SINGLE-PHASE TO THREE- PHASE CONVETER FOR LOW COST AC MOTOR DRIVES

T.S. Radwan*, Z.M.El-Barbary** and S.A. Mahmoud*
*Electrical Engineering Department, Faculty of Engineering, Minoufiya University, Egypt
** International Steal Rolling Mill "ISRM", Sadat City

ABSTRACT

In this paper, a single- phase to three- phase converter is proposed to provide variable output voltage and frequency. The proposed topology employs only six IGBT switches, which form the front-end rectifier and the output inverter for the one step conversion from single-phase supply to output three-phase supply. The front-end rectifier permits bidirectional power flow and provides excellent regulation against fluctuations in source voltage. Moreover, it incorporates active input current shaping feature. An easy method to implement control strategy is proposed. This control strategy ensures nearly unity input power factor with sinusoidal input current over the operating range. Based on vector control technique, the proposed converter is used for speed control of three-phase induction motor. A low cost of motor drive can be achieved using the proposed technique. Simulation and experimental results are carried out to analysis and explore the characteristics of the low cost drive system.

KEYWORDS

Voltage source reversible rectifier, Four-switch inverter, Induction motor, Vector control, Digital Signal Processor (DSP).

1. INTRODUCTION

In rural electric systems and remote areas the cost of bring three-phase power to a remote location is often high. This is due to high cost of a three-phase extension. Further, the rate structure of a three-phase service is higher than the single-phase service. A single-wire earth return transmission line has proved to be a cost-effective solution for delivering power to rural communities. Single-phase power is adequate for many domestic applications such as lighting, heating and powering small appliances. However, problems arise as soon as applications demand the use of medium electric motors. These applications include pumps for

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irrigation or drinking water, machinery for mills or small industries.

The single-phase induction motor, up to sizes of 5kw, has been an ubiquitous motor for many years in applications where a single- phase supply is readily available, such as in the household. These are used in refrigerators, washing machines, fans, food mixers, pumps and air conditioners. In most of these applications the motor is actually a two-phase motor. Usually the motor runs at one, two or at most three speeds obtained through manual intervention. Obviously, the motor operates at non-optimum efficiency and at low power factor. Variable speed operation has been obtained through voltage control using triacs or back-to-back thyristors. However, these suffer from large harmonic injection into the supply network and low power factor.

Fixed speed operation often means that the process, which is not controlled to the desired extent, consequently can not be made to respond well to what is desirable and adequate. A wide speed range, truly variable-speed motor, frees the designers to come up with better operating features. This can be achieved by using three-phase motor, which are more readily available. Therefore, single-phase to three-phase converters is a suitable solution for these situations where three-phase power is not available. Various methods, including static phase converter, rotary phase converters, reduced topology power electronic converters and novel motor designs have all been investigated [1-5].

Variable-speed drives using three-phase inverters and ac motors have already found widespread practical application in various forms. Certain applications, however, still require a further cost reduction for the drive. The reduction of the number of power semiconductor device components in the converter should be the main consideration for reducing the cost of the control subsystem. The use of transistors instead of thyristors eliminates the forced commutating components and presents distinct advantages for drives in the power range up to 50KVA. A further reduction in components is possible by considering alternatives to the conventional three- phase bridge configuration for a voltage- fed inverter, since this circuit uses six switching devices and six reactive power diodes. These considerations lead to a bridge circuit with only four switching devices and four reactive power diodes. It has been shown that a two-level current control can be implemented to yield quasi-sinusoidal currents in the three- phase load [6,7].

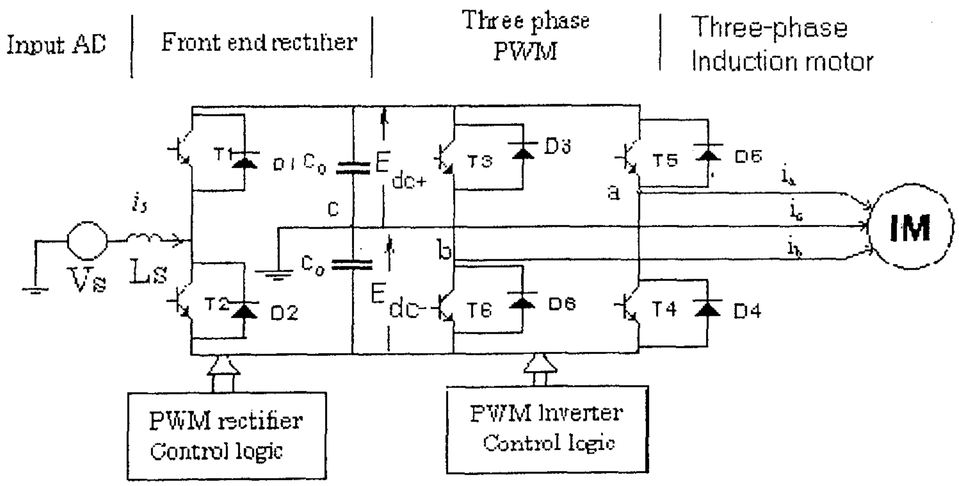
This paper presents single-to three-phase converter for low cost ac motor drives. The proposed system is based on vector control technique for operation over wide speed range. The proposed converter configuration employs only six IGBT transistors for rectifier and inverter structure. The proposed configuration incorporates front-end half bridge active rectifier. This rectifier provides the dc-link voltage with active input current shaping feature. A four-switch inverter with split capacitors in the dc-link provides a balanced three-phase output to the motor. The proposed converter draws near sinusoidal current from the ac mains at close to unity power factor and therefore satisfies strict harmonic current standards. Bi-directional power flow is possible between the ac mains and dc-link. This feature provides excellent voltage regulation against voltage fluctuations. A laboratory drive system is built and tested using the proposed converter to explore the most important feature of the low cost drive.

2. PROPOSED CONVERTER

The proposed single-phase to three-phase converter is configured as shown in Fig.1. The switches T_1 and T_2 form the front-end rectifier. Two split capacitors form the dc-link. The output inverter converts the dc-voltage to a balanced three-phase output with adjustable voltage and frequency. This inverter is configured with four switches T_3, T_4, T_5 and T_6, respectively. Two output phases are taken from the inverter legs directly where the third output is taken from the midpoint of the two capacitors.

2.1 Front-end rectifier

The front-end rectifier converts a single-phase utility ac supply voltage to dc-voltage. A boost inductor, L_s, is connected in series with the utility supply voltage. The power transistors T_1 and T_2 are switched on using current forced control (CFC) strategy that enables the device to operate with a sinusoidal line current. As a result, little distortion, unity power factor and regulated dc busbar voltage can be achieved. Moreover, this control strategy permits power transfer in either direction between ac mains and dc busbar voltage. So, this rectifier calls single-phase voltage source reversible rectifier (VSRR).



The two capacitors (designated C_0) initially have to charge up to the peak voltage of the utility supply via the inverse parallel diodes across the transistors. The two capacitors must be large enough to appear as an essentially constant dc voltage source with low ripple contents. The large size of C_0 also means that some form of soft start must be included during the initial charge up to prevent transient inrush current from harming circuit components. Once the capacitors have fully charged and the diodes are reversing biased, it becomes virtually constant dc voltages. The output voltages E_{dc+} and E_{dc-} can be kept constant by connecting the capacitors in series with the voltage source, Vs, via the two transistor switches T1 and T2. Note that only one-transistor turn on at a time, or else a short circuit of the dc voltages will occur. If a transistor that conducting current is turned off, owing to the inductive nature of the circuit, the current instantaneously freewheels through the diode across complementary transistor. If the same transistor is turned back on, then the conducting diode will reverse biased and the current will switch to flow back through the transistor. In a CFC scheme, whenever a transistor is turned on the complementary transistor is always turned off. So, the two transistors / diodes essentially behave as a two pole bi-directional switch. This switch conveniently is represented as switching logic variable NS. Using this logic variable, the circuit differential equations can be written as:

$$i_s = \frac{1}{L_s} \int (V_s - NS * E_{dc+} - (1-NS) * E_{dc-}) dt \tag{1}$$

$$E_{dc+} = \frac{1}{C_o} \int (NS * i_s - i_l) dt \tag{2}$$

$$E_{dc-} = \frac{1}{C_o} \int ((1-NS) * i_s + i_l) dt \tag{3}$$

Where, i_l is the load current which may be i_a or i_b or i_c. NS=1,if T1 or D1 is on and NS=0, if T2 or D2 is on. From the above equations it can be seen that if T1 is on the supply current i_s will decrease. On the other hand, if T2 is on the current will increase. Hence, the current can be forced to track a reference waveform i_r simply by switching on T2 if the current is higher than i_r, or by switching T1 on if the current is lower than i_r. For the device to operate with a sinusoidal input current and unity power factor it is necessary to produce a sinusoidal-current reference waveform that is either in phase or 180^0 out of phase with ac utility supply voltage. In addition, the dc voltage E_{dc} must be maintained equal to a reference E_{dcr} under all load conditions. Both these can be obtained using the control scheme of Fig.2. Comparing a dc reference, E_{dcr}, with the actual dc–link voltage, E_{dc}, the output error, e, is filtered and modified by a PI controller. The filter output, U, multiplied by a unit vector signal derived from the supply voltage to produce the required current reference i_r. The current reference is then compared with the actual supply current to determine which transistor should be turned on. If E_{dc} is too low then the error is positive and i_r is produced in phase with V_s .The power is transferred from the ac to the dc side (rectifying) and consequently increasing E_{dc}.

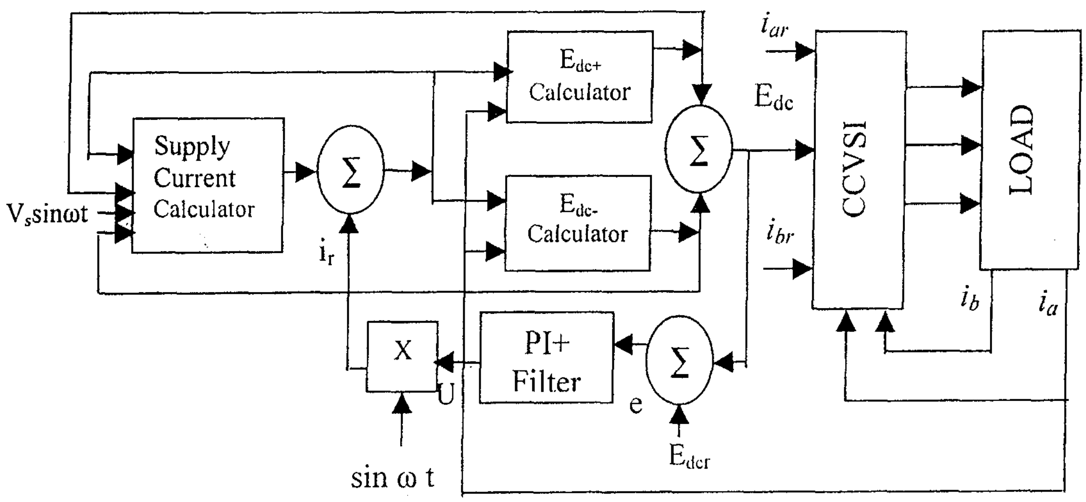


Fig. 2 Block diagram of the control scheme

Alternatively, if E_{dc} is too high then the error is negative and i_r is produced 180^0 out of phase with V_s. So the power is transferred from the dc to the ac side (inverting) to reduce E_{dc}. The control loop including the second order harmonic filter has a total transfer function:

$$\frac{U_{(s)}}{e_{(s)}} = \frac{K_C(S+45)}{S(S+1.8)} \tag{4}$$

Where,

$$e_{(s)} = E_{dcr}(s) - E_{dc}(s) \tag{5}$$

$$E_{dc} = E_{dc+} - E_{dc-} \tag{6}$$

And

$$i_r = U_{(t)}.sin(\omega t) \tag{7}$$

2.2 Four-switch three-phase output inverter

The output side of the proposed single-phase to three-phase converter consists of a four switch (T_3 to T_6) inverter. The center point of the capacitors forms the third phase "c", where the current in this phase is the result of the currents in the two controlled phases. A detailed comparison of the four-switch inverter with the conventional six-switch inverter configuration is given in [8]. Two control possibilities exist to control the four-switch bridge inverter, i.e., two- level current control to force the two controlled phases currents to sinusoidal, or using PWM to control the voltages applied to the three-phase quasi- sinusoidally. The two- level current control of the four-switch bridge inverter used to control the load current by forcing it to follow a reference one. This is achieved by the switching action of the inverter to keep the current within the hysteresis band. The load currents are sensed and compared with respective command currents using two independent hysteresis comparators. The output signals of the comparators are used to activate the inverter power switches. This controller is simple and provides excellent dynamic performance.

In this paper, the four-switch three-phase output inverter is controlled using a fixed band hysteresis current control scheme, which has the following mathematical model:

$$i_{ref} = i_{max} sin(\omega t) \tag{8}$$

$$i_{up} = i_{ref} + H \tag{9}$$

$$i_{lo} = i_{ref} - H \tag{10}$$

Where, i_{ref} is the reference current (may be i_{ar} or i_{br}), i_{up} is the upper band, i_{lo} is the lower band and H is the hysteresis band. For i_{ar}>0.0: if $i_a > i_{up}$, then NA=0, this means that the inverter output voltage switches to negative in order to reduce the line current. In the same manner if $i_a < i_{lo}$, then. NA=1, where the inverter output voltage switches to positive in order to increase the line current. The same sequence is followed for phase b. Hence, the control logic for the two phases (a and b) are given as follows:

For i_{ar}>0.0: if $i_a > i_{up}$, then NA=0, else if $i_a < i_{lo}$ then NA=1
For i_{ar}<0.0: if $i_a < i_{up}$, then NA1=1, else if $i_a > i_{lo}$ then NA1=0
For i_{br}>0.0: if $i_b > i_{up}$, then NB=0, else if $i_b < i_{lo}$ then NB=1
For i_{br}<0.0: if $i_b < i_{up}$, then NB1=1, else if $i_b > i_{lo}$ then NB1=0

where, NA1 and NB1 are complementary of NA and NB, respectively. The modulated phases voltages of four switch inverter are introduced as a function of switching logic NA, NA1, NB and NB1 of power switches by the following relations:

$$V_a = NA.\frac{E_{dc}}{2} + NA1.\frac{E_{dc}}{2} + V_c = (2.NA - 1).\frac{E_{dc}}{2} + V_c \tag{11}$$

$$V_b = NB.\frac{E_{dc}}{2} + NB1.\frac{E_{dc}}{2} + V_c = (2.NB - 1).\frac{E_{dc}}{2} + V_c \tag{12}$$

$$V_c = -(V_a + V_b) = (1 - NA - NB).\frac{E_{dc}}{3} \tag{13}$$

3. CONTROL SCHEME OF LOW COST AC MOTOR DRIVE

The control scheme of low cost ac induction motor drive is shown in Fig.3. It incorporates the speed controller which receives the error signal between the preset speed and the actual measured speed of the motor shaft and then generates the torque command (T_e^*) through a PI speed controller. This torque command produces the quadrature current command I_{qs}^{*e} in the synchronous reference frame (SYRF) [9]. The direct current I_{ds}^{*e} is sets by the rotor flux level, λ_{dr}^{*e}. This flux level is calculated according to the method described in [10]. The two current commands are then transformed to the stationary reference frame (STRF) with the aid of the calculated command angle (θ_e^*).This command angle is

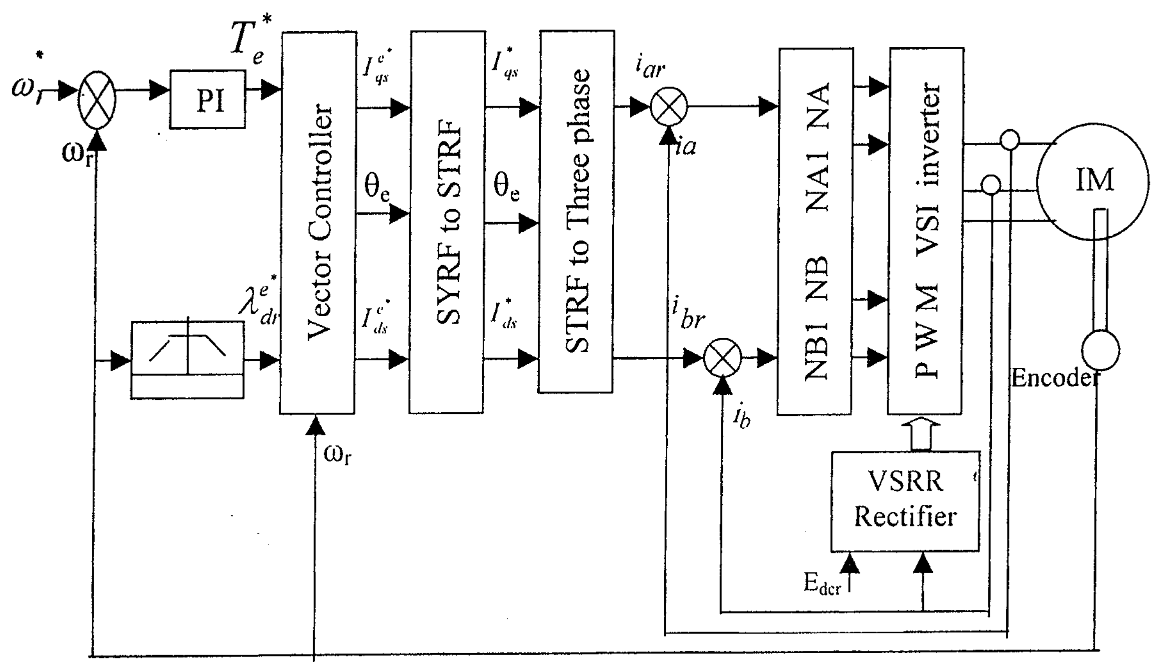


Fig.3 Block diagram of the proposed drive system

calculated in such away that aligns the d^e-axis of SYRF with the rotor flux axis. The two stator current commands, I_{qs}^{*e} and I_{ds}^{*e} in the SYRF are then transformed to STRF and then transformed to three phase references current i_{ar}, i_{br} and i_{cr}. Only, two currents reference i_{ar} and i_{br} required for the hystersis current controller that generates the switching function for the pulse width modulated voltage source inverter. The VSI takes the dc voltage from the front-end rectifier.

3.1 Motor model and vector control

The mathematical model of three –phase squirrel cage induction motor in d^e-q^e is described in [9] as:

$$\begin{bmatrix} V_{qs}^e \\ V_{ds}^e \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_s + pL_\sigma & \omega_e L_\sigma & p\frac{L_m}{L_r} & \omega_e \frac{L_m}{L_r} \\ -\omega_e L_\sigma & R_s + pL_\sigma & -\omega_e \frac{L_m}{L_r} & p\frac{L_m}{L_r} \\ -R_r L_m & 0 & R_r + pL_\sigma & (\omega_e - \omega_r)L_m \\ 0 & -R_r L_m & -(\omega_e - \omega_r)L_m & R_r + pL_\sigma \end{bmatrix} \begin{bmatrix} I_{qs}^e \\ I_{ds}^e \\ \lambda_{qr}^e \\ \lambda_{dr}^e \end{bmatrix} \tag{14}$$

The electromechanical equations are given by

$$T_e - T_L = J\frac{d\omega_r}{dt} + B\omega_r \tag{15}$$

$$T_e = \frac{3}{2}\cdot\frac{P}{2}\frac{L_m}{L_r}(I_{qs}^e \lambda_{dr}^e - I_{ds}^e \lambda_{qr}^e) \tag{16}$$

Equation (16) denotes that the torque can initially proportional to the quadrature component of the stator current I_{qs}^{*e} if the q^e-axis component of the flux becomes zero (d^e-axis is aligned with the rotor flux axis), and the d^e-axis component λ_{dr}^{*e} is kept constant. This is the philosophy of the vector control technique. In accordance, Eq.(16) is linearized as :

$$T_e = K_t \left|\lambda_{dr}^e\right| I_{qs}^e \tag{17}$$

This equation is similar to that of the separately excited dc motor. The angular slip frequency command (ω_{sl}^*) is:

$$\omega_{sl}^* = \frac{L_m}{\tau_r^*}\cdot\frac{I_{qs}^{e*}}{\lambda_{dr}^{e*}} \tag{18}$$

Also from Eq.(14)

$$I_{ds}^{e*} = \frac{1}{L_m}(1 + \tau_r^* p)\lambda_{dr}^{e*} \tag{19}$$

Angular frequency is obtained as follows,

$$\omega_e^* = \omega_r^* + \omega_{sl}^* \tag{20}$$

$$\theta_e^* = \int \omega_e^* . dt \tag{21}$$

The torque producing current component is calculated from:

$$I_{qs}^{e*} = \frac{1}{k_t}\,\frac{(\omega_r^{*} - \omega_r)}{\lambda_{dr}^{e*}}\,\frac{K_{ps}[1+\tau_{cs}S]}{\tau_{cs}S} \tag{22}$$

4. HARDWARE IMPLEMENTATION

The induction motor drive system is controlled using dSPACE DS1102 controller board [11]. This board is specifically designed for the development of high-speed multivariable digital controllers and real time simulations as shown in Fig.4. The feedback signals to the controller board are the measured motor line currents and the motor shaft angular positions. Also, the supply voltage and supply current are fed to the board. The actual currents are measured using Hall-effect devices. The currents are then buffered and fed to the A/D converters on the board. An optical incremental encoder installed at the motor shaft measures the motor shaft position. The encoder has a resolution of 2048 pulses/rev. to enable for fast and accurate speed control. The outputs of the controller board in the form of digital pulses are sent directly to the base drive circuit of the converter power transistors.

5. SIMULATION AND EXPERIMENTAL RESULTS

The simulation of the entire drive system of Fig.3 is carried out using Matlab-simulink ® toolbox. The motor parameters are given in Appendix

- ***Step change in the reference speed***

The response due to a step change in the command speed is used to evaluate the performance in terms of steady state errors and stability. The motor is subjected to step increase and decrease in the reference speed under loading conditions to evaluate the performance. Figure5(a) shows the rotor speed response with a command speed of 1000 rpm at full load. At t=1 second, the speed reference has been changed to 1100 rpm and returned to 1000 rpm again after 1second. It can be seen that the rotor speed is accelerated and decelerated smoothly to follow its reference value with nearly zero steady state error. Figure 5(b) shows the motor developed torque. The torque shows correspondingly increase and decrease during the step changes in the reference speed due to the dynamic states. The motor phase current is illustrated in Fig.5(c) and shows good dynamic response. Figure5(d) shows that the dc-link voltage is not affected by speed changes. This is one aim of rectifier design. It is shown that the control of the rectifier works independently. While, Fig. 5(e) shows the supply current with supply voltage. It is noticed that the supply current is in phase with supply voltage even during the period of speed changes. This means that unity power factor can be achieved irrespective of the motor dynamics.

- ***Load change***

The ability to withstand disturbances in IM control system is another important feature. A step change in the motor load is considered as a typical disturbance. Figure 6(a) shows the speed response when a full load is applied for 500ms, The motor started at no load and the full load is applied after one second. After 500ms the load is released. It is clear from Fig.6(a)that the speed is recovered very fast with acceptable under and overshoots .The corresponding developed torque is shown in Fig.6(b). A motor phase current is increased with applying the load and vise versa as shown in Fig.6(c). Figure 6(d) shows the supply current with supply voltage. It is noticed that the supply current is in phase with supply voltage a long the period of load changes. This ensures unity input power factor for different operating conditions. The dc-link voltage waveform is shown in Fig.6(e), it is clear that the voltage is not affected by motor loading and has good dc voltage regulation.

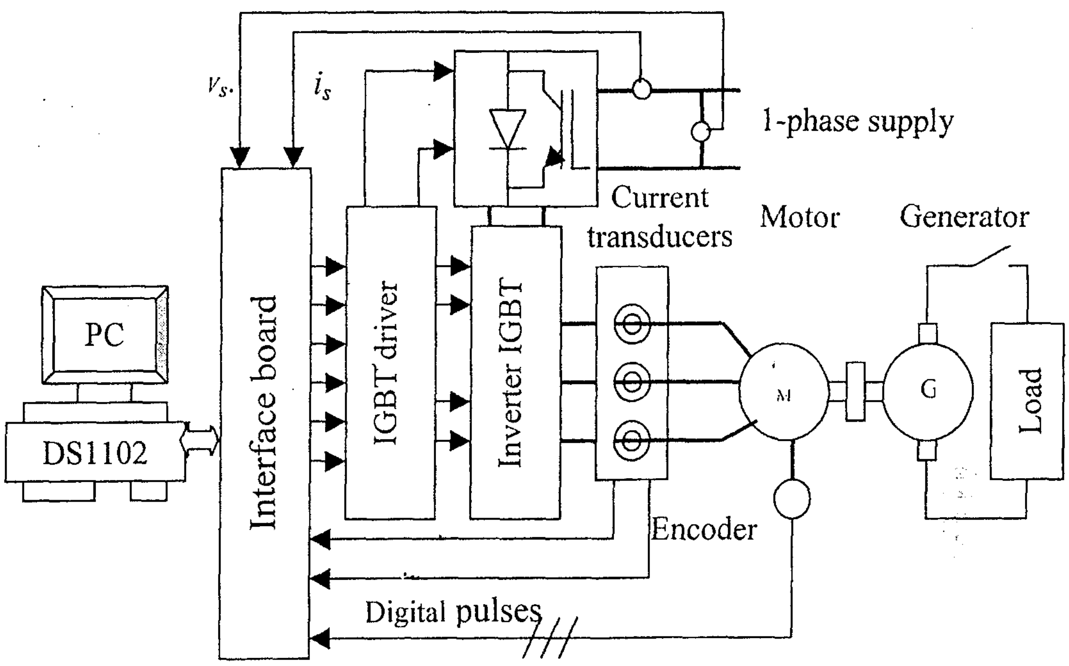


Fig. 4 Experimental set-up for DSP-based control of induction motor

The experimental investigation has been carried out to verify the validity of the control scheme of the drive system. The results were obtained at different operating points including drive response due to load changes and step change in the speed command. Figure 7 shows the experimental drive response when a full load is applied. The motor started with no load and after 900ms a full load is applied for 600ms and then is released. It can be seen that, undershoot due to load impact and overshoot due to load release are limited to an acceptable values as shown in Fig.7(a). The motor torque is show in Fig.7(b). Figures7(c) and (d), show the response of q-axis and d-axis stator current components during the load changes. It is noted that the quadrature axis current is changed related to the load and the direct axis current is maintained constant during the load change. This response clarifies that the motor is being driven under correct rotor field oriented control. Figure 8(a) shows the step change in speed from 900 rpm to 1000 rpm. It is shown that the rotor is accelerated smoothly to follow the speed reference command with nearly zero steady state error. It is noted that the system is fully damped. Figure 8(b) and (c) show the motor phase a and b currents waveforms. It is noted that the current is increased during the step change of speed.

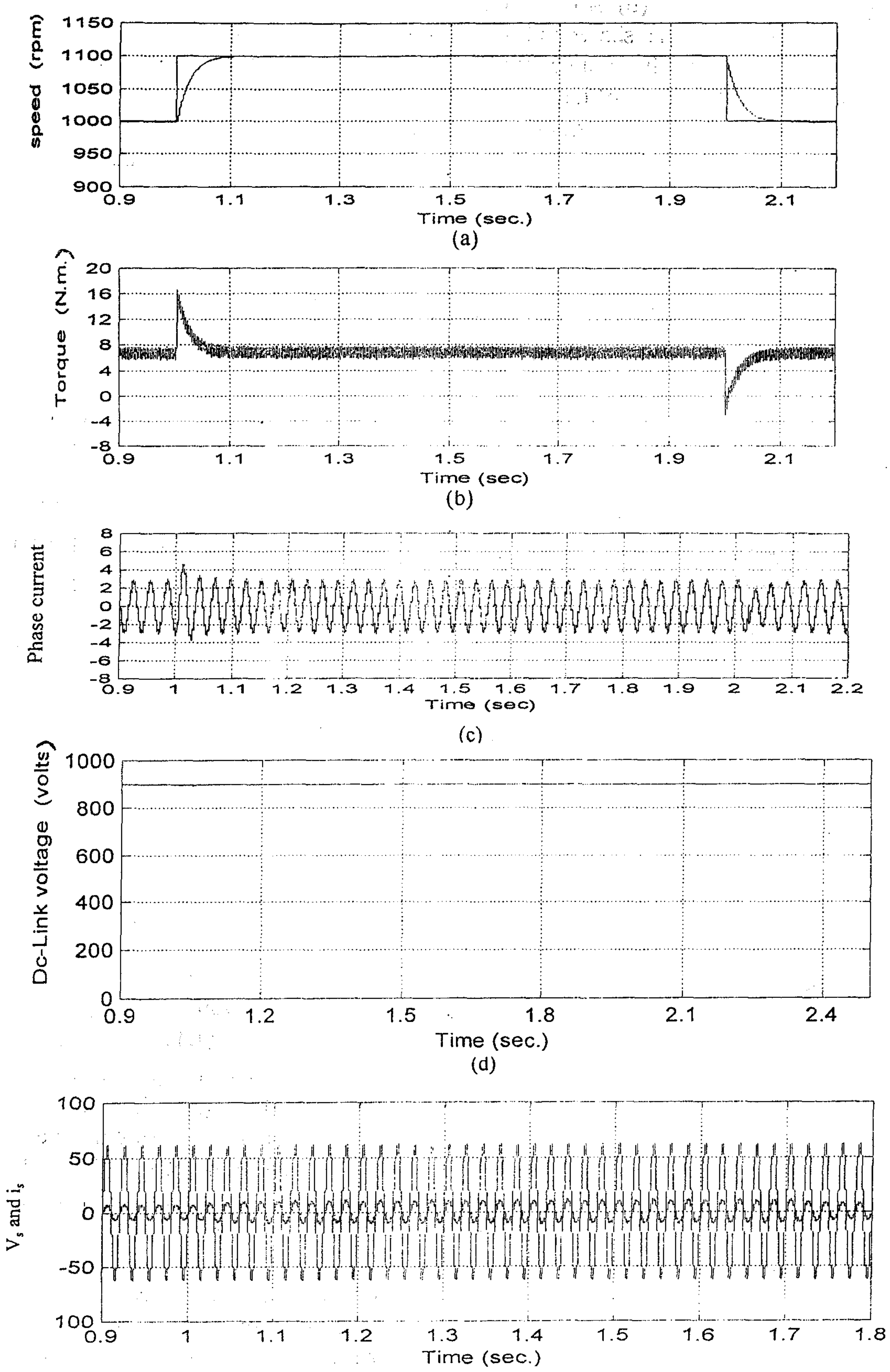
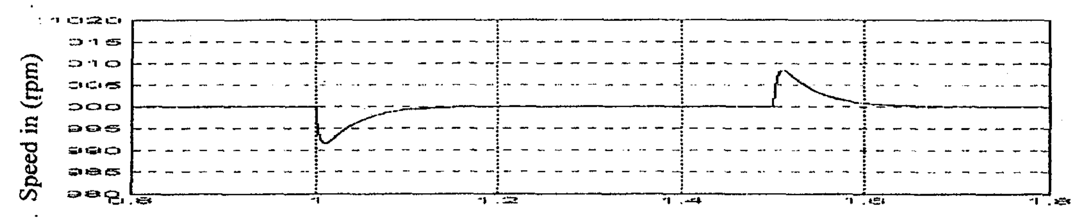
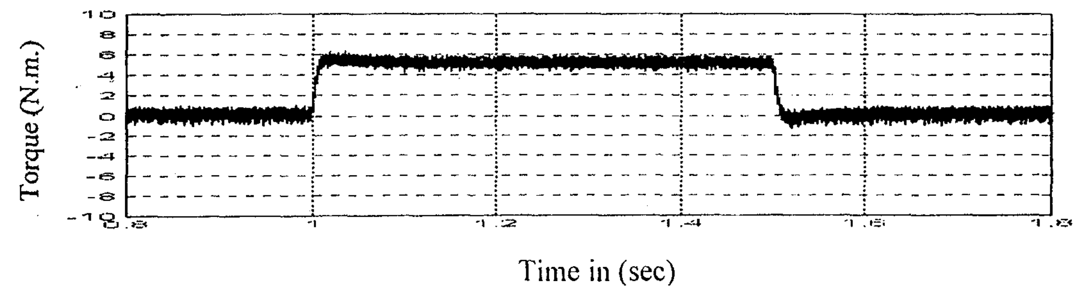


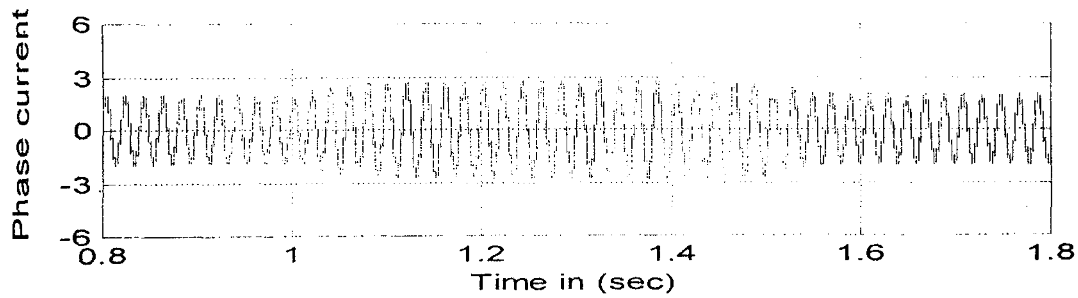
Fig.5 Simulation drive response for step changes in speed reference at full load:(a) Speed response; (b) Torque response; (c) motor phase current (d) Dc-link voltage (e) Supply current and supply voltage



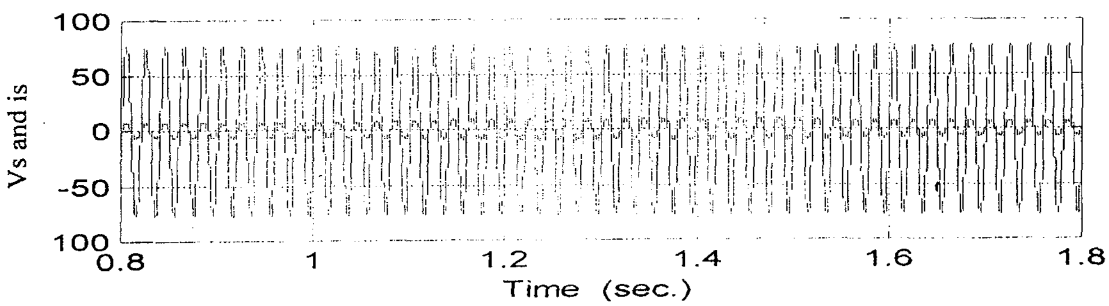
(a)



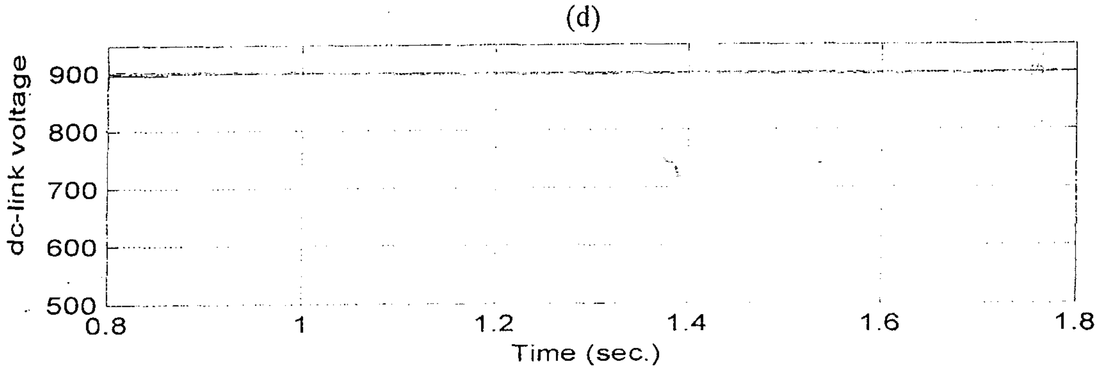
(b)



(c)



(d)



(e)

Fig.6. Simulation drive response due to load changes: (a) Speed response; (b) torque response; (c) phase current (d) dc-link voltage (e) supply current and supply voltage

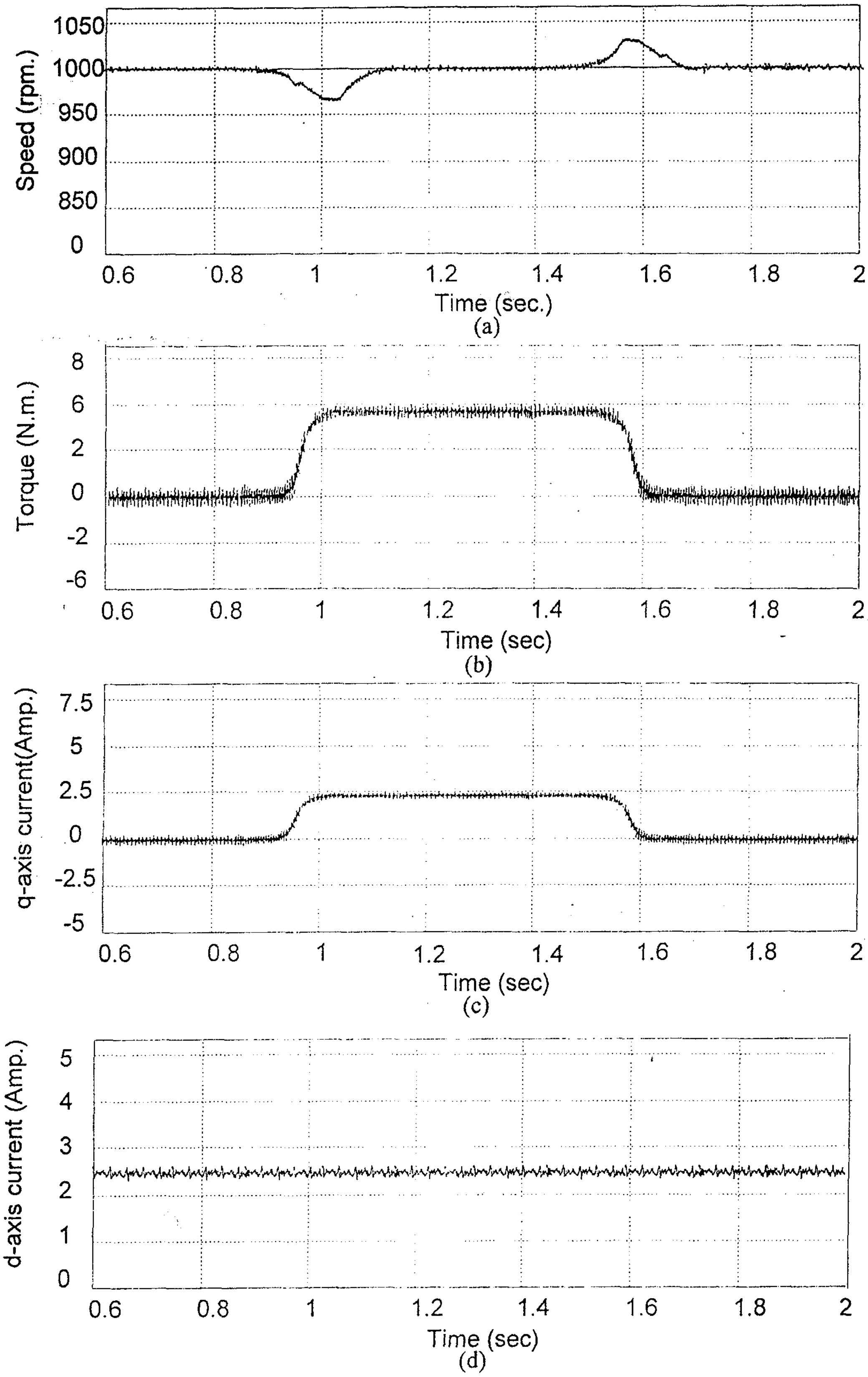
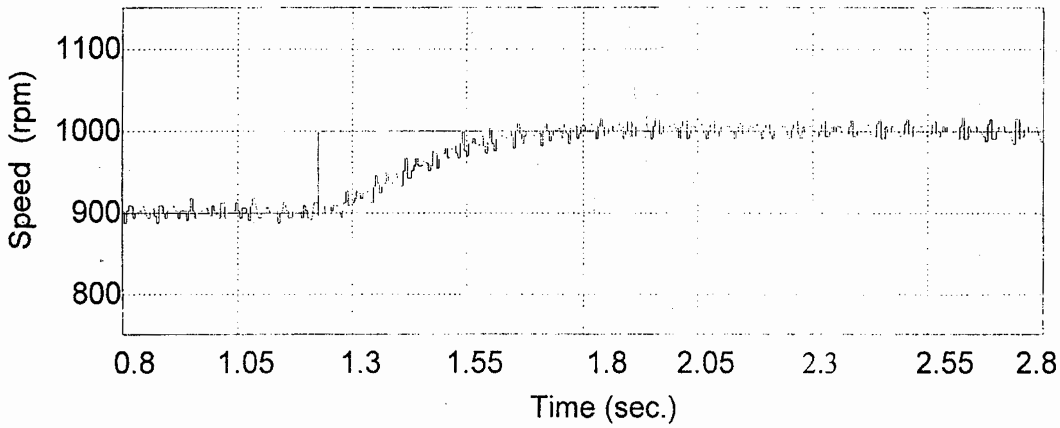
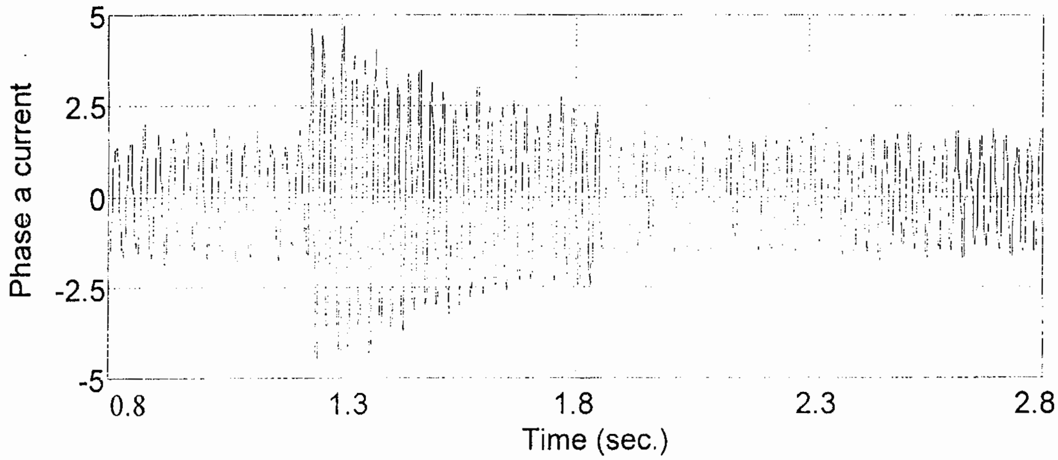
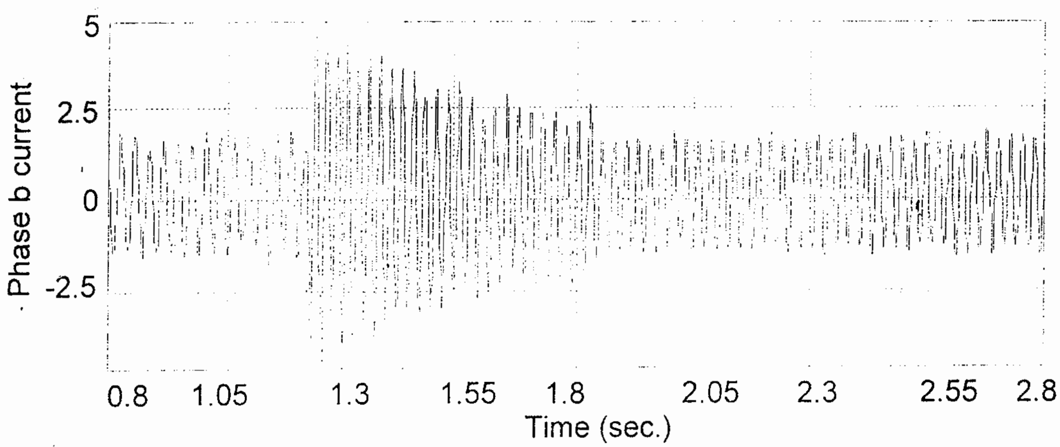


Fig. 7 Experimental drive response due to load changes (a) motor speed (b) motor torque (c) quadrature axis current component (d) direct axis current component







6. CONCLUSIONS

This paper has presented a single-phase to three-phase converter. This converter provides variable output voltage and frequency. The proposed converter has maintained sinusoidal input current with unity input power factor.

Bidirectional power flow and regenerative breaking have been achieved. Based on vector control technique, a control strategy of the proposed converter fed induction motor has been implemented .The low cost ac drive has been ensured using the proposed topology. A laboratory drive system has been built and tested to verify the most important features of the proposed control system. The results have been shown the efficacy of the low cost ac drive system.

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LIST OF SYMBOLES:

K_c	: Controller gain constant of front end rectifier
NS	: logic variable of rectifier
V^e_{qs}, V^e_{ds}	: q^e-d^e –axis stator voltage
I^e_{qs} , I^e_{ds}	: q^e-d^e –axis stator current
λ^e_{qs} , λ^e_{ds}	: q^e-d^e –axis stator flux linkage
R_s , R_r	: Stator and rotor resistances
J, B	: Moment of inertia and viscous coefficient of the motor
L_s, L_r, L_m	:Stator, rotor and mutual inductances
ω_e , ω_r	:Stator angular frequency ,and rotor speed
p	:Differential operator ($\frac{d}{dt}$)
K_{ps}	: Proportional gain
τ_{cs}	:Time constant of the PI controller .
τ_r^*	:Rotor time constant (L_r/R_r).,The synchronous reference frame
NA, NA1	:Phase 'a ' pulses logic
NB, NB1	:Phase 'b' pulses logic
V_a,V_b,V_c	:Motor phase voltage
C_o	:Output capacitance
E_{dc}	:Total dc busbar voltage
E_{dc-}	:Negative half DC busbar voltage
E_{dc+}	:Positive half DC busbar voltage
E_{dcr}	:Reference DC busbar voltage
i_s	:Supply current
i_r	:Supply reference current
L_s	:Supply inductance
v_s	:Phase to neutral supply voltage
T_e, T_L	:Electromechanical torque and mechanical load torque

APPINDEX

The induction motor is a three-phase squirrel cage and has the following parameters;

Rated power	: 1.1 kw
Rated load torque	: 7.5 N.m.
No. of pole pair	: 2
Stator resistance	: 7.4826 ohm
Rotor resistance	: 3.6840 ohm
Rotor leakage inductance	: 0.0221 H
Stator leakage inductance	: 0.0221 H
Mutual inductance	: 0.4114 H
Supply frequency	: 50 Hz
Motor speed	: 1400 r.p.m.
Supply voltage	: 220 volts

مغير الوجه الواحد إلى ثلاثة اوجه لتخفيض تكلفة تسيير محركات التيار المتردد

د/توفيق سعد رضوان*،م/زكريا محمد سالم البربري**،أ.د/صبري عبد اللطيف محمود*

* كلية الهندسة بشبين الكوم–جامعة المنوفية

** المصانع الدولية لدرفلة الصلب–مدينة السادات

ملخص البحث

ظهرت في ألا ونه الأخيرة بعض مغيرات التيار للتحويل من منبع قدرة أحادي الوجه إلى منبع قدرة ثلاثي الأوجه وذلك للاستخدام مع محركات التيار المتردد ذات السر عات المتغيرة و القليلة التكاليف عالية الكفاءة . و تتميز هذه المتغيرات باستخدام عدد قليل من المفاتيح الإلكترونية .

يقترح هذا البحث دراسة أداء مغير تيار لتحويل القدرة الكهربية من منبع أحادي الوجه إلى منبع قدرة ثلاثي الأوجه متغير التردد و الجهد. و الدائرة المقترحة تحتوي فقط على ستة مفاتيح إلكترونية سريعة التوصيل التي تشكل دائرة الموحد ودائرة العاكس حيث يتم تحويل الجهد أحادي الوجه إلى ثلاثة اوجه في خطوة واحدة . وتتكون دائرة الموحد من مفتاحين إلكترونيين سريعي التوصيل يوصلان مع منبع الجهد أحادي الوجه من خلال ملف تعضيد الجهد كذلك يوصل خرج الموحد على مكثفين لتثبيت الجهد الخارج وتوصيل نقطة الانتصاف للمكثفين. بالطرف الراجع لمنبع الجهد. إما العاكس فيتكون من أربعة مفاتيح سريعة التوصيل تشكل وجهين من خرج العاكس إما الوجه الثالث فيوصل بنقطة الانتصاف للمكثفين.

وتم التحكم في الموحد بحيث يسمح بمرور القدرة الكهربية من المصدر إلى المكثفين و العكس , كذلك أظهرت طريقة التحكم تثبيت الجهد الخارج على المكثفين ضد أي تغيرات في منبع الجهد , وكذلك تم الحصول على شكل جيبي لموجة تيار المنبع و معامل قدرة يساوي الوحدة. ثم تطبيق المغير المقترح على محرك تأثيري ثلاثي الأوجه و تم استخدام طريقة التحكم المتجه للتحكم في سرعة المحرك .

تم بناء نموذج معملي للنظام المقترح و قد أظهرت النتائج العملية تطابقا مع النتائج النظرية مما يشجع على استخدام المغير المقترح في التطبيقات التي تحتاج تكلفة اقل مع أداء عالي. و يفيد هذا المغير في التطبيقات المنزلية مثل أجهزة التكييف و الثلاجات و الغسالات الكهربائية وكذلك فى التطبيقات التي تستخدم فى المناطق الزراعية .