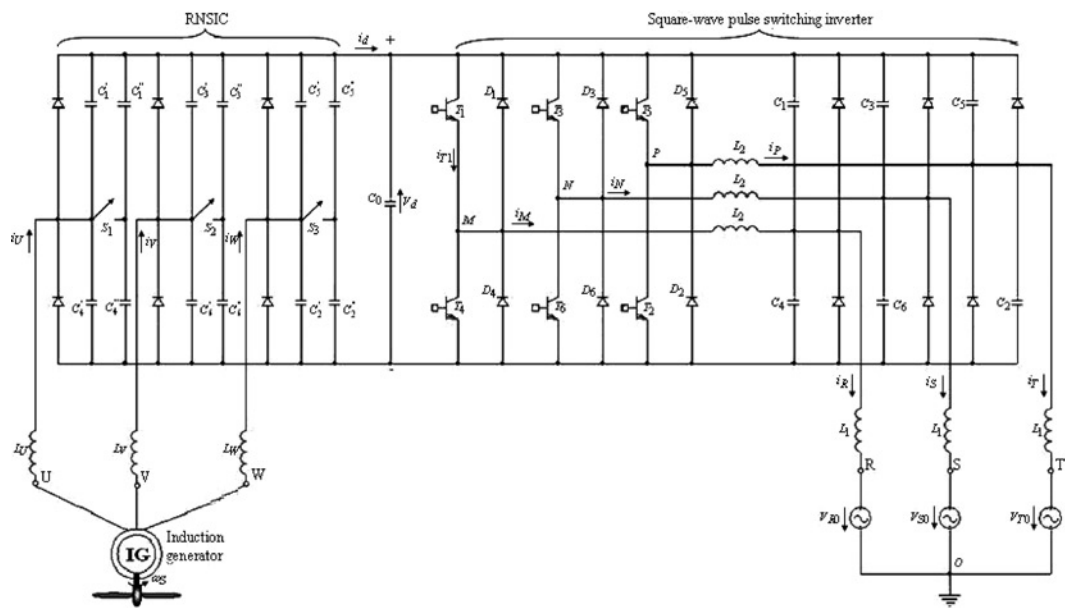

Figure 8 *System of wind/hydro generator with square wave pulse switching inverter*

values of V_1, the RNSIC diodes are off and the magnetising current I_m varies linearly with V_1 according to

$$I_m = \frac{2\omega_s C_{eq} V_1}{(1 - 2LC_{eq}\omega_s^2)} \quad (9)$$

in which the equivalent capacity C_{eq} equals C' or $(C' + C'')$, depending on the state of the switches.

For this case, when the diodes are off, the RNSIC converter represents a pure capacitive load for the induction generator. The diodes begin conducting when the voltage V_1 reaches the value $(1 - 2LC_{eq}\omega_s^2)V_d/\sqrt{3}$. In the operating region, pointed out in Fig. 10*a* for two values of ω_s, the magnetisation currents I_m are practically constant as function of V_1 and the mean value of i_d, denoted I_d. When voltage V_1 exceeds the value $\pi(1 - 2LC_{eq}\omega_s^2)V_d/3\sqrt{3}$, the current I_m decreases quickly to zero, the RNSIC converter begins behaving as a resistive load and the induction generator de-excites itself. In the operation regions, the current I_d increases from zero to a maximum value depending on ω_s, V_1 and the load resistor R_L, as shown in Fig. 10*b*.

The output inverter which connects to the grid can be conveniently chosen according to the variant shown in Fig. 2*a*, having the semiconductor devices

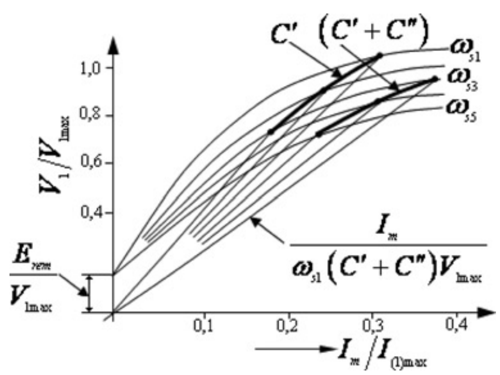

Figure 9 *Variations ratio V_1/V_{1max} as a function of $I_m/I_{(1)max}$ for different value*

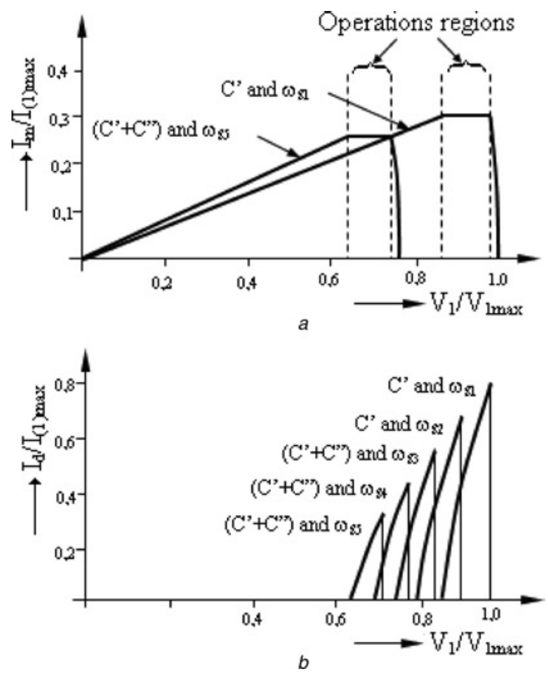

Figure 10 *Characteristics of RNSIC converter as a function of ratio V_1/V_{1max}*

a Ratio $I_m/I_{(1)max}$
b $I_d/I_{(1)max}$

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controlled over continuous time intervals. Such an inverter can be adopted also for small hydro interconnections with induction generators. Slightly modifying the angles α and θ, as shown in Fig. 6*c*, one can vary the active power delivered in the distribution grid with unity power factor.

3 Experimental and simulations results

Laboratory experiments and simulation results have proved the effectiveness of the proposed converter shown in Fig. 2*a*. The laboratory prototype consists of a three-phase voltage source (with $V_{m1} = 311$ V and $f = 50$ Hz), an RNSIC converter and a three-phase module with IGBTs and diodes. The RNSIC is composed of six diodes, three inductors L_1 with inductance 25 mH and six DC capacitors C_1-C_6 with capacitance 24 μF. For the three inductors L_2, we have adopted the value 12.5 mH.

Figs. 11 and 12 show the simulations results. In Figs. 11*a* and 11*b*, the waveforms of the phase current i_R and the DC current i_d are shown, for the case of the IM operating in motoring regime. Table 1 presents the important values obtained for the rectifier operation mode. Transistors T_1-T_6 are off and, at the RNSIC output, in parallel with C_0 (4000 μF), the load resistor R_L is connected. Table 1 gives the values of V_d, $I_{(1)}$, φ, the phase displacement angle between the phase voltage and the fundamental of the phase current, the total harmonic distortion (THD) factor for phase currents, as well as the ratio $I_{(5)}/I_{(1)}$ between the fifth harmonic and the fundamental $I_{(1)}$. It is known that the RNSIC converters have the fifth harmonic as the largest input harmonic [8–9].

For larger load currents I_d, the amplitude of the fundamental harmonic current $I_{(1)}$ is increased, the ratio $I_{sc}/I_{(1)}$ can be reduced (e.g. $<20\%$) and so, the THD of the phase currents has to be $<5\%$, according to the IEEE Standard 519/1992. I_{sc} denotes the amplitude of the short-circuit currents at the R, S and T terminals. For small load currents I_d, the amplitude of the fundamental harmonic current $I_{(1)}$ is reduced, the ratio $I_{sc}/I_{(1)}$ can be between 20 and 50 or 50 and 100 and so, the THD of the phase currents have to be $<8\%$ or 12% [1]. The optimal value of the parameter $L_1C\omega^2$, for acceptable RNSIC converter dimensions and THD current factor, is 0.06. The output rated power $P_{dr} = V_{dr}^2/R_{Lr}$ of an RNSIC converter, defined for $\varphi = 0$, is directly proportional to $C\omega$ and inversely proportional to $L_1\omega$.

For power circulation in the opposite direction, we consider the application shown in Fig. 7, implementing a static frequency converter with RNSIC. Fig. 12 shows the simulated waveforms of the phase current i_R, transistor currents i_{T1} and i_{T4}, diode currents $i_{D'1}$ and $i_{D'4}$, capacitor current i_{C1} and DC current i_d, when the IM is operating in braking mode. When i_{C1} is zero, only one of the diodes D_1' or D_4' is conducting. When transistors T_1 and T_4 are off, the current i_R does not flow instantly through capacitors C_1 and C_4, but after the current i_U becomes zero, it

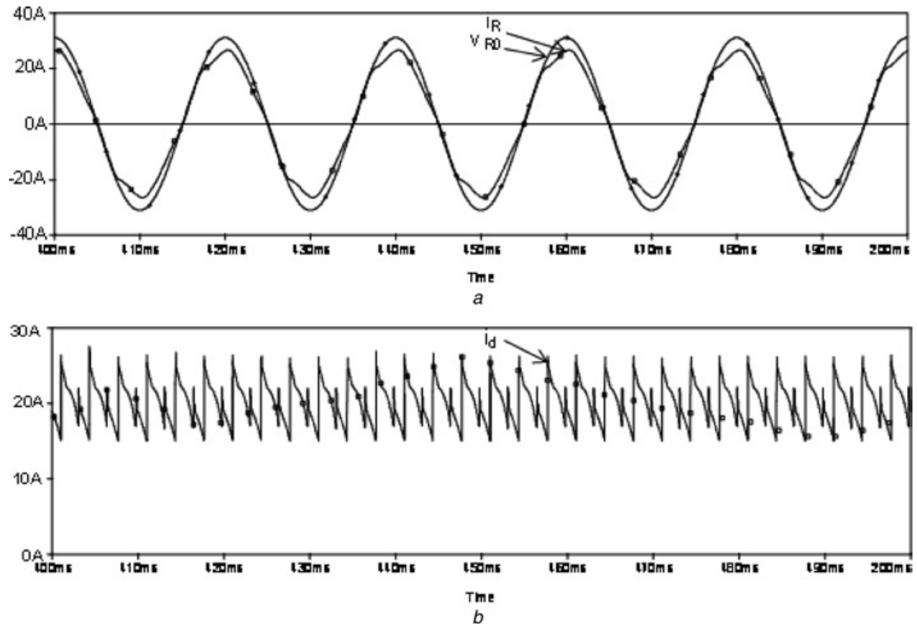
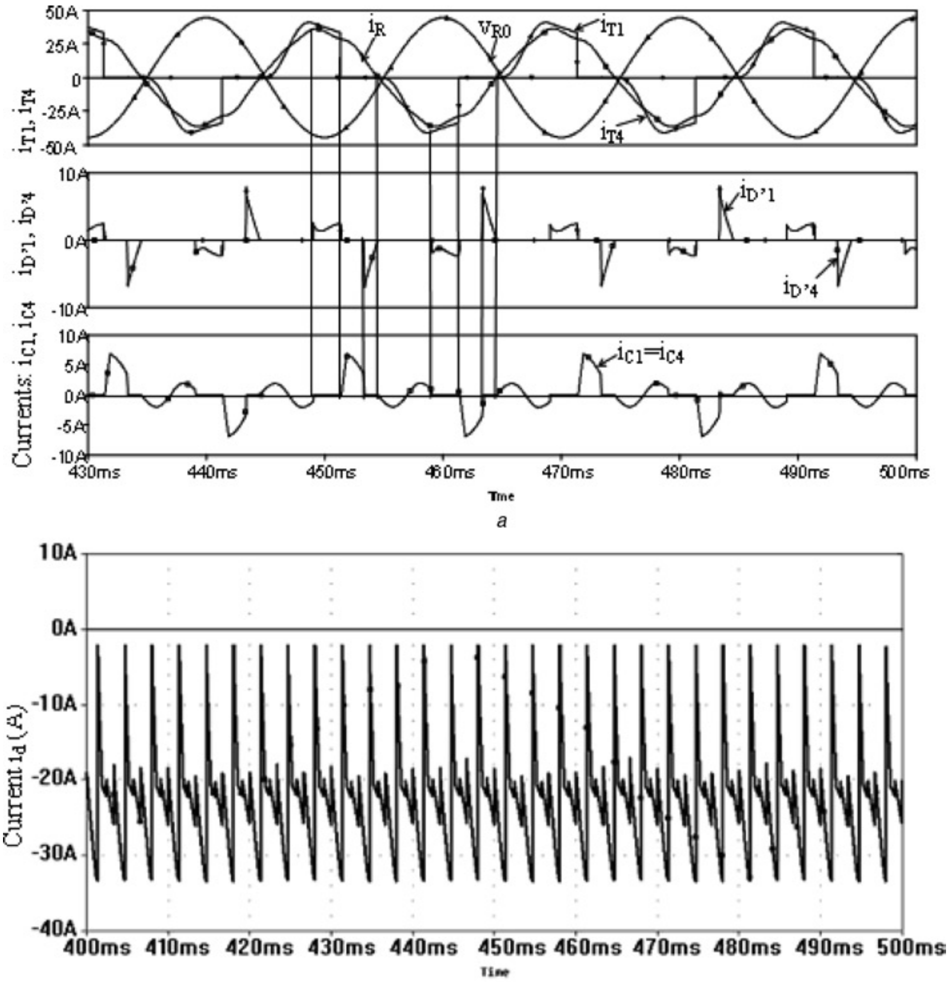


Figure 11 *Rectifier operation mode of the proposed converter with an RNSIC for $V_d = 600$ V*

a Waveforms of the phase current i_R and the phase voltage v_{R0} in steady state
b Waveform of the DC current i_d



a

b

Figure 12 *Inverter operation mode of the proposed converter with an RNSIC for $V_d = 800$ V*

a Waveforms of the phase current i_R, the transistor currents i_{T1} and i_{T4}' the phase voltage v_{R0}, currents $i_{D'1}$ and $i_{D'4}$ and current i_{C1}
b Waveform of the current i_d

flows through diode D_1 or D_4. From Fig. 12, one can deduce that, taking into account the waveforms of the currents i_{T1}, i_{T4}, $i_{D'1}$, $i_{D'4}$ and $i_{C1} = i_{C4}$, the waveform of the current i_R is practically sinusoidal.

The value of the power transmitted to the AC source can be varied by modifying the angles α and θ, as shown in Fig. 6. The amplitude V_{m2} of the inverter output voltage, which can be modified by the help of

Table 1 Rectifier operation mode, $L_1 = 25$ mH, C = 24 μF, $V_{m1} = 311$ V, $f = 50$ Hz, $P_{dr} = 12$ kW

R_L, Ω	V_d, V	$I_{(1)}$, A	φ, º	THD, %	$I_{(5)}/I_{(1)}$, %
20	526	31.4	+19.8	4.05	3.75
30	590	25.0	+1.8	5.15	4.92
40	612	20.7	−10.8	5.53	5.30
70	622	14.1	−32.4	5.41	5.15
100	624	11.3	−43.2	4.87	4.48
200	629	8.0	−59.4	5.23	4.85
600	648	6.02	−72.0	5.81	5.65
5 k	688	5.36	−88.2	1.77	1.64
50 k	702	5.32	−90	0.24	0.17

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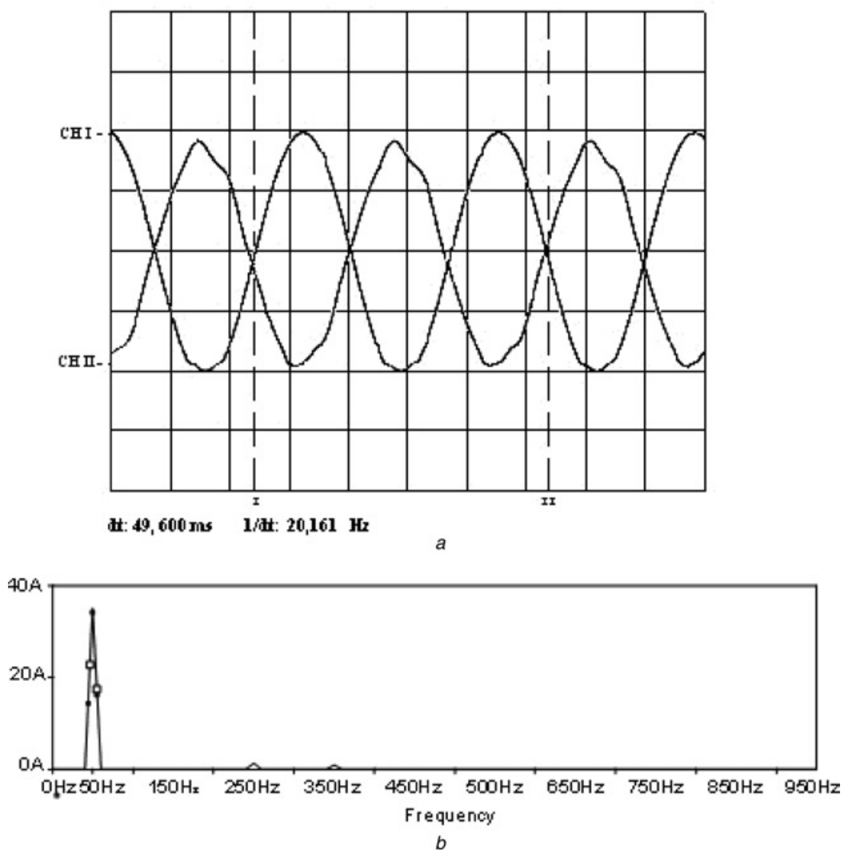

Figure 13 *Experimental waveforms of phase current i_R and voltage v_{R0} for inverter operation mode with $V_d = 800$ V*

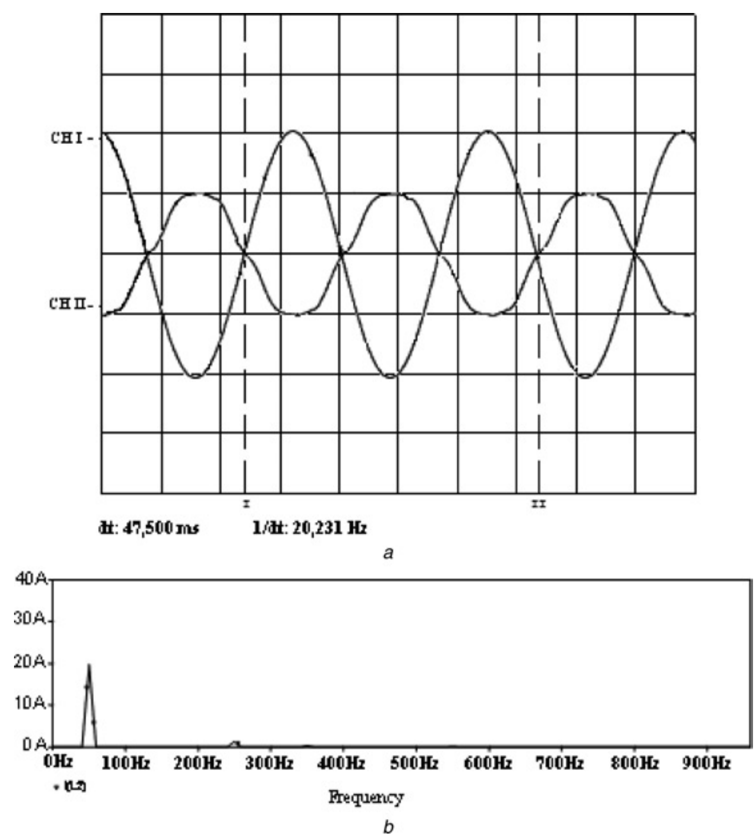

Figure 14 *Experimental waveforms of phase current i_R and voltage v_{R0} for inverter operation mode with $V_d = 700$ V*

the transistors command angle θ, is larger than that obtained from a PWM inverter fed with the same voltage V_d, which is an advantage of the proposed converter.

Figs. 13 and 14 show the experimental waveforms for the phase current i_R for two different values of the power transmitted to the AC source, when the converter operates as inverter, according to Fig. 2*a*. In Fig. 13, the transmitted power is 16.3 kW, equal to that obtained by simulation, as shown in Fig. 12*a* for $V_d = 800$ V and angle $\theta = 144^\circ$. The experimental waveforms are similar to those obtained by simulation. For this case, the only significant harmonics are those of order 5 and 7, which represent 3.65% and 2.45% of the fundamental line current, respectively.

In Fig. 14, the transmitted power is 9.28 kW for $V_d = 700$ V and angle $\theta = 126^\circ$. In this case also, the only significant harmonics are those of order 5 and 5, which represent 6.07% and 2.19% of the fundamental line current, respectively

Experimental tests have shown that the voltage V_d can be reduced practically till the limit of 600 V in order for the two-quadrant converter to operate still as inverter, delivering into the AC mains 6.28 kW. Of course, in this case, the THD factor has a high value of 8.66%, but this drawback is admissible because of the large value of the ratio $I_{sc}/I_{(1)}$. It is interesting to emphasise that, for $V_d = 600$ V, if the transistors T_1-T_6 are off, the converter operates in rectifier regime as RNSIC.

The transition from rectifier to inverter regime and reverse is very fast, during a semi-period $T/2 = \pi/\omega$. This fact is an advantage in the case when the converter, according to Fig. 2*a*, is used in UPS systems.

4 Conclusions

Comparing the two-quadrant PWM converter with the converter proposed in this paper, one can make the following considerations.

- The proposed topology reduces the commutation losses and EMI problems. Transistors T_1-T_6 begin to conduct only once in a utility grid cycle, at zero currents i_U, i_V and i_W (i.e. according to zero current switching principle).

- One of the advantages of the continuous functioning of the controllable switches (in square-wave pulse and not PWM switching) is that each inverter switch changes its state only twice per cycle, which is important at high power levels where the solid-state switches generally have slower turn-on and turn-off speeds.

- The proposed converter has increased safety due to the fact that controllable switches have much smaller total conducting durations, being blocked while the converter operates as RNSIC.

- In the case of DC to AC conversion, for the same values of the voltage V_d and AC inductances, the proposed converter provides larger output voltages V_{U0}, V_{V0} and V_{W0} and thus allows a more efficient energy transfer (obviously, at the PWM inverter, the fundamentals of the output voltages are smaller and so the transferred energy is smaller).

The simulation and experimental results proved that the fifth current harmonic is the most significant one generated in the AC mains and that its value is within the limits imposed by the IEEE Standard 519/1992.

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