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High performance direct instantaneous power control of PWM rectifiers

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ABSTRACT

This paper presents a new direct instantaneous power control (DPC) strategy for active rectifiers. In this novel scheme the PWM modulator has been utilized instead of the hysteresis comparators and switching table. The required converter voltage in each sampling period is directly calculated based on the reference and measured values of powers, system parameters, and the measured voltage of the AC source through simple equations which are wisely compensated for variations of the grid voltage during a sampling period. Then, the PWM generator generates the switching pulses for the voltage source converter. It is shown that the proposed DPC–PWM exhibits several features, such as a simple algorithm, constant switching frequency, robust to sampling frequency changes, robust to inductance values mismatch, and particularly it provides low sampling frequency. Extensive simulation and experimental results have proven the excellent performance and verify the validity and effectiveness of the proposed instantaneous power control scheme.

1. Introduction

The increasing power demand in most of the consumer centers has forced the planning engineers to look for new solutions to improve the quality of electrical utility systems. Conventional applications of VSC such as AC and DC drives, AC and DC power supplies, active filters, electrolyze systems, and electric furnaces require controlling the active power or accurate control of DC link voltage at set value. DC link voltage control (usually through PI regulator) leads to active power control. As power electronic systems are extensively used, not only in industrial applications, but also in consumer products, several problems with regard to their diode rectifiers have arisen in recent years. One of the problems is a low input power factor, and another problem is caused by harmonics in input currents. Research interest in three-phase PWM rectifiers has grown rapidly over the past few years due to some of their important advantages, such as power regeneration capabilities, control of DC link voltage, low harmonic distortion of input currents, bidirectional control of active and reactive powers, small DC link capacitor, and high power factor (usually, near unity) [1]. Various control strategies have been proposed in recent works on this type of rectifiers. A well-known method of indirect active and reactive power control is based on the current vector orientation with respect to the line voltage vector called voltage-oriented control or VOC. The VOC guarantees high dynamics and static performance via internal current control loops. The scheme decouples the converter currents into active and reactive power components. Control of the active and reactive powers is then achieved by controlling the decoupled converter currents using current controllers. One main drawback of such a system is that the performance is highly dependent on the applied current control strategy and the connected AC network conditions [1]. Another control strategy called direct power control (DPC) is based on the instantaneous active and reactive power control. In DPC, there are no internal current control loops and no PWM modulator block, because the converter switching states are appropriately selected by a look-up table based on the instantaneous errors between the commanded and measured values of the active and reactive powers. Compared to the VOC, there is a simpler algorithm, no current control loops, no coordinate transformation and separate PWM voltage modulator, no need for decoupling between the control of the active and reactive components, and better static and dynamics performance. However, among the well-known disadvantages of the DPC scheme are [2–18]: variable switching frequency (difficulties of converter and filter design); high sampling frequency needed for digital implementation of hysteresis comparators; large inductance needed between the AC source and the converter; and some problems due to the high gain of the hysteresis controllers.

In this research work a novel method for direct power control of three-phase pulse-width-modulated converters is presented. In this method hysteresis comparators and switching table are replaced by PWM voltage modulator. The required converter voltage in each sampling period is directly calculated based on only

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reference and measured values of active and reactive powers, system parameters, and the measured voltage of the AC source. Then, the PWM generator synthesizes the reference voltage and generates the switching pulses for the voltage source converter. Compared to the VOC, and conventional DPC there is a simpler algorithm, no current control loops, there is no need for decoupling between the control of active and reactive components, and finally, no hysteresis controllers are required. Simulation and experimental results show that the proposed method is more robust to sampling frequency variations than the conventional method do. Also parameter value mismatch has negligible effect on power tracking performance. The proposed strategy besides having the conventional switching table based DPC method advantages, offers many unique features such as:

- fixed and low switching frequency;
- low sampling frequency needed for digital implementation;
- simple and easy for real time implementation;
- no hysteresis controller and linear PI controller;
- small inductance needed between the AC source and the converter;
- robust to sampling frequency variations; and
- inductance values mismatch has negligible effect on active and reactive power tracking performance.

2. Principles of classical DPC

Fig. 1 shows the configuration of the direct instantaneous active and reactive power controller for the PWM converter. Direct power control is based on the instantaneous active and reactive power control loops [2,3]. With DPC there are no internal current control loops and no PWM modulator block, because the converter switching states, in each sampling period, are selected from a switching table based on the instantaneous errors between the commanded and measured or estimated values of active and reactive powers, and the angular position of the source voltage vector. In this configuration, usually, the DC link voltage is regulated by controlling the active power, and the unity power factor operation is achieved by controlling the reactive power to be zero. The DPC idea has been proposed by Ohnishi [2]. For the first time he used the instantaneous active and reactive power values as control variables instead of instantaneous three phase line currents ever used. He established first a proportional relationship between the instantaneous power values and the currents expressed in the rotational reference frame which only holds for the balanced sinusoidal operation. Since the converter voltage is related to the time derivatives of the line currents, so there is a relationship between the injected converter voltage and the time derivatives of the instantaneous active and reactive powers. Thus, the reference voltage for the PWM block is proposed in such a way that the sign of these derivatives opposes the sign of the errors in the active and reactive powers. For this purpose, hysteresis controllers are utilized which are simple and

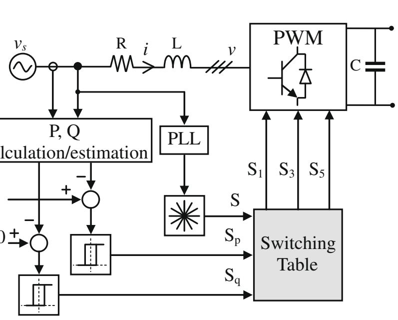

Fig. 1. Basic configuration of DPC for three-phase PWM converters [3].

have a high gain. Because this method still needs a PWM block, so it cannot yet be considered as direct, however, the principle of DPC is based on the Ohnishi's idea. The term "Direct Power Control" or DPC for the first time was used by Noguchi et al. for the control scheme depicted in Fig. 1 [3]. This method is based on selecting a voltage vector from a look-up table, Table 1, according to the errors of active and reactive powers as well as the angular position of the source voltage vector. The entries of the table which hereafter named the switching table was determined in order to minimize the errors between the commanded and measured or estimated powers in each sampling period. Also to achieve a better performance, they proposed to divide the vector space into twelve sectors and then determine the position of the source voltage vector accordingly.

The most significant drawback of the DPC is the variable switching frequency which mainly depends on the sampling frequency, the switching table structure, system parameters, reference values of the active and reactive powers, hysteresis bands, and finally the converter switching status. This variable switching frequency will produce a broadband harmonic spectrum in the AC line currents. Because of these harmonics the design of filters will be difficult. On the other hand, DPC controllers are hysteresis type. These controllers cannot guarantee the perfect tracking of a time varying signal, unless arbitrarily high sampling/switching frequencies are used. Besides, due to their high gain, they are too much sensitive to current ripples which may disturb the control. So, in order to achieve an acceptable performance, large values for the sampling frequency and the filter inductance should be selected to attenuate the current ripples. Large inductance value leads to increased cost, dimensions, weight, and losses, and also reduces the system dynamics. Above mentioned problems can be eliminated by avoiding the hysteresis controllers and also introducing a space vector modulator (SVM) in control strategy [4–8]. Moreover, the line voltage sensors can be replaced by virtual flux (VF) estimator, which introduces technical and economical advantages to the system such as: simplification, reliability, galvanic isolation, and cost reduction. In this method hysteresis comparators and switching table are replaced by linear PI controllers and SVM. The main drawback for such a system is that the performance is highly dependent on the tuning of the PI controller. Rodriguez et al. proposed a new strategy that eliminates the hysteresis controllers and switching table [9,10]. A predictive DPC is presented in their work for the control of the AC/DC/AC converter. In the proposed control strategy, the finite possible switching states of the AC/DC/AC are considered, the effect of each one on the load current and input power is evaluated, and the switching state that minimizes a quality function is selected and applied during the next sampling period. The quality function evaluates the load current error for the inverter, and the input active and reactive power error for the rectifier. Restrepo et al. conducted a similar work in which the quality function minimizes the active and reactive power errors [11,12]. Predictive approaches have also been employed in order to overcome the variable switching frequency problem of the DPC strategy [13,14]. Instead of selecting an instantaneous optimal voltage vector, these approaches select an optimal set of concatenated voltage vectors, which is the so-called "voltage-vectors sequence." The control problem is solved by computing the application times of the sequence vectors in such a way that the controlled variables converge toward the reference values along a fixed predefined switching period. In this way, constant switching frequency operation is obtained. Several authors have developed this concept in multilevel converter topologies linked to different kind of machines, but there are few predictive control applications on line-connected VSC systems. They called their proposed method P-DPC. Unfortunately, these methods require complex computation intensive and may not be viable in industrial applications. Also

Table 1
Switching table for DPC [3].

S_p	S_q	S 1	2	3	4	5	6	7	8	9	10	11	12
0	0	101	100	100	110	110	010	010	011	011	001	001	101
1	0	101	111	100	000	110	111	010	000	011	111	001	000
0	1	100	110	110	010	010	011	011	001	001	101	101	100
1	1	111	111	000	000	111	111	000	000	111	111	000	000

their performance is highly sensitive to system parameters. Some authors proposed predictive current control algorithms related to power control requirements, but these works present variable switching frequencies [15,16]. There are some reports, in which authors have tried to improve the performance of the conventional switching table based DPC, by introducing new switching tables, however, the results are not considerable [17,18].

3. Proposed DPC

3.1. Principles of proposed DPC

The proposed novel method to direct instantaneous power control of three-phase PWM converters has the following advantages:

- no hysteresis controller and linear PI controller are required and reference values in each sampling period are directly computed based on measurements and system parameters;
- decoupled control of active and reactive powers;
- no need for evaluation of any quality function or any other optimization which are time consuming calculations;
- fast calculation of reference voltage value for the modulator by using simple mathematical operations such as plus, minus, multiplication, and division;
- simple algorithm besides strong theoretical background;
- proper compensation of grid voltage variations during a sampling period;
- it operates at constant switching frequency thanks to the PWM generator, which makes the use of advanced modulation techniques possible;
- filter design is simple because of the constant switching frequency and smaller inductance value due to the elimination of the hysteresis controllers;
- low switching and sampling frequencies;
- higher dynamic behavior due to lower inductance values and fast control strategy;
- robust to sampling frequency variations; and
- inductance values mismatch has negligible effect on active and reactive power tracking performance.

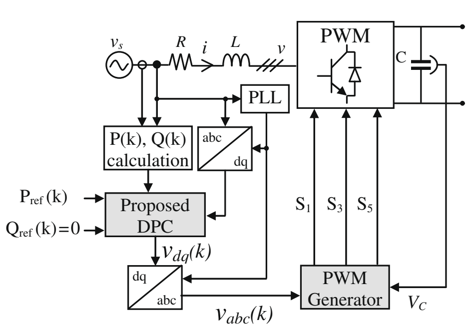

Fig. 2. Proposed DPC for three-phase PWM converters.

Fig. 2 shows the block diagram of the proposed method. In this configuration, the reference value of the active power usually comes from the DC link voltage regulator, and the unity power factor operation is achieved by controlling the reactive power to be zero.

3.2. Equations of proposed DPC

The following equations can be obtained from Fig. 2.

$$L\frac{d}{dt}\begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = -R\begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} + \begin{bmatrix} v_{sa} \\ v_{sb} \\ v_{sc} \end{bmatrix} - \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix}$$

Or:

$$L\frac{d}{dt}\vec{i}_{abc} = -R\vec{i}_{abc} + \vec{v}_{sabc} - \vec{v}_{abc} \quad (1)$$

where v is the converter voltage, v_s is the AC source voltage, i is the line current, R and L are equivalent resistance and inductance between the source and the converter, respectively. By applying the Park transformation in the stationary reference frame to (1), then we will obtain:

$$L\frac{d}{dt}\vec{i}_{\alpha\beta} = -R\vec{i}_{\alpha\beta} + \vec{v}_{s\alpha\beta} - \vec{v}_{\alpha\beta} \quad (2)$$

If one chose ω as the angular speed of the AC source voltage, and considering (3), the Eq. (2) will change to (4) in the rotating reference frame.

$$\vec{x}_{\alpha\beta} = \vec{x}_{dq}e^{j\omega t} \quad (3)$$

$$\frac{d}{dt}\vec{i}_{dq} = \left(-\frac{R}{L} - j\omega\right)\vec{i}_{dq} + \frac{1}{L}\vec{v}_{sdq} - \frac{1}{L}\vec{v}_{dq} \quad (4)$$

Eq. (4) is discretized using (5) in each small sampling period (T_{sp}) which then can be decoupled to d and q components as shown in equations set (6).

$$\frac{d}{dt}\vec{i}_{dq} = \frac{\vec{i}_{dq}(k+1) - \vec{i}_{dq}(k)}{T_{sp}} \quad (5)$$

$$\begin{cases} i_d(k+1) = \left(1 - \frac{T_{sp}R}{L}\right)i_d(k) + T_{sp}\omega i_q(k) + \frac{T_{sp}}{L}(v_{sd}(k) - v_d(k)) \\ i_q(k+1) = \left(1 - \frac{T_{sp}R}{L}\right)i_q(k) - T_{sp}\omega i_d(k) + \frac{T_{sp}}{L}(v_{sq}(k) - v_q(k)) \end{cases} \quad (6)$$

The active and reactive powers in the rotating reference frame are:

$$P(k+1) = v_{sd}(k+1)i_d(k+1) + v_{sq}(k+1)i_q(k+1) \quad (7)$$

$$Q(k+1) = v_{sq}(k+1)i_d(k+1) - v_{sd}(k+1)i_q(k+1) \quad (8)$$

During a small sampling period the AC source voltage can be assumed constant. However, if T_{sp} is not small enough the future grid voltage $v_s(k+1)$ can be calculated by compensating the angle of the voltage vector for one sampling period:

$$v_s(k+1) = v_s(k)e^{j\Delta\theta} \quad (9)$$

where $\Delta\theta = \omega T_{sp}$. Eq. (9) can be decoupled into direct and quadrature components:

$$v_{sd}(k+1) = v_{sd}(k)\cos(\Delta\theta) - v_{sq}(k)\sin(\Delta\theta) \quad (10)$$
$$v_{sq}(k+1) = v_{sq}(k)\cos(\Delta\theta) + v_{sd}(k)\sin(\Delta\theta) \quad (11)$$

By substituting (10) and (11) in (7) and (8) we will have the following equations for the instantaneous powers:

$$\begin{aligned} P(k+1) = {} & \left(1 - \frac{T_{sp}R}{L}\right)(P(k)\cos(\Delta\theta) - Q(k)\sin(\Delta\theta)) \\ & - T_{sp}\omega(P(k)\sin(\Delta\theta) + Q(k)\cos(\Delta\theta)) + \frac{T_{sp}}{L} \\ & \times ((v_{sd}^2(k) + v_{sq}^2(k))\cos(\Delta\theta) - (v_{sd}(k)v_d(k) \\ & + v_{sq}(k)v_q(k))\cos(\Delta\theta)) + \frac{T_{sp}}{L}(v_{sq}(k)v_d(k) \\ & - v_{sd}(k)v_q(k))\sin(\Delta\theta) \end{aligned} \quad (12)$$

$$\begin{aligned} Q(k+1) = {} & \left(1 - \frac{T_{sp}R}{L}\right)(P(k)\sin(\Delta\theta) + Q(k)\cos(\Delta\theta)) \\ & + T_{sp}\omega(P(k)\cos(\Delta\theta) - Q(k)\sin(\Delta\theta)) + \frac{T_{sp}}{L} \\ & \times ((v_{sd}^2(k) + v_{sq}^2(k))\sin(\Delta\theta) + (v_{sd}(k)v_q(k) \\ & - v_{sq}(k)v_d(k))\cos(\Delta\theta)) - \frac{T_{sp}}{L}(v_{sd}(k)v_d(k) \\ & + v_{sq}(k)v_q(k))\sin(\Delta\theta) \end{aligned} \quad (13)$$

The target of the control is to make the load active and reactive powers at the sampling point $(k+1)$, equal to the reference active and reactive power values currently available at the sampling point (k), so we will have:

$$P(k+1) = P_{ref}(k) \quad (14)$$
$$Q(k+1) = Q_{ref}(k) \quad (15)$$

By substituting (14) and (15) in (12) and (13) and rearranging them we will obtain an equations set which seems too complicated to be solved. In order to achieve a unity power factor the reactive power is assumed to be zero ($Q_{ref}(k) = 0$). Besides, by using a PLL, the control system will be well synchronized with the AC source voltage and the quadrant component of the source voltage will be zero ($v_{sq}(k) = 0$). Using these two assumptions, we get the following simplified equations:

$$\begin{aligned} & v_{sd}(k)v_d(k)\cos(\Delta\theta) + v_{sd}(k)v_q(k)\sin(\Delta\theta) \\ & = -\frac{L}{T_{sp}}P_{ref}(k) + \left(\frac{L}{T_{sp}} - R\right)(P(k)\cos(\Delta\theta) - Q(k)\sin(\Delta\theta)) \\ & \quad - L\omega(P(k)\sin(\Delta\theta) + Q(k)\cos(\Delta\theta)) + v_{sd}^2(k)\cos(\Delta\theta) \end{aligned} \quad (16)$$

$$\begin{aligned} & v_{sd}(k)v_d(k)\sin(\Delta\theta) - v_{sd}(k)v_q(k)\cos(\Delta\theta) \\ & = \left(\frac{L}{T_{sp}} - R\right)(P(k)\sin(\Delta\theta) + Q(k)\cos(\Delta\theta)) + L\omega(P(k) \\ & \quad \times \cos(\Delta\theta) - Q(k)\sin(\Delta\theta)) + v_{sd}^2(k)\sin(\Delta\theta) \end{aligned} \quad (17)$$

If one solve the above equations set for $v_d(k)$ and $v_q(k)$ the following results will be obtained:

$$\begin{aligned} v_d(k) = {} & v_{sd}(k) + \left(\frac{L}{T_{sp}} - R\right)\frac{P(k)}{v_{sd}(k)} - \frac{L}{T_{sp}}\frac{P_{ref}(k)\cos(\Delta\theta)}{v_{sd}(k)} \\ & - L\omega\frac{Q(k)}{v_{sd}(k)} \end{aligned} \quad (18)$$

$$v_q(k) = -\left(\frac{L}{T_{sp}} - R\right)\frac{Q(k)}{v_{sd}(k)} - \frac{L}{T_{sp}}\frac{P_{ref}(k)\sin(\Delta\theta)}{v_{sd}(k)} - L\omega\frac{P(k)}{v_{sd}(k)} \quad (19)$$

The above equations are the v_d and v_q components in the rotating reference frame of *dq* that will satisfy the control conditions of (14) and (15).

Eqs. (18) and (19) show that the *dq* components of the converter voltage can be directly controlled according to only the reference and measured values of active and reactive powers, system parameters, and the measured voltage of the AC source. The gating signals of the PWM converter will then be produced according to these *dq* voltages components. The proposed scheme's flowchart is shown in Fig. 3. In the proposed strategy, the voltage modulator has the dominant dynamics and the controller can almost reach the maximum system dynamic response. Since in the proposed DPC, in spite of the classical one which is usually called switching table based DPC, the hysteresis controllers are eliminated, so the problems of their high gain have been avoided. For example, the control sensitivity to AC current ripples is minimized and consequently the switching and sampling frequencies as well as the inductance value between the AC source and the converter can be chosen to be small. Furthermore, in the proposed DPC, the gate signals are generated by a PWM modulator instead of the hysteresis regulators, so the switching frequency is constant and much lower than the classical DPC case, and also advanced modulation techniques can be used to achieve higher efficiencies and better harmonics performance.

4. Simulation results

A digital computer simulation model has been developed in MATLAB/SIMULINK platform in order to verify the effectiveness and feasibility of the proposed configuration and its control strategy. The system parameters are: $L = 10$ mH, $R = 200$ mΩ, $C = 470$ uF, $v_s = 70$ V_{peak} and $f_s = 50$ Hz and the carrier frequency for SPWM generation is 4.5 kHz. In order to evaluate the system performance, extensive simulations have been done based on the proposed strategy in steady-state and transient conditions. The validity of the proposed system will be verified by simulations and will be compared with the simulation results of the classical switching table based DPC. Figs. 4–7 show the steady-state and transient response of both classical and proposed DPC schemes for two different sampling frequencies. These waveforms for classical DPC and proposed DPC schemes confirm the superiority of the proposed method in providing more precise current control with minimum distortion and less harmonic noises (THD) and at the same time, more accurate regulation and less distortion in the output active and reactive powers. Figs. 4 and 5 show the transient response of both STB–DPC and the proposed DPC schemes. At $t = 0.025$ s the reference value of

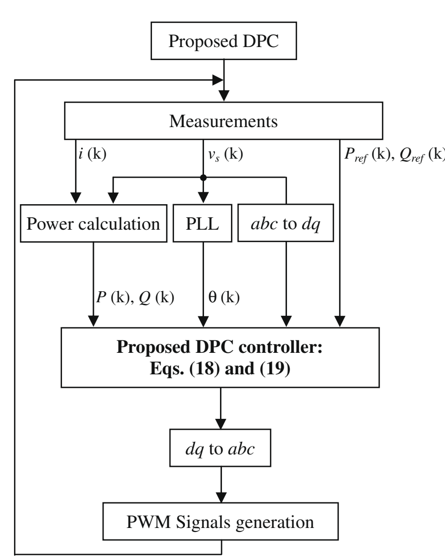

Fig. 3. Flowchart of the proposed DPC.

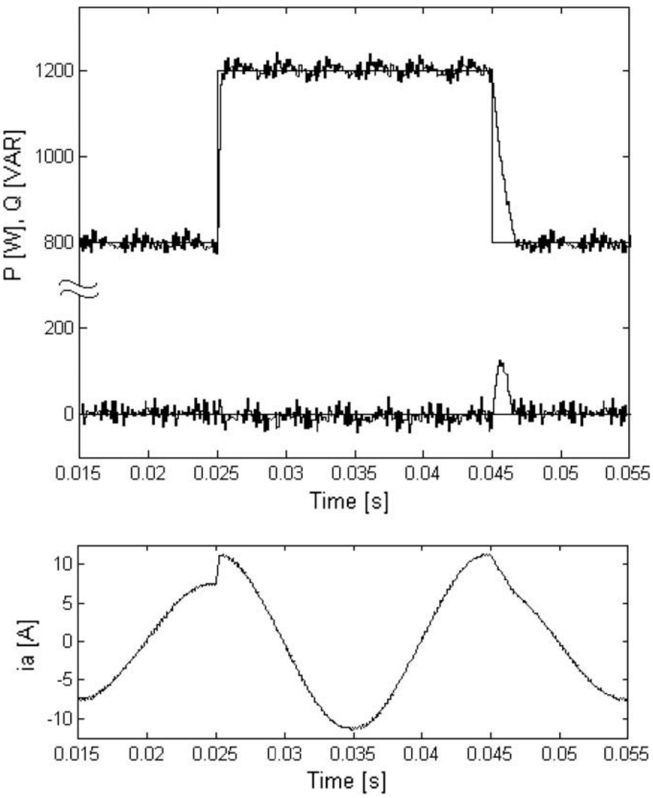

Fig. 4. Simulated waveforms for step change of active power; proposed DPC and f_{sp} = 10 kHz.

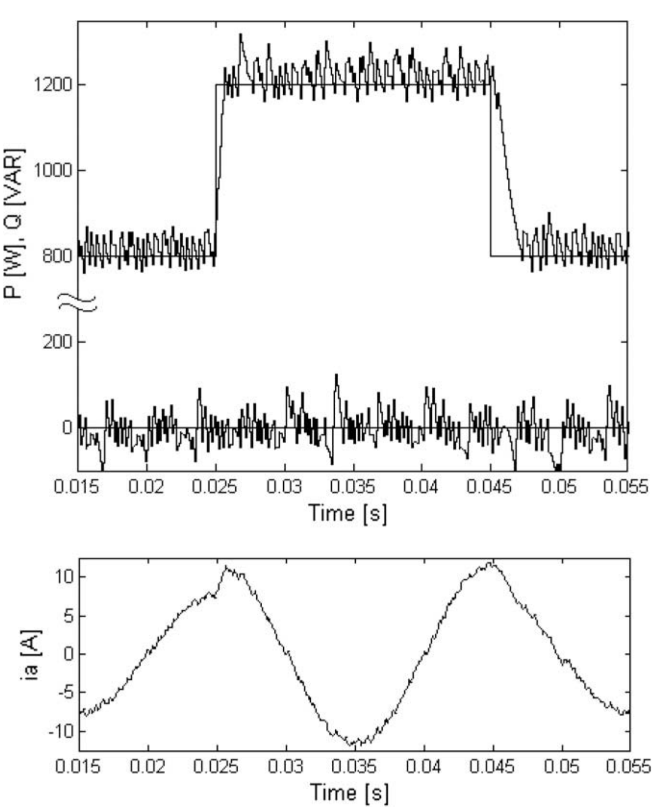

Fig. 5. Simulated waveforms for step change of active power; classical DPC and f_{sp} = 10 kHz.

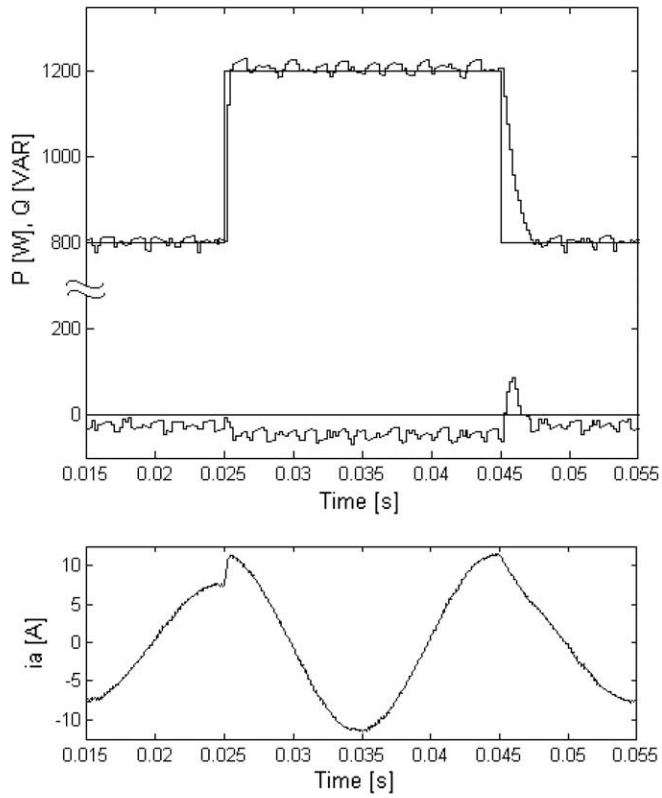

Fig. 6. Simulated waveforms for step change of active power; proposed DPC and f_{sp} = 5 kHz.

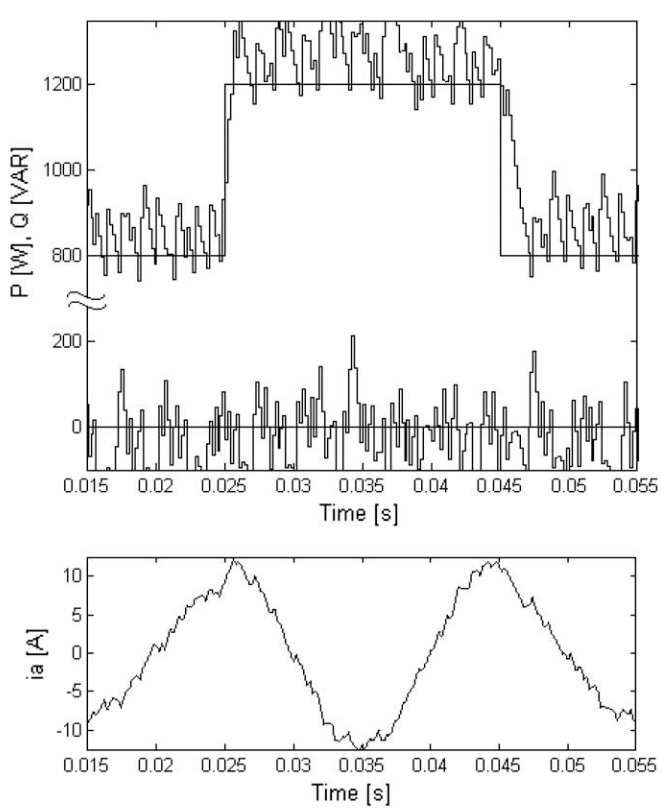

Fig. 7. Simulated waveforms for step change of active power; classical DPC and f_{sp} = 5 kHz.

the active power stepped up to 1200 W. Fig. 4 shows that the estimated value of the active power converges to the reference value in 0.001 s, whereas this time for the STB–DPC is around 0.003 s as shown in Fig. 5. Also it can be seen from Fig. 7 that the behavior of the classical DPC is not acceptable with reduced values of sampling frequency while with the same sampling frequency in Fig. 6 shows the proposed method provides acceptable performance. This can simplify the microcontroller tasks in an industrial application. In other words the proposed method is more robust to sampling frequency variations than the classical one do. The harmonic spectrum of the phase current i_a is shown in Figs. 8 and 9. One can recognize that the classical DPC in Fig. 9 has broadband harmonic spectrum that spread over a wide low frequency range. Because of these low frequency harmonics the design of filters will be difficult in order to avoid possible grid resonances. Table 2 summarizes the total harmonic distortion (THD*i*) values of both proposed and classical methods for two different sampling frequencies which confirm the superiority of the proposed method. In another study, the effect of the parameters' values mismatch on active power and reactive power tracking performance is presented in Fig. 10. It can be seen that the ripple is slightly increased when a wrong value is used in the model; however, the active power reference is well achieved in all cases. Errors in the resistance value have negligible effect in the performance of the proposed DPC scheme.

5. Experimental results

Fig. 11 shows the experimental setup of the proposed scheme. As depicted in Fig. 12, the prototype IGBT converter includes a power circuit module, an interfacing and sensing module, PWM generator system and IGBT driver modules. The converter is equipped with software and hardware protections including over-voltage, over-current and over-temperature protections. The control strategy is developed using MATLAB/SIMULINK Real Time Windows Target and the interface between the model and the converter module is provided using a high speed data acquisition interface card. The sampling frequency for experimental implementation was chosen 10 kHz, and experimental results are obtained with the same parameters used in the simulations except that the value of input inductance is 15 mH. The DC side of the converter is connected to a fixed 42 Ω resistive load. Experimental results for the proposed DPC inside a DC link voltage control loop are presented in Fig. 13. The dynamics of the PWM converter with the proposed scheme is fast and there is no coupling between the active and reactive powers. This figure shows the transient response

Table 2
THD*i* for different values of sampling frequency (Q_{ref} = 0, P_{ref} = 1000 W).

f_{sp}	Classical DPC (%)	Proposed DPC (%)
10 kHz	4.17	1.82
5 kHz	7.40	1.83

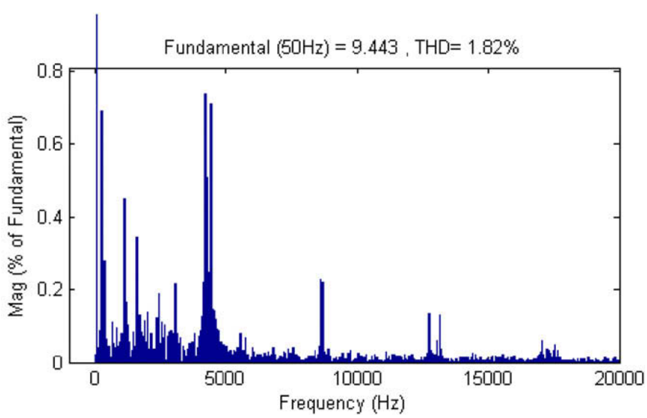

Fig. 8. Simulated line current harmonic spectrum (P_{ref} = 1000 W); proposed DPC and f_{sp} = 10 kHz.

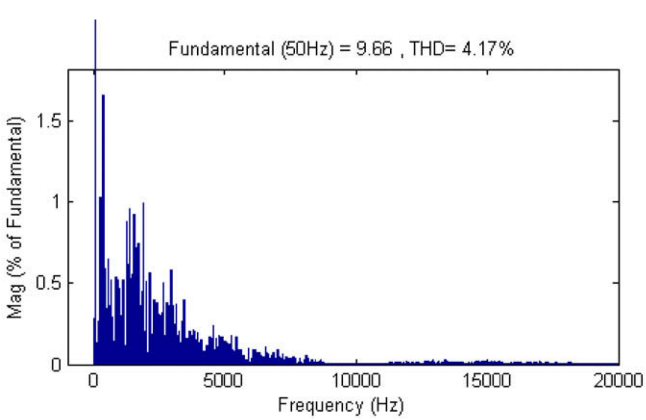

Fig. 9. Simulated line current harmonic spectrum (P_{ref} = 1000 W); classical DPC and f_{sp} = 10 kHz.

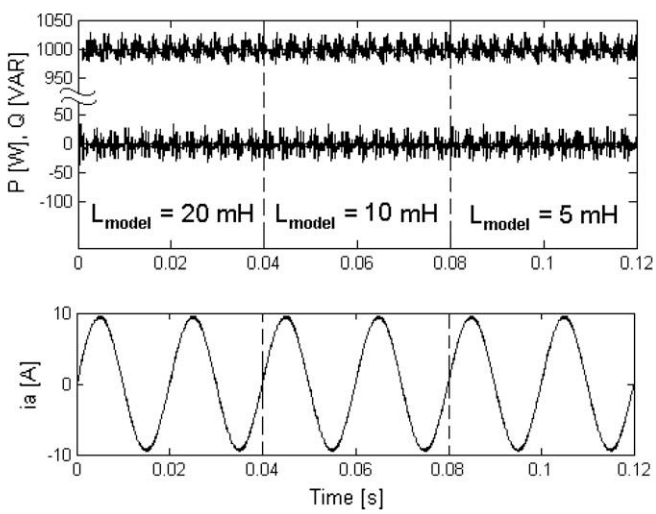

Fig. 10. Simulated behavior of the proposed DPC considering errors in the inductance value ($L_{circuit}$ = 10 mH).

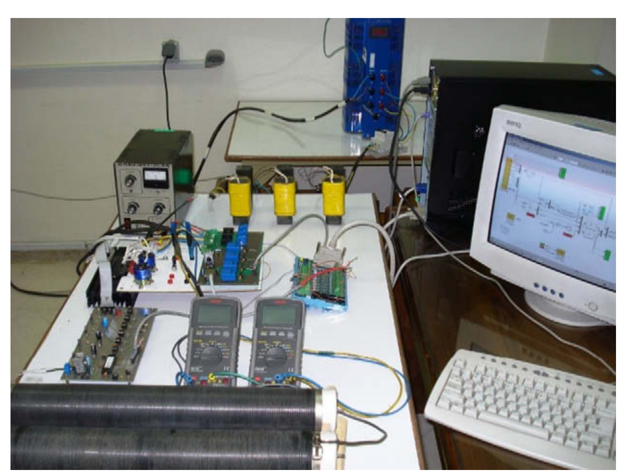

Fig. 11. Laboratory setup.

of the proposed DPC scheme due to step change of dc link voltage. At $t = 5$ s the reference value of the active power stepped up to 700 W Fig. 13 shows that the measured value of the active power converges to the reference value in less than 0.01 s. The experimental behavior of the proposed DPC considering mismatch in the inductance value is studied by disconnecting the external PI controller used for voltage control. In general, the measured and simulated results depicted in Figs. 14 and 10, respectively, show good agreement. At low and high inductance values, there is a slight active power tracking error with wrong values of input inductance. Steady-state operation, which confirms the proper operation of the proposed DPC, is presented in Fig. 15. The quality of practical converters is normally evaluated in terms of total harmonic distortion (THD). Based on the IEEE standard this value should be less than 5%. The proposed strategy, also, provided an acceptable THD value.

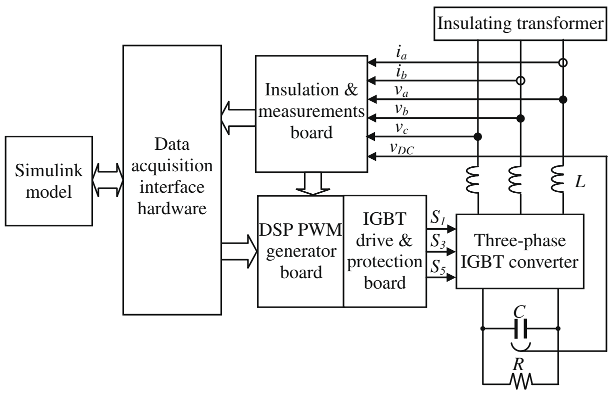

Fig. 12. Configuration of laboratory setup.

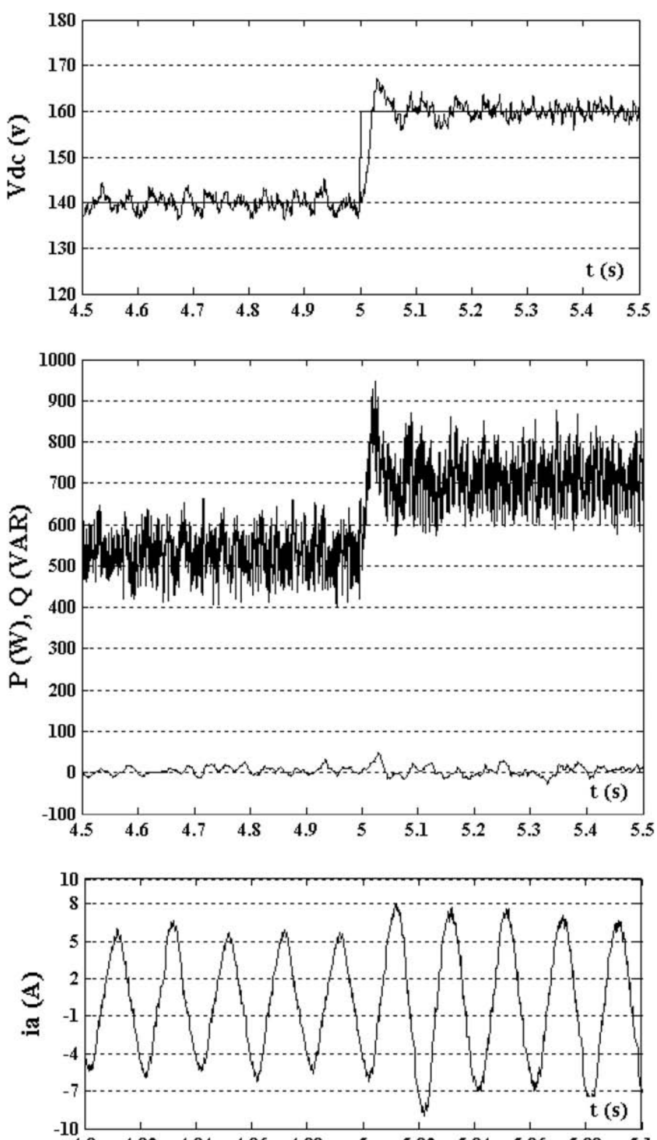

Fig. 13. Experimental waveforms for step change of DC link voltage.

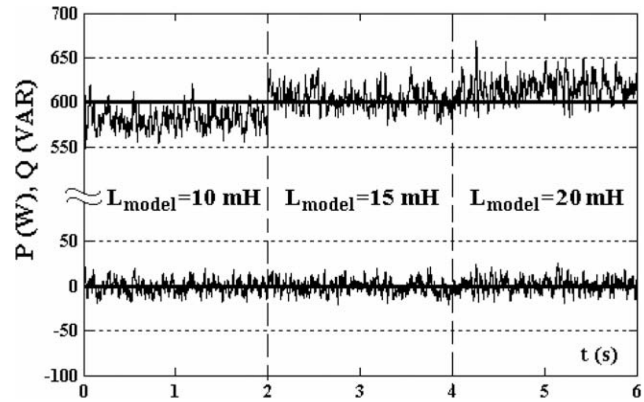

Fig. 14. Experimental behavior of the proposed DPC considering mismatch in the inductance value ($L_{circuit}$ = 15 mH).

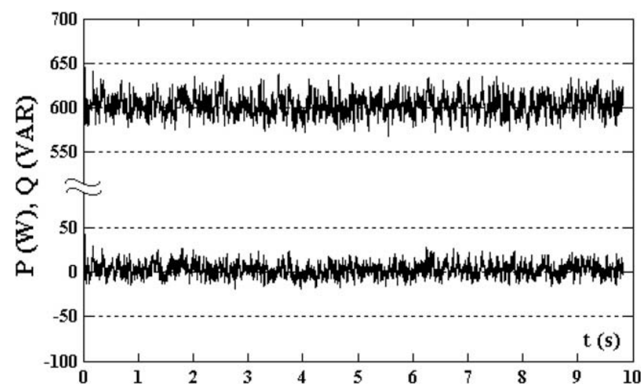

Fig. 15. Experimental results for steady-state operation; active power, and reactive power (P_{ref} = 600 W).

6. Conclusion

In the proposed method the hysteresis comparators and the switching table are replaced by a PWM modulator. Simulation and experimental results confirm the superiority of the proposed method in providing more precise power control with minimum distortion and harmonic noises (THD*i*), and at the same time, less distortion in active and reactive powers and narrower current harmonic spectrum in compare to conventional DPC scheme. Simulation results confirm that the proposed strategy is more robust to sampling frequency variations than the classical one do. In other words the behavior of the classical DPC is not acceptable with reduced values of sampling frequency while with the same sampling frequency the proposed method provides acceptable performance. From simulation and experimental results the proposed strategy besides having the conventional methods advantages offers many unique features such as: more precise control and better regulation performance than the classical switching table based DPC; higher dynamic behavior (less convergence time); has narrower harmonic spectrum compared to the classical DPC; it provides less total harmonic distortion (THD*i*); the switching frequency is fixed, also it requires less sampling and switching frequencies and inductance value which is mainly due to a simpler and more precise algorithm which eliminates the hysteresis controllers and also compensates the source voltage variation during a sampling period; it is robust to sampling frequency changes.

Despite these advantages, the main shortcoming of this topology may be a slight sensitivity to parameter mismatch especially to line inductance error, where it shows some active power tracking errors.

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