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# A DC to 3-Phase Series-Resonant Converter with Low Harmonic Distortion

H. HUISMAN AND S. W. H. DE HAAN

**Abstract**—A type of dc to 3-phase series-resonant converter (s.r.-converter) for potentially submegawatt industrial applications is presented. The converter provides variable-frequency sine-wave currents, with low harmonic distortion at the output terminals, and with the frequency ranging from  $-200$  through dc to  $+200$  Hz. The converter can transfer power in both forward and reverse power-flow directions to almost any type of load circuit. The methods of control are formulated such that they can be implemented easily with high-speed logical circuits. Test results for a 1-kW demonstration converter are supplied.

## I. INTRODUCTION

**D**URING the last decade the interest in electrical wind-energy conversion has increased significantly. This attention concerns both the generation of autonomous multiphase ac grids, as well as the delivery of power to existing utility grids. To carry out the electrical conversion process, one often uses a cycloconverter or a cascaded ac-dc and dc-ac converter. The dc to 3-phase converter presented in this paper is considered as an intermediate step in the development of a novel type of 3-phase-to-3-phase cycloconverter. Apart from wind-energy conversion this converter can be applied to other areas of technology including power generation and variable frequency-machine driving.

Fundamentals of dc to 3-phase ac power conversion have been known for many decades. Because of the simplicity of control and the clarity of insight into the method of operation, some of these systems, such as the load-commutated bridge inverter, the square-wave inverter, and the high-frequency pulsewidth-modulated inverter, are well developed [1] and widely spread. The advantages of these systems are, however, countered by one or more of the following disadvantages:

- 1) the stress on electronic switches is high due to forced commutation techniques,
- 2) heavy and costly filters are required to remove the low-frequency harmonic content of the electrical waveforms, and
- 3) the power factor is poor and cannot be controlled freely.

The s.r. dc to 3-phase ac converter presented, which is based on a concept presented by Schwarz [2], does not possess these disadvantages. Moreover, the system has the common features of series-resonant converters (s.r.-converters) such as, 1) the system is inherently short-circuit proof, 2) the size of the system is small due to the high internal frequency, and 3) the system is capable of controlling and turning off power by a fraction of a millisecond. The technology of the single-quadrant

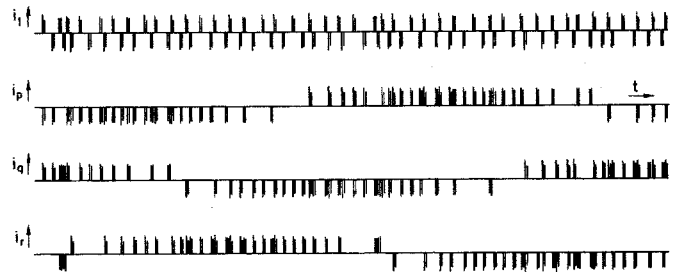


Fig. 1. Example of the distribution of the current pulses in the resonant current  $i_r$  over the line currents  $i_p$ ,  $i_q$ , and  $i_r$ .

dc-dc s.r.-converter has been well established by now and many authors have discussed the subject [3]–[6].

Since the beginning of the eighties attempts have been made to break through to the development of ac and multiphase s.r.-converters. Some successful experiments of ac-dc s.r.-converters were reported by Schwarz [7], [8]. Recently, a single-phase dc-ac converter has been presented [9]. The 3 $\phi$ -dc (3-phase to dc) s.r.-converters described up to now [7], [8] do, however, extract currents from the power grid which are similar to the currents which are extracted by a 3-phase diode rectifier. These square-wave-like currents cause harmonics in the supply lines, while the power factor is theoretically limited to 0.955 [8].

From a fundamental point of view, the s.r.-converter is capable of dealing with multiphase ac power in a more sophisticated way [2]. A s.r.-converter can be considered as a network which generates an array of current pulses at a high frequency (5–50 kHz). By properly distributing the pulses from this high-frequency carrier over the 3 lines of the grid, it is possible to generate current waveforms, which do contain one single frequency in the low-frequency range ( $<3$  kHz). The principle of the process is illustrated in Fig. 1, where resonant current pulses are represented by impulses. Fig. 1 is obtained from a simulation program, which operates similarly to the one described in [10]. The high-frequency content of the output current is removed by a relatively small filter, with the result being a low-frequency phase current with low harmonic distortion. The pulse distribution process can be applied both to 3 $\phi$ -dc converters as well as to dc-3 $\phi$  converters.

In this paper a dc-3 $\phi$  converter is described which is based on the above principle. The converter is capable of generating a 3-phase output waveform with frequencies ranging from ca  $+200$  via 0 to  $-200$  Hz (phase reversal). Moreover, the converter can be connected to leading as well as lagging loads ( $+180^\circ$  to  $-180^\circ$ ), both linear and nonlinear, in both the forward and

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The authors are with the Delft University of Technology, Mekelweg 4, 2624, CD Delft, The Netherlands.

reverse power-flow directions. The critical points that can be identified in the system design of the dc-3 $\phi$  converter are: 1) the pulse-distribution algorithm (line selection), 2) the algorithm for the selection of particular switches in the selected lines (switch selection, and 3) the method of control to guarantee thyristor turnoff and continued oscillation ( $V_c$ -control). The method of control is in accordance with a control mode, which is described in a companion paper [11]. The theory is verified on a small-scale (1 kW) demonstration converter.

## II. PROPERTIES OF THE GENERALIZED DC-DC S.R.-CONVERTER

The operation modes of the dc-3 $\phi$  s.r.-converter is based on the principles of the dc-dc s.r.-converter, which is treated in literature [3]–[6], [11], [12]. To enable greater insight, some properties of the dc-dc s.r.-converter will be reconsidered in a nonorthodox manner. Fig. 2 shows the basic, generalized dc-dc s.r.-converter. The input switch matrix consisting of switches  $Q_A$  and  $Q_B$  connects the resonant circuit to  $+E_s$  or  $-E_s$ , where  $E_s$  is the dc source voltage. The output matrix consisting of switches  $Q_C$  and  $Q_D$  connects the resonant circuit to  $+V_o$  or  $-V_o$ . By appropriate control of the switch matrices a bipolar voltage, compounded of square waves, is generated across the resonant circuit, thus inducing a resonant current of the well-known shape (Fig. 3(a)). In this paper the resonant current is considered as a series of current pulses. The time of initiation of the  $K$ th current pulse is denoted by  $\beta_k$ , where  $\beta$  is the normalized time defined by  $\beta = t/\sqrt{L_1 C_1}$ .

The current gap which is not present in most dc-dc converters is a prerequisite in ac-converters for the purpose of commutation and control [11]. Depending on the application, the switches are implemented as single diodes or as combinations of diodes and/or controllable switches such as thyristors and MOSFET's. In multiphase converters all symbolical switches should be implemented as thyristors. Although the multiphase converter does not contain diodes, we will stick to the term "diode current" for each first current segment following a current zero crossing; the second current segment will be designated as "thyristor current". The  $K$ th event between the time  $\beta_k$  and  $\beta_{k+1}$  will also be referred to as the  $K$ th converter cycle.

It is well known that a number of switching modes can be distinguished for the generalized s.r. dc-dc converter [4], [11]. It will be shown that all switching modes have the following properties in common:

- 1) the diode current is (at least) opposed by the highest terminal voltage (this follows from the commutation condition),
- 2) the thyristor current at the high-voltage side extracts power from that source (this follows from the power balance), and
- 3) both diode and thyristor currents on the low-voltage side transfer power in the desired direction.

Conversely, one may state that if the above rules are obeyed, one can transfer power in any desired direction in any combination of input and output voltages.

The statements are based on the following considerations.

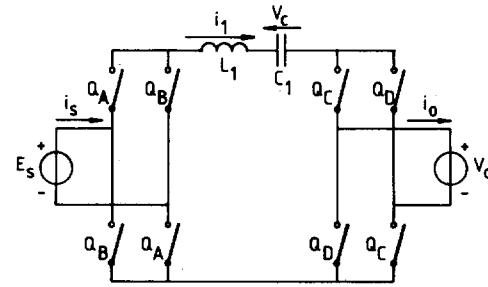


Fig. 2. Basic schematic and the definition of the voltages and currents in a generalized s.r. dc-dc converter.

Suppose that in the symbolic converter from Fig. 2, the source voltage  $E_s$  and the output voltage  $V_o$  are both allowed to be positive or negative. In either an input or an output line the current will be prescribed by some reference. The desired power direction is defined by the sign of this reference current and the sign of the actual voltage at the associated port. As stated before, a resonant current phase consists of a diode current and a thyristor current.  $Q_D$  and  $Q_T$  are defined as the charge transferred by the diode and the thyristor current, respectively. Both  $Q_D$  and  $Q_T$  are defined such that they are always positive. The input and output matrices facilitate  $Q_D$  and  $Q_T$  being routed in any desirable direction on input and output terminals.

If one bears in mind that the energy transferred by a current in a certain interval is equal to the product of voltage and transferred charge, it follows from considering the energy balance applied to the input and output port of the converter (Fig. 2)

$$E_s(Q_D S_{11} + Q_T S_{12}) = V_o(Q_D S_{21} + Q_T S_{22}). \quad (1)$$

Here  $S_{11}$  to  $S_{22}$  indicate in which direction the charge flows on input and output, so  $S_{ij}$  is either +1 or -1. Note that (1) is valid under nonsteady-state conditions, provided that  $v_{cD}(k) = v_{cD}(k+1)$  and that both  $E_s$  and  $V_o$  are constant during the cycle considered.

From (1) it follows that

$$-\frac{Q_D}{Q_T} = \frac{S_{12}E_s - S_{22}V_o}{S_{11}E_s - S_{21}V_o}. \quad (2)$$

Because both  $Q_D$  and  $Q_T$  are positive, the sign of the left-hand term is negative. To satisfy (2), the coefficients  $S_{11}$  and  $S_{12}$  of the highest voltage, either  $E_s$  or  $V_o$ , should be opposite in sign. On the low-voltage side both  $S_{21}$  and  $S_{22}$  are unconstrained, so that they are chosen such that the power is transferred in the desired direction. For a proper commutation from the diode to the thyristor current the diode current is required to be opposed by the highest voltage. By way of example Fig. 3(b) shows the resulting current waveforms of  $i_s$  and  $i_o$ , if  $E_s > 0$ ,  $V_o > 0$ ,  $|E_s| < |V_o|$  and  $i_{o\text{ref}} < 0$ .

The elaboration of the switch selection mechanism in the dc-3 $\phi$  converter presented is primarily based on the previous discussion. The basic diagram of the power network of the converter is depicted in Fig. 4. The switches symbolize appropriate electronic switches such as antiparallel thyristors.



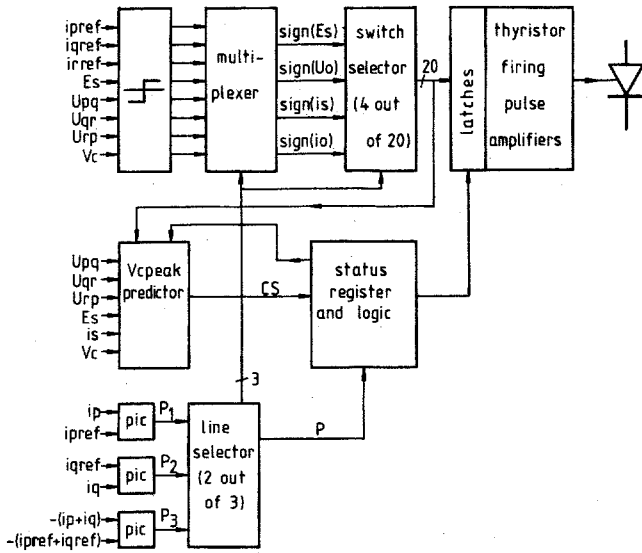


Fig. 5. Block diagram of the controller for the dc-3φ s.r.-converter. The PIC and the  $V_{cp}$ -predictor do process analog signals. All other circuits process booleans.

desired power transfer direction, defined by  $I_{oref} \cdot U_o$ , is satisfied.

So, if

$$|V_o| < |E_s|$$

$$\text{then sign}(i_o) = \text{sign}(i_{oref})$$

$$\text{and sign}(i_s) = -\text{sign}(E_s)$$

and if

$$|V_o| > |E_s|$$

$$\text{then sign}(i_o) = \text{sign}(V_o)$$

$$\text{and sign}(i_s) = \frac{\text{sign}(i_{oref} V_o)}{\text{sign}(E_s)} \quad (4)$$

If the directions of  $i_s$  and  $i_o$  are known, and if the direction of  $i_1$  is known, a unique set of four appropriate thyristors can be designated. Note that  $\text{sign}(i_1)$  in the  $K$ th cycle follows from  $\text{sign}(v_c(\beta_k))$ .

2) *Thyristor Cycle*: The current direction on the low-voltage side occurs during the thyristor cycle such that the desired power-transfer direction is satisfied. On the high-voltage side the direction of the thyristor current is opposite to that of the diode current on that side.

So, if

$$|V_o| < |E_s|$$

$$\text{then sign}(i_o) = \text{sign}(i_{oref})$$

$$\text{and sign}(i_s) = \text{sign}(E_s)$$

and, if

$$|V_o| > |E_s|$$

$$\text{then sign}(i_o) = -\text{sign}(V_o)$$

$$\text{and sign}(i_s) = \frac{\text{sign}(i_{oref} V_o)}{\text{sign}(E_s)} \quad (6)$$

If the signs of  $E_s$ ,  $V_o$ ,  $I_{oref}$ , and  $v_c$  are determined, the appropriate switches can be designated easily from (3) to (6).

#### D. Line Selection

In the following section it is assumed that a s.r.-converter should feed an  $M$ -phase network with ac power. For the purpose of driving machinery, it is suitable if the converter behaves like a multiphase current source, although for multiphase power generation a voltage-source character is more appropriate. The line-selection mechanism will be explained with respect to a current-source character. The multiphase converter should be controlled such that the current  $i_m$  in line  $m$  ( $m = 1, 2, \dots, M$ ) is a replica of a reference current  $i_{mref}$  ( $m = 1, \dots, M$ ). The reference current in lines  $1, \dots, M-1$  can be chosen freely; the reference current in line  $M$  follows from

$$\sum_{m=1}^M i_{m,ref} = 0. \quad (7)$$

The current waveform synthesis is based on the concept of pulse-integral control [11], [12]. In a dc-dc converter the resonant current  $i_1$  is measured, subsequently rectified and compared to a reference  $i_{ref}$ . The error signal ( $|i_1| - i_{ref}$ ) is fed to an integrator. Whenever the integrator output  $e_c$  exceeds some threshold  $e_T$ , the power switches are actuated, thus generating another current pulse. When two successive threshold crossings are denoted by  $t_a$  and  $t_b$ , the following holds:

$$\int_{t_a}^{t_b} (i_o - i_{ref}) dt = 0. \quad (8)$$

For dc and single-phase converters,  $t_a$  and  $t_b$  coincide with the beginning of two successive converter cycles.

For normal unipolar dc-dc s.r.-converters the implementation of the pulse-integral controller (PIC) for the output current  $i_o$  is relatively simple because  $i_o = |i_1|$ . In multiphase converters only two lines are served during a converter cycle. For that reason (8) is implemented such that the interval ( $t_a$ ,  $t_b$ ) effectively spans several cycles. This implies that during each cycle a limited error occurs for each line. Moreover, the implementation should be modified such that the PIC can handle both positive and negative values of  $i_1$  and  $i_{ref}$ . Fig. 6(a) gives the implementation of one of the  $M-1$  PIC's and Fig. 6(b) shows the associated waveforms. Whenever the integrator output  $e_c$  gets into one of the arced areas, a pulse-request signal  $P_m$  is generated.

The gap between the comparator levels is intended to prevent the immediate generation of a pulse request whenever the reference signal changes polarity. If  $\Delta = 0$ , this undesirable side effect will occur. For optimal functioning the gap should





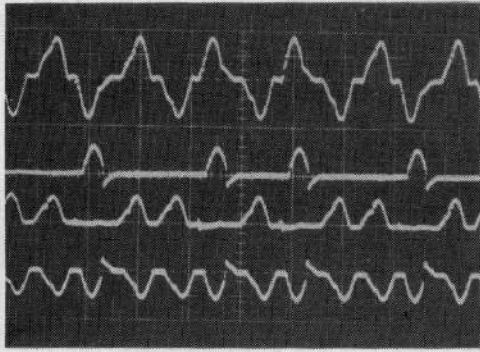


Fig. 7. Pulse-distribution process at a low output frequency. Upper trace: resonant current  $i_a$ , 40 A/div. Lower traces: unfiltered line currents, 50 A/div, 100  $\mu$ s/div.

The converter was primarily intended to demonstrate the most important features of the dc-3 $\phi$  s.r.-converter such as harmonic distortion, power factor regulation and load independence. Optimizing the performance with respect to efficiency and size was not a goal, although a maximum efficiency of 85 percent was established. The relatively high losses are due to the low source voltage in combination with the oversized power network.

Fig. 7 shows the resonant current and the unfiltered output currents at a relatively low line frequency. Note the difference in shape of the current pulse in the lines  $p$  and  $q$ , which is due to the fact that  $V_{pr} < E_s$  while  $V_{qr} > E_s$ .

Fig. 8 gives an impression of the pulse-distribution mechanism. Simultaneously, it is shown that the capacitor peak voltage is maintained within 10 percent at a constant level.

The filtered and unfiltered currents in line  $p$  are shown in Fig. 9. The frequency of the reference current is approximately 60 Hz. From the frequency spectrum of Fig. 10(a) it can be concluded that the harmonic distortion of the filtered line current is extremely low. The 3rd, 5th, and 7th harmonics are at least 40 dB below the 100-Hz fundamental component. For comparative purposes, the harmonics of a line-commutated converter are indicated in the same graph. The difference is self-explanatory.

As stated in relation to (7), two reference currents can be chosen independently. For the purpose of examining the cross distortion between lines, a frequency spectrum (Fig. 10(b)) was recorded for a condition in which line  $p$  generated a 100-Hz sinoid, and line  $q$  a 130-Hz sinoidal current. The cross distortion is approximately -25 dB. Note that the 3rd, 5th, and 7th harmonics have disappeared from the graph. The "grass" at -60 dB between the harmonics is due to the more or less stochastic character of the unfiltered line current, which is rooted in the fact that the inverter cycle is not synchronized with the ac-cycle. From the last graph it may be concluded that the converter presented can be considered as a multiphase current amplifier. Any arbitrary current  $i_{ref}$  can be reproduced in the output line.

To investigate the ability of the converter to feed currents in nonlinear and active loads, two test circuits were set up. In the first circuit the converter was connected to an asynchronous machine. The converter was able to accelerate and decelerate the machine in both rotation directions. When the axis of the machine was connected to a mechanical drive, it was possible

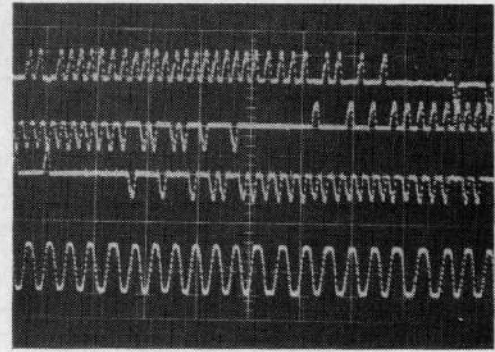


Fig. 8. Sinusoidal (200 Hz) modulated output currents. Upper traces: unfiltered line currents 50 A/div. Lower trace: resonant capacitor voltage  $v_c$ , 700 V/div.

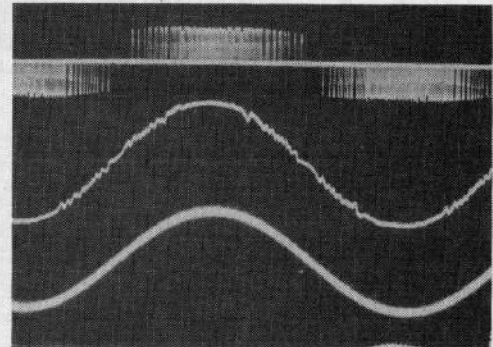


Fig. 9 Sinusoidal (60 Hz) modulated currents in line  $p$ . Upper trace: unfiltered current in line  $p$ , 50 A/div. Middle trace: filtered current in line  $p$ , 10 A/div. Lower trace: reference for line  $p$ , 2 ms/div.

to use the machine as a generator, thus transferring power to the dc "source"  $E_s$ . The reactive power for the machine was in both cases supplied by the converter.

In the second-setup the converter was connected via a transformer to the 3-phase utility grid. In order to test the ability of the converter to supply current independent of the actual output voltage, a 33 1/3-Hz 3 $\phi$  reference was phase locked to the 50 Hz grid, forcing the converter to supply a 33 1/3-Hz current. The filtered output currents, shown in Fig. 11, arise from a superposition of the 33 1/3-Hz converter current and a 50-Hz current, which is caused by the connection of the converter filter capacitors to the grid.

#### IV. CONCLUSIONS

A dc-3 $\phi$  s.r.-converter is presented which has a combination of favorable properties compared to other types of multiphase converters.

These properties include:

- low size and weight of the converter;
- low harmonic distortion of the generated currents;
- ability to generate low-frequency ac currents (up to several hundreds of Hz) of any arbitrary shape;
- ability to feed currents to almost any type of load including existing utility grids, machinery and passive loads;
- potential ability to process hundreds of kW of power with an efficiency of more than 90 percent;
- high speed of reaction;
- inherently short-circuit proof; and
- both forward and reverse power flows.

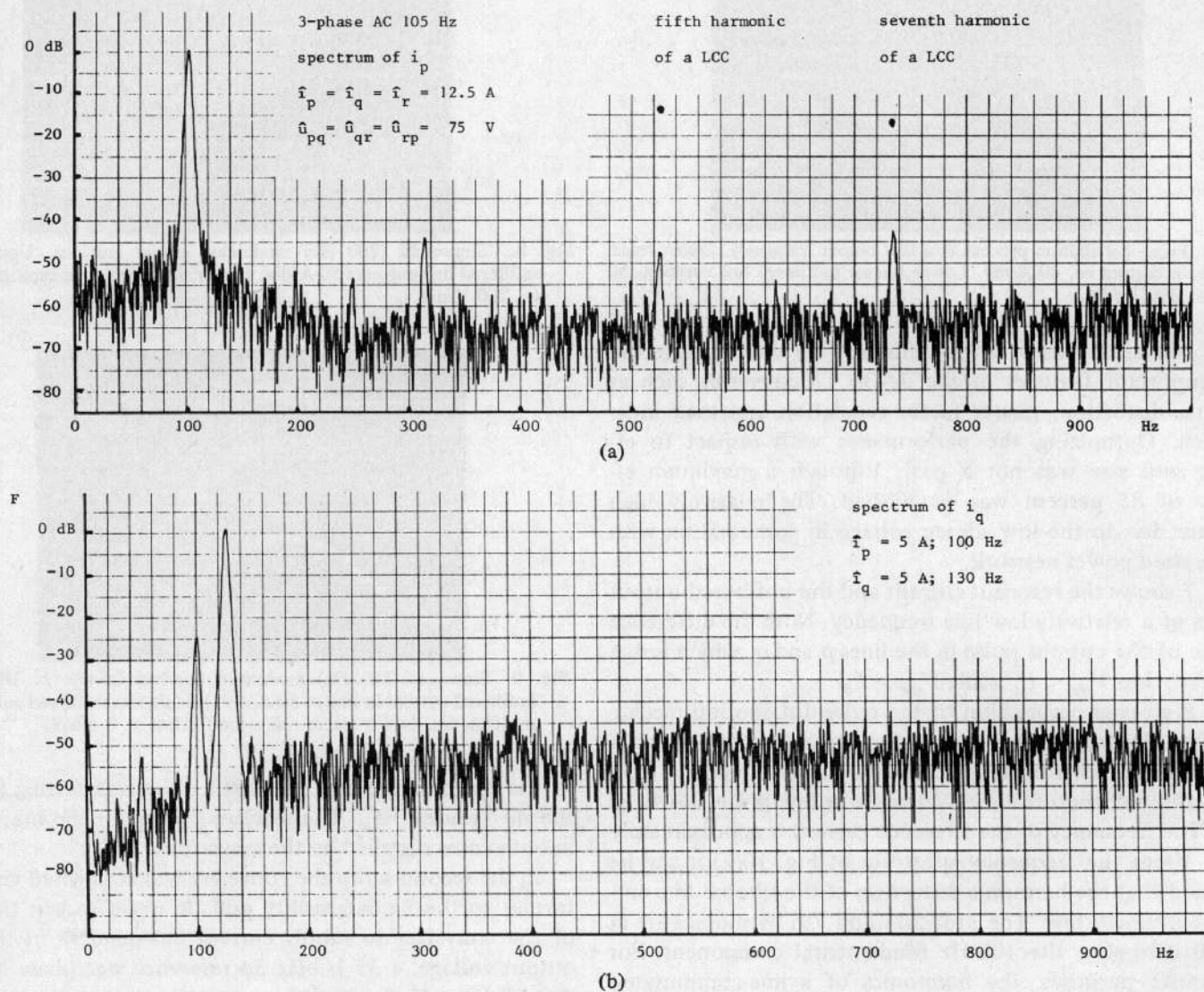


Fig. 10 Frequency spectra of the line currents. (a) Spectrum of  $i_p$ ; the reference for line  $p$  is a 100-Hz sinoid. (b) Spectrum of  $i_p$ ; the references for the lines  $p$  and  $q$  are 100- and 130-Hz sinoids, respectively.

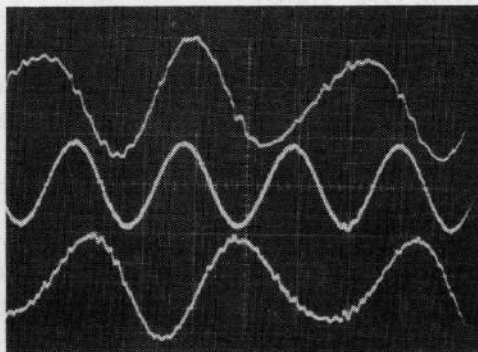


Fig. 11 Filtered load currents in lines  $p$  and  $q$  when the converter is connected to a 50-Hz 3-phase grid and while the current reference is set to  $33 \frac{1}{3}$  Hz. Upper trace:  $i_p$ , 10 A/div. Middle trace:  $V_{pq}$ , 100 V/div. Lower trace:  $i_q$ , 10 A/div; 10 ms/div.



The properties of the converter are rooted in combining a well-known s.r. power network topology with a novel method of preserving the energy content of the converter ( $v_{cp}$ -control), and of distributing current pulses over the output lines.

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