Topologies of three-phase rectifiers with near sinusoidal input currents

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Abstract: Several topologies of three-phase low-harmonic diode rectifiers equipped with inductors, capacitors and diodes, based on an original solution, are presented. Inductors and capacitors are used in conjunction with the three-phase diode rectifier bridge to improve the waveform of the currents drawn from the utility grid. For the basic variant of the three-phase rectifier, representative characteristics and design elements are presented. Starting from this configuration, an AC/DC converter with small holding current and a four-quadrant frequency converter are proposed. The operation of the proposed converters is analysed and, on this basis, design considerations are reviewed. Several possible applications of the converter variants are mentioned. Analytically obtained results have bee experimentally verified.

1 Introduction

In most power electronics applications, the input power supply is in the form of a 50 or 60 Hz sine wave AC voltage provided by the electric utility, that is eventually converted to DC. As power electronic system proliferate, AC-to-DC rectifiers are playing an increasingly important role. A large majority of power electronics applications, such as switching DC power supplies, AC motor drives, static frequency converters, DC servo drives and so on, use such uncontrolled three-phase rectifiers [1]. The three-phase, six-pulse, full-bridge diode rectifier is a commonly used circuit configuration. A filter (L_f C_f) is connected at the DC side of the rectifier. In the three-phase rectifier, shown in Fig. 1*a*, the AC side inductor is assumed to be zero and the DC side is replaced by a constant DC current I_d. The RMS harmonic components $I_{(n)}$ of the phase current can be determined in terms of the fundamental frequency component $I_{(1)}$ as: $I_{(n)} = I_{(1)}/n$, where n represents the harmonic order: $n = 5,7,11,13,\ldots$ To draw a conclusion, typical AC current waveforms in the three-phase diode rectifier circuits are far from a sinusoid, in accordance with Fig. 1*b*. The power factor is also very poor because of the harmonic content in the line current. Moreover, these harmonics cause additional harmonic losses in the utility grid and may excite electrical resonance, leading to large overvoltages [2].

A first alternative to reduce the current harmonics is the use of classical passive filters made up of LC series circuits, in accordance with Fig. 1*a*. However, passive filters have the following drawbacks [1, 2]:

(i) Filtering characteristics are strongly affected by the source impedance. As a result of the resonant nature of passive filters, there may be unwanted resonant interactions with the supply system.

(ii) Amplification of the currents on the source side at specific frequencies can appear owing to the parallel resonance between the source and the passive filter.

(iii) Excessive harmonic currents flow into the passive filter owing to the voltage distortions caused by the possible resonance with the source.

Active power filters, consisting of voltage or current source PWM inverters, have been studied and put into practical use because they have the ability to overcome the drawbacks inherent in passive filters [2]. The active filter

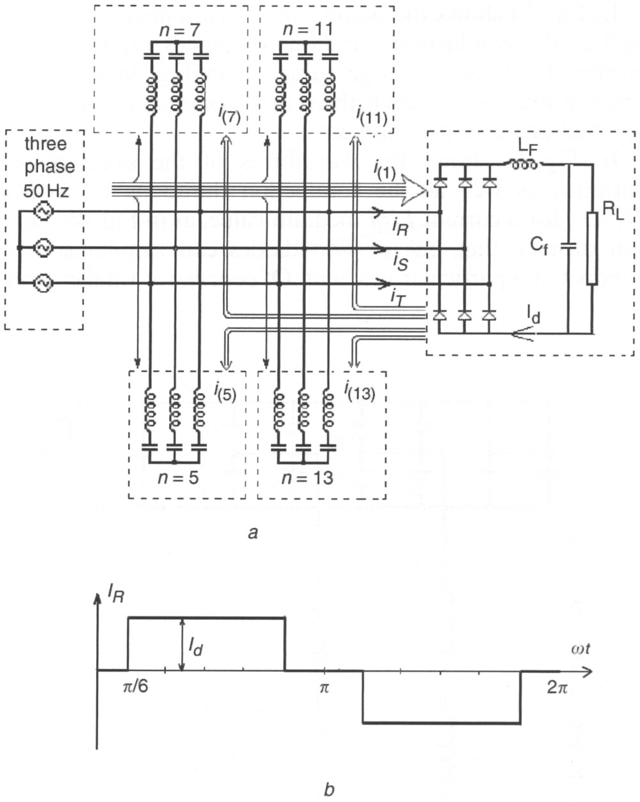

Fig. 1 *Three-phase, six-pulse, full-bridge diode rectifier with passive filters*
a Classical configuration
b AC current waveform i_R

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eliminates the harmonics that are present in the AC line by injecting a compensating current into the AC side. However, active filters have the following drawbacks:

(i) Difficulty to construct large-rated current sources with rapid current response.

(ii) High initial and running costs.

Obviously, the reduction of higher-order current harmonics generated by a three-phase AC–DC converter can also be obtained by using a PWM rectifier [3, 4]. Because of the rapid changes in voltages and currents within a switching converter, PWM rectifier equipment is a source of EMI with other equipments as well as with its own proper operation. The EMI is transmitted in two forms : radiated and conducted [1].The PWM rectifier, even though it has near sinusoidal input currents, has the following important drawbacks compared with the diode three-phase rectifier: larger commutation losses, higher costs and less reliability.In this paper we present several topologies of AC–DC converters having near sinusoidal input currents, implemented with diodes, inductors and capacitors, following a previous original circuit solution, [5–7].

2 New converter configuration

Figure 2 show an AC–DC converter generating reduced higher order current harmonics in the mains, called for short in what follows RNSIC (rectifier with near sinusoidal input current) [5]. The capacitors C_1–C_6 have the same value C and they are DC capacitors. The inductors L_R, L_S and L_T have the same value, denoted by L, and they are connected on the AC side. L and C fulfil the condition $0.05 \leq LC\omega^2 \leq 0.10$ in order for the phase currents i_R, i_S, i_T to be practically sinusoidal. The working principle of this rectifier can be explained by analysing the functioning stages, each lasting one-third of the mains period $T = 2\pi/\omega$ (where ω denotes the mains angular frequency).

In Fig. 3*a* shows the waveforms of currents i_R, i_S, i_T, as well as the conduction durations of diodes D_1–D_6, for a current I_d of relative large value, as in Fig. 3*b*. One can observe that two or even three diodes can simultaneously conduct at particular moment.

In Fig. 4*a* shows the waveforms of the same phase currents, as well as the conduction durations of diodes D_1–D_6, for a current I_d of medium value, as in Fig. 4*b*. One can observe that one or two diodes can simultaneously conduct at a particular moment. Of course, for smaller load currents i_L and, accordingly, smaller I_d values, no or only one diode can conduct at a particular moment.

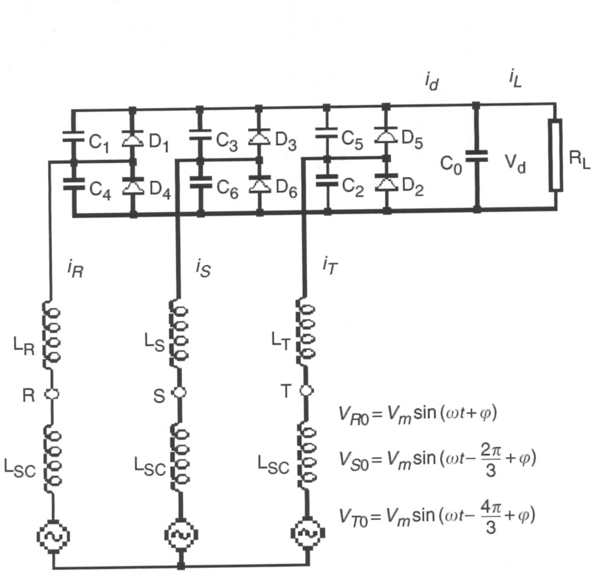

Fig. 2 *Configuration of RNSIC converter*

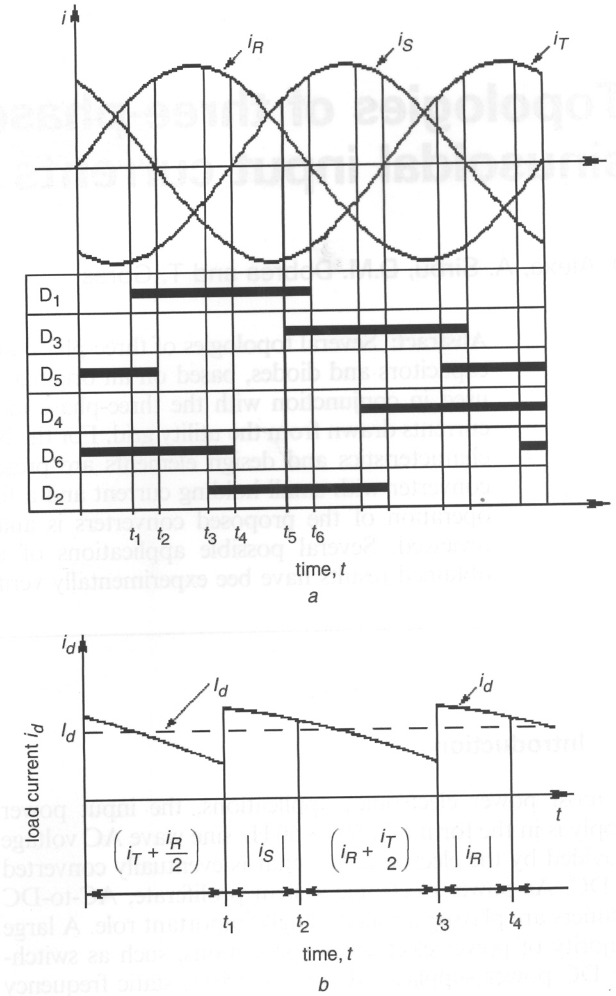

Fig. 3 *Waveforms for large values of I_d*
a AC current waveforms
b Load current i_d

For the three cases mentioned above, considering that the currents i_R, i_S, i_T are practically sinusoidal and have the amplitude $I_{(1)}$, a function of the load resistor R_L, the current I_d can be calculated from the following relation:

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

where t_1 is given in Figs. 3 and 4.

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 $\varphi = +90°$ 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 $\sqrt{3}V_m/(1 - 2LC\omega^2)$, the diodes D_1–D_6 do not conduct any more and the angle $\varphi = -90°$ 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_{max}/I_{min} has the value:

$$\left|\frac{I_{max}}{I_{min}}\right| = \frac{1 - 2LC\omega^2}{2LC\omega^2} \tag{2}$$

Figure 5*a* shows the variation of the angle φ, the phase displacement angle between the phase voltage and the

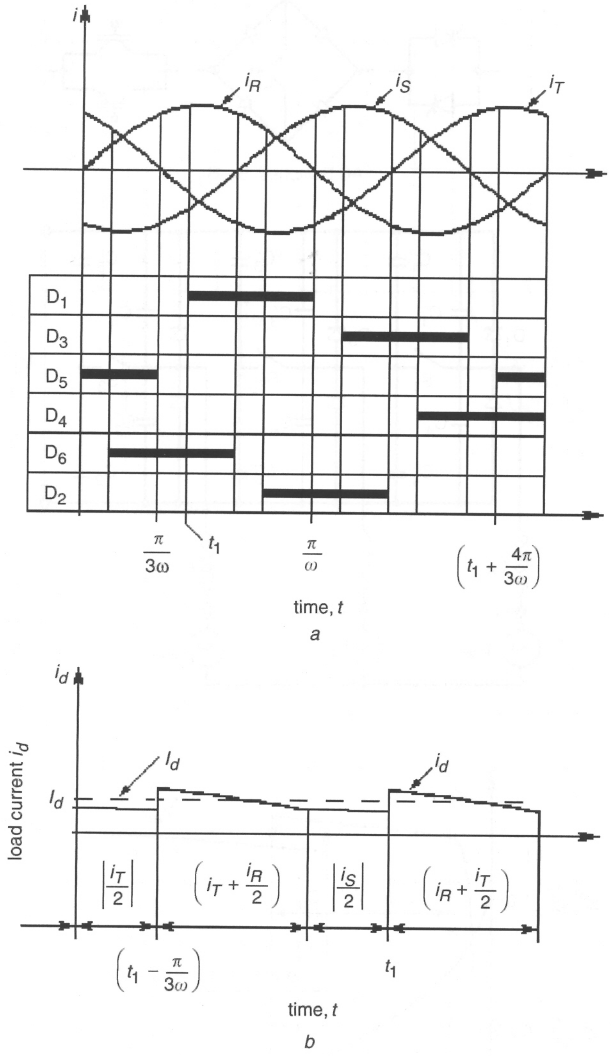

Fig. 4 *Waveforms for small values of I_d*
a AC current waveforms
b Load current i_d

fundamental of the phase current, as a function of the mean rectified voltage V_d rated to the reference value $V_{ref} = 3\sqrt{3}V_m/\pi$ specific for the classical three-phase rectifier [1]. The voltage V_d can be established at a certain value by the load current.

Figure 5*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).

To choose the capacitors C_1–C_6, beside the necessity to determine the maximum mean rectified voltage $\sqrt{3}V_m/(1 - 2LC\omega^2)$, one has to determine the *RMS* current I_{CRMS} that flows through such a capacitor:

$$I_{CRMS} = \sqrt{\frac{1}{2\pi}\int_0^{\omega t_1} 2\left(\frac{i_R}{2}\right)^2 dt} \tag{3}$$

if the waveform of the phase current is practically sinusoidal:

$$i_R \simeq I_{(1)} \sin \omega t \tag{4}$$

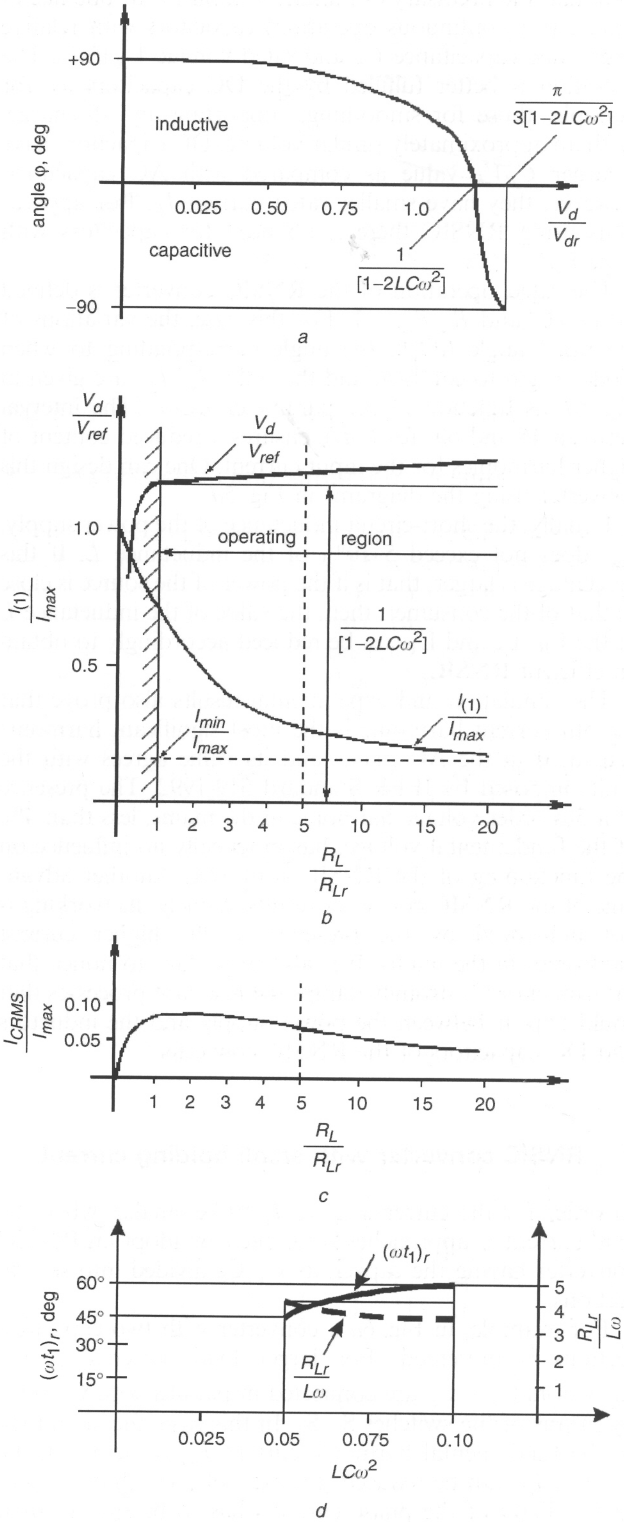

Fig. 5 *Characteristics of RNSIC converter*
a Angle φ as a function of ratio V_d/V_{ref}
b Ratios V_d/V_{ref} and $I_{(1)}/I_{max}$ as a function of R_L/R_{Lr}
c Ratio I_{CRMS}/I_{max} as a function of R_L/R_{Lr}
d Rated angle $(\omega t_1)_r$ and the ratio $R_{Lr}/(L\omega)$ as a function of $LC\omega^2$

it implies:

$$I_{CRMS} = \frac{I_{(1)}}{2}\sqrt{\frac{1}{2\pi}\left[\omega t_1 - \frac{\sin 2\omega t_1}{2}\right]} \tag{5}$$

Figure 5*c* shows the variation of the ratio I_{CRMS}/I_{max} as a function of R_L/R_{Lr}. Because the currents that flow through capacitors C_1–C_6 have small values as compared with I_{max}, to obtain the necessary ωt_1 angle, it implies that one has to choose (for continuous operation) capacitors with relative large rated capacitance C_R and rated voltage V_R [5–7]. The condition is better fulfilled by the DC capacitors as, for example, those for smoothing, supporting and discharge. With an approximately similar volume, DC capacitors have a larger $C_R V_R$ value as compared with AC capacitors; however, they have smaller rated current I_R. For applications using RNSIC there is no need for capacitors with large I_R.

The rated operation of the RNSIC converter is defined for $\varphi = 0°$ and $R_L/R_{Lr} = 1$. For this case, the variations of the rated angle $(\omega t_1)_r$, the angle corresponding to when diodes begin to conduct, and the ratio $R_{Lr}/L\omega$ are given in Fig. 5*d*, as function of the parameter $LC\omega^2$. The interval between 45 and 60° for $(\omega t_1)_r$ ensures a reduced content of higher harmonics for the input current. One can design this converter using the diagrams in Fig. 5*d*.

Usually, the short-circuit inductance of the power supply, L_{sc}, does not exceed 5–10% of the inductance L. If this percentage is larger, that is if the power of the source is close to that of the consumer, then, the value of the inductance L of the L_R, L_S and L_T can be reduced accordingly to obtain an efficient RNSIC.

The simulation and experimental results also prove that the 5th current harmonic is the most significant harmonic generated in the AC mains and its value agrees with the limits imposed by IEEE Standard 519-1992. The presence of a 5th order voltage harmonic in the mains, less than 4% of the fundamental voltage, has practically no influence on the functioning of the RNSIC converter. Another advantage of the RNSIC converter results: namely, its working is not influenced by the presence of the higher current harmonics in the mains. It is also important to notice that the capacitor C_0 strongly damps the resonant processes that could appear between the power supply and the inductors and DC capacitors of the RNSIC converter.

3 RNSIC converter with small holding current

In order for the currents i_R, i_S, i_T to be smaller when the load current i_L approaches zero, one can adopt an RNSIC converter having the capacitors C_1–C_6 divided into several sections.

For example, in Fig. 6*a* a converter with two capacitive sections is presented. For larger load currents i_L, the capacitors $C_1' - C_6'$ are connected in parallel with $C_1'' - C_6''$ by means of the switches S_1–S_3. In this case, the amplitude of the fundamental harmonic current $I_{(1)}$ is increased, the ratio $I_{sc}/I_{(1)}$ can be reduced (for example, less than 20) and so the THD of the phase currents has to be smaller than 5%, according to 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_L, the capacitors $C_1'' - C_6''$ are decoupled from the converter, but are still connected in parallel with the capacitor C_0. In this case, the amplitude of the fundamental harmonic current $I_{(1)}$ is reduced, the ratio $I_{sc}/I_{(1)}$ can range between 20 and 50, and so, the THD of the phase currents must be smaller than 8%.

Using this method, presented in Fig. 6*b*, the holding current I_{min} can be reduced approximately with the ratio $C_1'/(C_1' + C_1)$. One can observe that the RMS currents flowing through the switches S_1–S_3 have small values, less than 10% of I_{max} and so they need not be dimensioned for high currents.

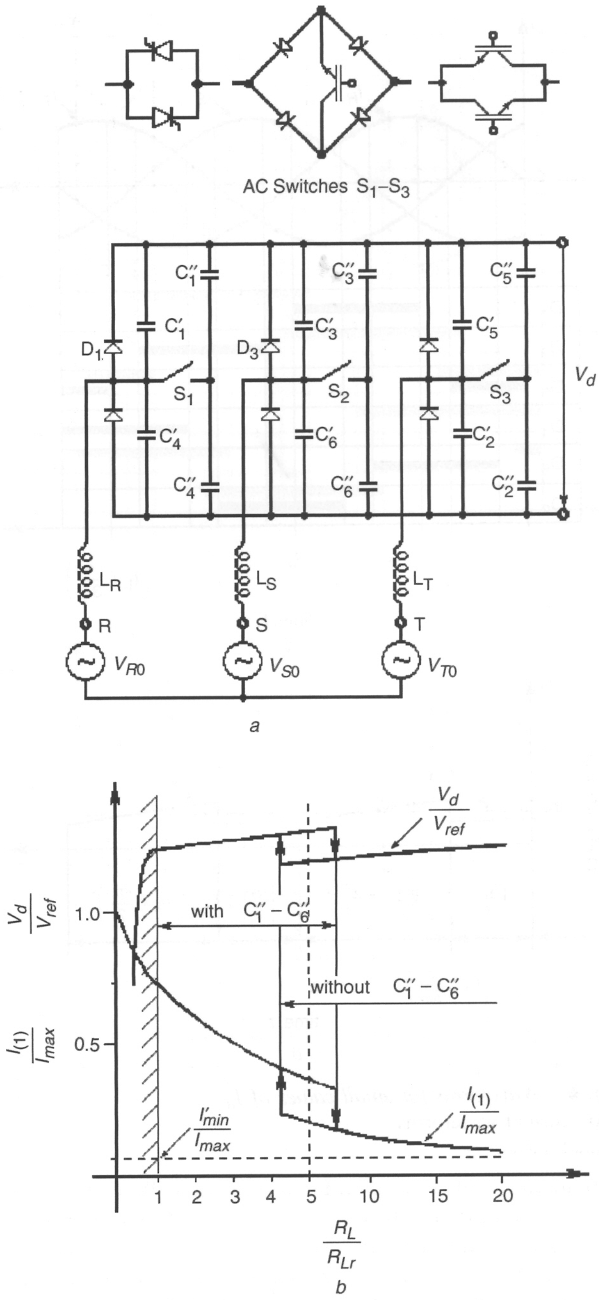

Fig. 6 *RNSIC converter with small holding current*
a Basic configuration
b Variations of the ratio V_d/V_{ref} and $I_{(1)\max}$ as a function of R_L/R_{Lr}

4 Frequency converter with RNSIC

A possible applications for RNSIC converters is their usage in static frequency converters with DC voltage link, designed for supplying with variable voltage and frequency the three-phase induction motor drive, as in Fig. 7.

For the time intervals when the induction motor drive is in the motoring regime the proposed converter in the input becomes an RNSIC converter. For this case the transistors T_1–T_6 are off and the switches S_1–S_3 are closed. The output switch-mode converter operates as a PWM inverter. The energy is transmitted from the power supply to the motor and the voltage on the capacitor C_0 is less than $\sqrt{3}V_m/(1 - 2LC\omega^2)$.

During the time interval while the induction machine operates in breaking 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

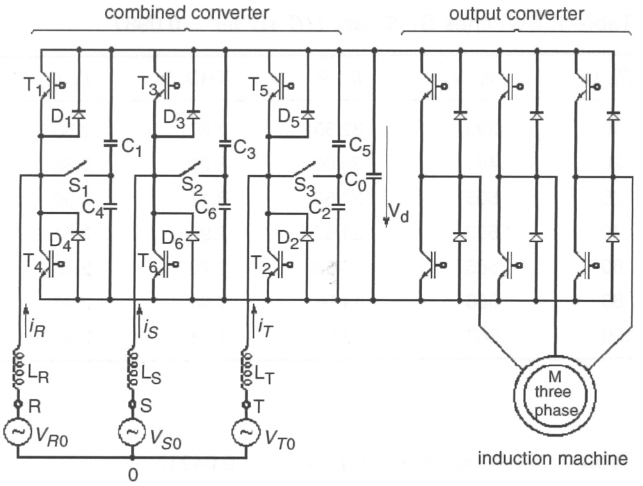

Fig. 7 *Four-quadrant converter with RNSIC*

greater than $\sqrt{3}V_m/(1-2LC\omega^2)$. Further on, the energy is transmitted into AC mains by means of a PWM inverter made up of transistors T_1–T_6, inductors L_R, L_S and L_T and diodes D_1–D_6. Obviously, in this case, the switches S_1–S_3 must be off and the capacitors C_1–C_6 remain connected in parallel with C_0 .

One can observe from the above that diodes D_1–D_6 and inductors L_R, L_S and L_T are always in use whether the combined converter operates as a rectifier or a inverter. One must mention also the fact that the total duration of operation as a generator for the asynchronous machine is much smaller as compared with the total motor functioning duration. Switches S_1–S_3 have to be dimensioned for small currents (maximum 20% of I_{max}).

As compared with a classical PWM rectifier connected at the input of a frequency converter, the combined converter presented in Fig. 7 is characterised by smaller commutation losses and greater reliability. The higher efficiency is obtained because the combined converter in the input operates mostly as an RNSIC. The greater reliability is explained by the fact that, supposing that the transistors or their command circuits are damaged, the functioning of the combined converter in the input as PWM inverter is abandoned. In this case, the energy received by C_0 from the induction motor drive operating as a generator can be dissipated on a discharge resistor in series with a switch.

Because the output of a RNSIC converter voltage V_d 15–25% larger than the reference voltage V_{ref} obtained from a three-phase classical diode rectifier, it implies that at the output of the PWM inverter one can get the rated voltages for the three phases supplying the induction motor drive. In this way, there is no need to apply on overmodulation technique (as, for example, methods of PWM pattern generation with third-harmonic injection or with partially constant modulating waves) [8, 9].

The rectifier rated voltage $V_{dr}=V_{ref}/(1-2LC\omega^2)$ surpasses the reference value V_{ref} ($V_{dr}=kV_{ref}$, where $k=1/(1-2LC\omega^2)$ is an overvoltage coefficient varying between 1.15 and 1.25). This is the reason for which one can get stator phase voltages, V_S, applied to the induction machine, practically surpassing with the same coefficient k the rated voltage V_{Sr}, according to Fig. 8. From this figure, it can be seen that the motor supplied from a frequency converter with RNSIC connected at its input is used more efficiently even at frequencies larger than the rated one, f_r. In this case, the region of functioning at a rated nominal torque T_{emr} is larger, and the power obtained from the motor at frequencies higher than f_r is larger.

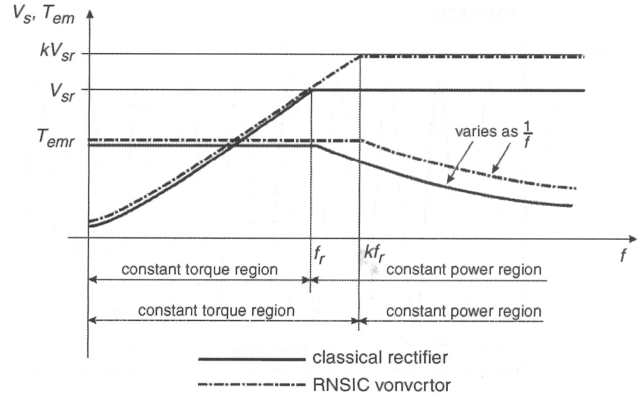

Fig. 8 *Induction machine characteristics and capabilities*

5 Experimental and simulation results

Laboratory experiments and simulation results have proved the effectiveness of the proposed three-phase low-harmonic rectifier. The laboratory prototype shown in Fig. 6*a* consists of a three-phase voltage source (with $V_m=250\,\text{V}$ and $f=50\,\text{Hz}$) and an RNSIC converter with small holding current. This converter is composed of six diodes, three inductors and 12 DC capacitors with capacitance $20\,\mu\text{F}$. For the inductors L_R, L_S and L_T we have adopted the value 25 mH. The filtering capacitor C_0 is $1000\,\mu\text{F}$ and the load resistor R_L can be varied between 15 and $1000\,\Omega$. The total power dissipation P for a capacitor with $C_R=20\,\mu\text{F}$ is composed of the dielectric losses (P_D) and the resistive losses (P_R). This power, P, does not exceed 2–3 W for $V_m/\sqrt{2}=220\,\text{V}$.

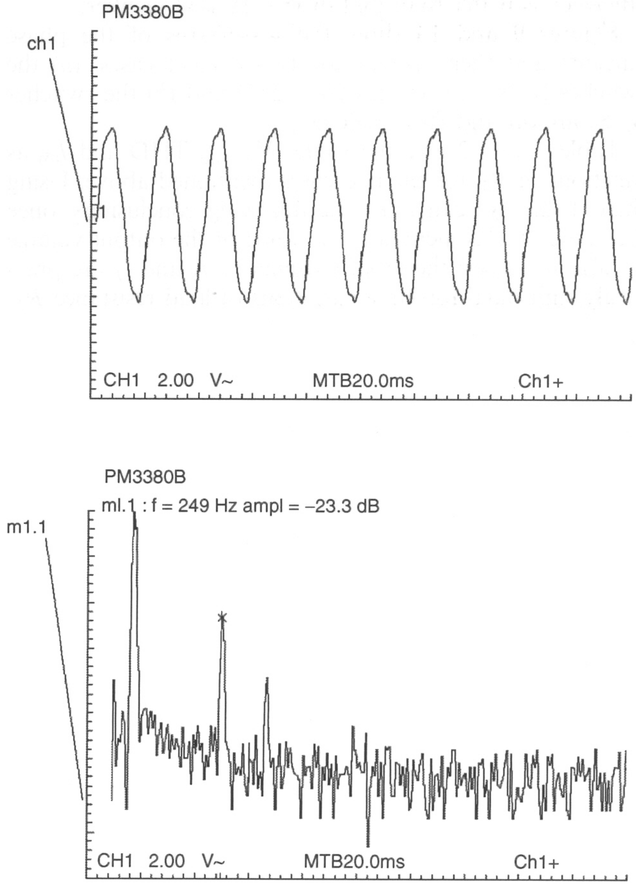

Fig. 9 *Waveforms of phase current and its spectrum for $R_L=25\,\Omega$ (switches S_1–S_3 on), probe 10:1*

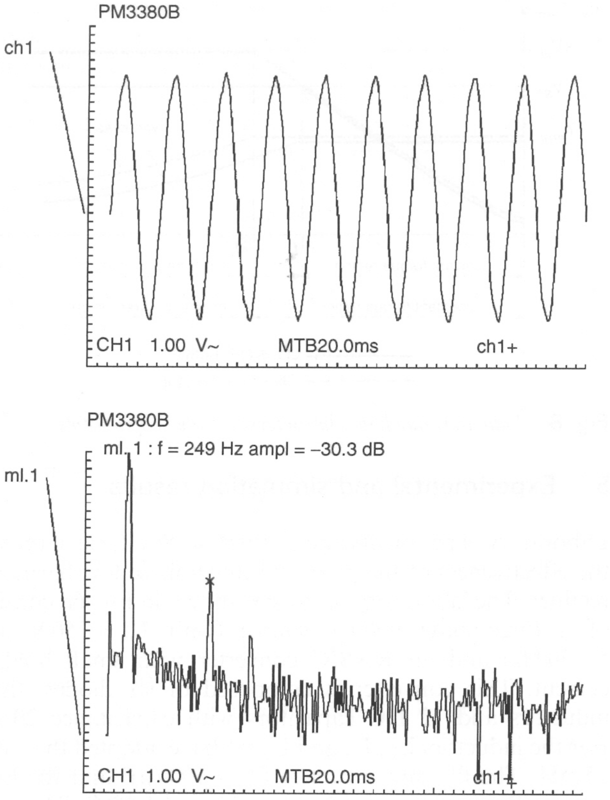

Fig. 10 *Waveforms of phase current and its spectrum for* $R_L = 1000\,\Omega$ *(switches* S_1–S_3 *off), probe 1:1*

For the same output power, $P_d = V_d I_d$, the current stress for the RNSIC diodes is 15–20% less as compared with that of the classical rectifier. This implies that the RNSIC efficiency is better than that of the classical rectifier.

Figures 9 and 10 show the waveforms of the phase currents and their spectra for two distinct cases: (*a*) the switches S_1–S_3 are on and $R_L = 25\,\Omega$ and (*b*) the switches S_1–S_3 are off and $R_L = 1000\,\Omega$.

Tables 1 and 2 give the values V_d, I_m, THD and $I_{(5)}$ as functions of R_L for the two cases mentioned above. Using these Tables we can draw the following conclusions: once the value of *C* is increased, the value of the output voltage V_d also increases; the input currents i_R, i_S and i_T are practically sinusoidal for large variations of load resistance R_L.

Table 1: Switches S_1–S_3 on, $L(C'+C'')\omega^2 = 0.0987$

R_L, Ω	V_d, V	I_m, A	THD, %	$I_{(5)}/I_{(1)}$, %
15	393	32.30	2.67	2.50
20	461	30.00	3.01	2.88
25	505	27.56	3.11	2.90
40	553	21.57	2.95	2.81
60	565	17.03	2.70	2.48
80	566	14.56	2.64	2.36
100	567	13.05	2.73	2.47

Table 2: Switches S_1–S_3 off, $LC'\omega^2 = 0.04935$

R_L, Ω	V_d, V	I_m, A	THD, %	$I_{(5)}/I_{(1)}$, %
80	482	9.97	6.61	6.35
100	484	8.18	6.30	5.95
150	485	6.01	6.45	5.45
200	488	5.60	6.28	5.74
400	498	4.40	7.49	7.24
1000	513	3.78	6.90	6.72

6 References

1 Mohan, N., Undeland, T., and Robbins, W.: 'Power electronics–converters, applications and design', (John Wiley & Sons Inc, 1995)
2 Akagi, H.: 'Trends in active power conditioners', *IEEE Trans. Power Electron.*, 1994, **9**, (3), pp. 263–268
3 Viriya, P., Kubota, H., and Matsuse, K.: 'New PWM-controlled GTO converter', *IEEE Trans. Power Electron.*, 1987, **2**, (4), pp. 373–381
4 Chung, D.W., and Sul, S.K.: 'Minimum-loss strategy for three-phase PWM rectifier', *IEEE Trans. Ind. Electron.*, 1999, **46**, (3), pp. 517–526
5 Alexa, D.: 'Three-phase rectifier with almost sinusoidal input current', *Electron. Lett.*, 2001, **37**, (19), pp. 1198–1150
6 Alexa, D.: 'Combined filtering system consisting of passive filter with capacitors in parallel with diodes and low-power inverter', *IEE Proc. Electr. Power Appl.*, 1999, **146**, (1), pp. 88–94
7 Alexa, D., and Sirbu, A.: 'Optimized combined harmonic filtering system', *IEEE Trans. Ind. Electron.*, 2001, **48**, (6), pp. 1210–1218
8 Hava, M.A., Kerkman, J.R., and Lipo, T.: 'Carrier-based PWM-VSI overmodulation strategies : analysis, comparison and design', *IEEE Trans. Power Electron.*, 1998, **13**, (4), pp. 674–689
9 Alexa, D., and Lazar, A.: 'Optimisation of PWM Techniques with Partially Constant Modulating Waves', *Electr. Eng.*, 2000, **82**, (5), pp. 263–272