Analysis and Control of PWM Buck-Boost AC Chopper Fed Single-Phase Capacitor Run Induction Motor

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ABSTRACT

Single phase capacitor-run induction motors (IMs) are used in various applications such as home appliances and machine tools; they are affected by the sags or swells and any fault that can lead to disturb the supply and make it produce rms voltage below or above the rated motor voltage, which is 220V. A control system is designed to regulate the output voltage of the converter irrespective to the variation of the load and within a specific range of supply voltage variation. The steady-state equivalent circuit of the Buck-Boost chopper type AC voltage regulator, as well as the analysis of this circuit are presented in this paper. Switching device for the regulator is an IGBT Module. The proposed chopper uses pulse width modulation (PWM) control technique to chop the input voltage into segments in order to guarantee rated rms voltage supplied to the load, which is capacitor-run induction motor. Proportional integral (PI) controller is used to obtain very small steady state error, stable and fast dynamic response, and robustness against variations in the line voltage. The complete system is simulated using software package, and the results are obtained to verify the proposed control method.

Keywords:- AC chopper, voltage controller, pulse width modulation, duty cycle, damped input filter.

تحليل وسيطرة لمنظم الفولتية المتناوبة الخافض- الرافع لسواقة محرك حثي احادي الطور ذو متسعة دائمة العمل

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الخلاصة

ان المحركات الحثية الاحاديه الطور ذات المتسعة الدائمة العمل، شائعة الاستخدام في التطبيقات المختلفه مثل بعض الاجهزة الكهربائية المستخدمه داخل البيت والمكائن، والتي تتاثر بالانخفاض اوالارتفاع بالفولتية او اي خطأ يحصل بالنظام والذي ممكن ان يؤدي الى ارباك مصدر الفولتيه ويجعله يولد متوسط قيمة فعالة مقدارها ادنى او اعلى من القيمة المطلوبة لتشغيل المحرك بصورة طبيعية، والتي مقدارها (220فولت). نظام السيطرة المقدم في هذا البحث تم تصميمة لتعديل الفولتية المسلطة على الحمل ضمن مدى معين من التغير في فولتية المصدر برغم التغير بالحمل.الدائرة المكافئة للمنظم الرافع-الخافض في الحالة المستقرة وكذلك طريقة التحليل لدائرة الخافض-الرافع للفولتية المتناوبه تم عرضها في هذا البحث. المفتاح المستخدم بجهاز السيطرة يمثل بالحقيقة مفتاح قدرة الكتروني شبة موصل نوع (IGBT Module). منظم الفولتيه المقترح يستخدم طريقة التحكم بعرض الموجة (PWM) للسيطرة على الفولتية الداخلة واعادة توليدها وفق مديات محددة لضمان القيمة المطلوبة للفولتية المسلطة على الحمل المتمثل بالمحرك الاحادي الطور ذو المتسعة دائمة العمل. في هذا البحث تم استخدام المسيطر نوع تناسبي تكاملي (PI) حيث ان هذا المسيطر يوفر الحصول على نسبة خطأ قليلة للتخلص من الفرق بين فولتية المصدرالمعرضة للخطأ والفولتية المطلوب الثبات عند قيمتها بسرعة استجابة عالية ويجعل سير العمل مستقر ومتين ضد اي اختلال يحصل بفولتية المصدر. تمت محاكاة النظام المصمم ككل نظريا باستخدام (Matlab/Simulink)، وتم الحصول على النتائج لتاييد صلاحية النظام المقدم.

كلمات رئيسية:- مقطع الجهد المتناوب، منظم فولتية، التحكم بعرض الموجة، الزمن الدوري للمقطع، مرشح الدخول المثبَط.

1. INTRODUCTION

Single-phase capacitor-run (IMs) are widely used because they have good power factor and efficiency under load. It is required to develop the methods of controlling its operation to reach the best performance. Several methods exist for variable speed operation of a single-phase(IMs). Considering simplicity and low cost, most common type is the line-frequency AC choppers, which can be found as a conventional phase-controlled AC controllers using thyristors, which have the advantages of simplicity of the control circuit and large power capability. However, these have the inherent drawbacks that power factor decreases when the firing angle increases and that, since the content of the line current harmonics is relatively large, the size of the passive filter circuit becomes bulky. There are also other methods which use a tapped winding transformer to regulate the input voltage to a lower or higher output voltage. However, because the winding ratio is changed by servo motor or by manual regulation, it has low regulation speed, **Kwon, et al., 1996** and **Nan, et al., 2009**.

The PWM Buck-Boost AC chopper can overcome all these drawbacks and guarantee the best control for this kind of motors since it offers several advantages such as sinusoidal input current, fast dynamics, and significant reduction in filter size. In addition the problems caused by the sags or swells of the input voltage can be solved by proposing voltage controller which uses output peak voltage as feedback signal and adopts proportional integral (PI) control strategy to regulate the output voltage, **Kown, et al., 1999**.

The object of this paper is to present the analyses and design of PWM AC chopper circuit suitable to drive a single phase capacitor run induction motor.

2. CIRCUIT CONFIGURATION AND PRINCIPLE OF OPERATION

The basic circuit configuration of the Buck-Boost AC converter is shown in **Fig.1**. It can operate directly from the single-phase line (source) voltage *vs* and regulate the output voltage higher or lower steplessly, by using two bidirectional standard switches modules capable of bidirectional current control and regenerative DC snubbers consisting of capacitor only C_b, to absorb bidirectional turn-off spike energy due to line stray inductance. The input filter consisting of inductor L_i and capacitor C_i , absorbs the harmonic currents. The used bi-directional switch module is composed of two insulated gate bipolar transistors (IGBT). The switches S_1, S_2, S_3 and S_4 are unidirectional. The inductor L is used to store the input energy and transfer it to the output side, the filter capacitor C_o at the output side reduces the output voltage ripple, **Kown, et al., 1999**.

A switching policy solving the commutation problem is based on the polarity of the switch-to-switch voltage v_t across two bidirectional switches: two unidirectional switches S_3 and S_4 are additionally turned on during the positive period of v_t and the switches S_1 and S_2 during the negative period of v_t, to avoid voltage spikes, without affecting the value of the duty ratio (D). Then the inductor current is bypassed through the input side or output side, depending on its direction during the dead-time. The control of the switches is based on the symmetrical PWM techniques. This ensures that the output voltage is sinusoidal for a sinusoidal AC input voltage. The output voltage is controlled by changing the duty cycle of the control pulses. Three modes of operation are possible during one switching cycle for v_t>0 :

1. Charging mode; the inductor current flows through the input side via S_1 and the diode across S_3 for i_L>0, or S_3 and diode across S_1 for i_L <0.
2. Discharging mode; the inductor current flows through the output side via S_4 and the diode across S_2 for i_L>0 , or S_2 and the diode across S_4 for i_L <0.
3. Dead-time mode; the inductor current is bypassed through the input or output side depending on its direction.

Fig. 2a-c show the inductor current and voltage waveforms for $v_t > 0$ during one switching cycle. The inductor voltage v_L is v_i during the charging mode and v_o during the discharging mode as shown in **Fig. 2b** and **c.** Because the switches S_3 and S_4 are turned on for $v_t > 0$, the inductor voltage v_L is v_o or v_i, according to the direction of i_L during the dead-time mode, **Kim, et al., 2011**.

3. ANALYSIS

To facilitate the analytical procedure in order to obtain an equivalent circuit for the buck-boost AC chopper, all components are assumed ideal and the switching frequency f_s is much greater than the line frequency f, so that during a switching period, the input and output voltage can be cosidered constant. The average inductor voltage during one switching period $T_s = 1/f_s$ is given by, **Kown, et al., 1999**:

$$v_L(t) = Dv_i(t) - (1-D)v_o(t) \tag{1}$$

Where $v_i(t)$ and $v_o(t)$ are the average AC input voltage and output voltage, respectively, during the switching period, and D is the duty ratio. The inductor voltage is given by:

$$v_L(t) = L\frac{di_L(t)}{dt} \tag{2}$$

Where $i_L(t)$ is the average inductor current during the switching period. If charging the inductor by the input current will take a time equals to (T_{on}), then discharging the inductor current to the load can be represented by the following equation:

$$i_o(t) = (1-D)i_L(t) \tag{3}$$

Where:

$$i_i(t) = Di_L(t) \tag{4}$$

Where $i_o(t)$ and $i_i(t)$ are the average output and input current respectively. From Eqs. (1) and (2), the following equation is obtained:

$$Dv_i(t) = L\frac{di_L(t)}{dt} + (1-D)v_o(t) \tag{5}$$

Substituting Eq. (3) in Eq. (5), yields:

$$\frac{D}{1-D}v_i(t) = \frac{L}{(1-D)^2}\frac{di_o(t)}{dt} + v_o(t) \tag{6}$$

Eq. (6) represents the steady-state equivalent circuit for the chopper type voltage regulator. The equivalent circuit is shown in **Fig.3**. At steady-state motor impedance including (R_o and X_o)

varies according to the load applied to the motor which is 175W single phase capacitor-run induction motor. From the equivalent circuit shown in **Fig.3**, the transfer function of the output voltage $V_o(s)$ with respect to the input voltage $V_i(s)$ is obtained as, appendix A, **Kamel, 2013**:

$$\frac{v_o(s)}{v_i(s)}=\frac{D(1-D)(R_o+sL_o)}{s^3LL_oC_o+s^2LC_oR_o+s[L+(1-D)^2L_o]+(1-D)^2R_o} \tag{7}$$

From Eq. (3) and Eq. (7), Eq. (8) is obtained

$$\frac{I_L(s)}{V_i(s)}=\frac{D[s^2L_oC_o+sC_oR_o+1]}{s^3LL_oC_o+s^2LC_oR_o+s[L+(1-D)^2L_o]+(1-D)^2R_o} \tag{8}$$

The source current $I_s(s)$ and source voltage $V_s(s)$ are obtained as

$$I_s(s)=DI_L(s)+I_{ci}(s)$$

$$=V_i(s)\cdot\left[\frac{D^2[s^2L_oC_o+sC_oR_o+1]}{s^3LL_oC_o+s^2LC_oR_o+s[L+(1-D)^2L_o]+(1-D)^2R_o}+sC_i\right] \tag{9}$$

$$\mathrm{v}_s(s)=sL_i\mathrm{I}_\mathrm{s}(\mathrm{s})+\mathrm{v}_\mathrm{i}(\mathrm{s})$$

$$=V_i(s)\cdot\left[\frac{sL_iD^2[s^2L_oC_o+sC_oR_o+1]}{s^3LL_oC_o+s^2LC_oR_o+s[L+(1-D)^2L_o]+(1-D)^2R_o}+s^2L_iC_i+1\right]$$
(10)

From Eq. (9) and Eq. (10), the following relation is obtained for simplicity:

$$\frac{Vs(s)}{Is(s)}=\frac{a_5s^5+a_4s^4+a_3s^3+a_2s^2+a_1s+a_0}{b_4s^4+b_3s^3+b_2s^2+b_1s+b_0} \tag{11}$$

Where: $a_5=C_iL_iLL_oC_o$, $a_4=LL_iC_oC_iR_o$, $a_3=[D_2L_iL_oC_o+ L_iC_i[L+(1\text{-}D)^2L_o+LL_oC_o]$, $a_2=[D^2L_iC_oR_o+C_iL_i(1\text{-}D)^2 R_o+LC_oR_o]$, $a_1=[L_iD^2+L+(1\text{-}D)^2L_o]$, $a_o=(1\text{-}D)^2R_o$, $b_4= C_iC_oLL_o$, $b_3=LC_oC_iR_o$, $b_2=[D^2L_oC_o+C_i[L+(1\text{-}D)^2L_o]]$, $b_1=[D^2C_oR_o+(1\text{-}D)^2C_iR_o]$, $b_o= D^2$

Under an appropriate selection of the parameters L, L_i, C_o and C_i satisfying $a_5\approx a_4\approx a_3\approx a_2\approx b_4\approx b_3\approx b_2\approx b_1\approx 0$, Eq. (10) can be approximated as follows:

$$\frac{\mathrm{Vs(s)}}{\mathrm{Is(s)}}\approx\frac{s[L_iD^2+L+(1-D)^2L_o]+(1-D)^2R_o}{D^2}$$

$$\frac{\mathrm{V_S(s)}}{\mathrm{I_S(s)}}\approx\frac{\mathrm{s[L_iD^2+L+(1-D)^2L_o]+(1-D)^2R_o}}{\mathrm{D^2}} \tag{12}$$

From Eq. (12), the angle of the power factor θ is given by

$$\theta \approx \tan^{-1}\frac{\omega[L_i D^2 + L + (1-D)^2 L_o]}{(1-D)^2 R_o}$$

$$\theta = \tan^{-1}\frac{\omega[L_i D^2 + L + (1-D)^2 L_o]}{(1-D)^2 R_o} \quad (13)$$

Where ω (2πf) is the angular frequency of the source voltage.
Derivation of equations: (7, 8, 9, 10 and 12) are shown in appendix A, **Kamel, 2013**.

4. DESIGN OF INPUT AND OUTPUT FILTERS AND VOLTAGE CONTROLLER

For fast dynamic response the voltage controller, which uses the output peak-voltage as the feedback signal, is designed to keep the stability of the output voltage in case of input voltage fluctuation. The peak-voltage detector system is shown in **Fig.4**. Where v_o is the rms voltage across the load terminals, v_{od} is the sensed output of the proposed detector, k_d is the detection gain and V_o is the peak value of the output voltage. A fast detection technique is composed of a phase shifter, two multipliers, and an adder. The detection technique utilizes the simple trigonometric principle as follows, **Kown, et al., 1999** :

$$\sin^2\omega t + \cos^2\omega t = 1 \quad (14)$$

The sensed output v_{od} is:

$$v_{od} \propto (V^2{}_o \sin^2\omega t + V^2{}_o \cos^2\omega t) = k^2{}_d\ V^2{}_o \quad (15)$$

When the disturbed input v_i , is applied to the system, the sensed output v_{od} (disturbed by the disturbance input) needs to be regulated to a desired constant reference V_r with no steady-state error where V_r is $k_d^2 V_{or}^2$ and V_{or} is the peak value of the desired output voltage. A good response is obtained by using a traditional PI controller. The integral part of the designed controller makes the steady-state output voltage error zero. The controller output is duty ratio D, **Nan, et al., 2009**:

$$D = k_p \Delta v + ki \int \Delta v dt \quad (16)$$

Where k_p and k_i are proper proportional and integral gains respectively. Δv is the controller input:

$$\Delta v = V_r - v_{od} \quad (17)$$

A filter must be added at the power input of a switching converter for improving power quality and interface issues. Low pass input LC filter needs to be damped at the corner frequency fc to prevent the gain of the filter to go to infinity otherwise, this rise would cause extreme current peaks which would make the system worse than if it was without filter. Taking into account only the dominant harmonic, the input current of the converter is represented with good approximation by, **Barbi, et al., 1991**:

$$i_i = I_{ip} \sin \omega t + I_{ip} \sin \omega t \ \sin \omega_s t \quad (18)$$

Where I_{ip} is the peak value of the input current, thus,

$$i_h = I_{ip} \sin \omega t \ \sin \omega_s t \tag{19}$$

The role of the input filter is to prevent the harmonic current i_h from circulating through inductor L_i . According to the harmonic equivalent circuit shown in **Fig.5**:

$$\frac{i_{Lh}}{i_h} = \frac{1}{\omega_s{}^2 L_i C_i - 1} \tag{20}$$

According to the usual specification, the total harmonic distortion (THD) of the input current i_i is ≤ 5%. Then reducing the value of i_{Lhp} (which is the maximum harmonic current circulating through inductor L_i) to 3% of input current, the THD ≤ 5% is ensured, **Barbi, et al., 1991**

$$\frac{i_{Lhp}}{I_{ip}} = 0.03 \ => i_{Lhp} = 0.03 \ I_{ip}, \text{ where } I_{hp} = \ I_{ip}$$

then using Eq. (20):

$$\frac{1}{\omega_s{}^2 L_i C_i - 1} = 0.03 \tag{21}$$

$\omega_s{}^2 = (2\pi f_s)^2 = 2.47 * 10\text{^}10$ rad/s, where

f_s=25KHZ, Let L_i=0.5mH , then

$C_i = \frac{32}{L_i \omega_s{}^2} = 2.59\mu F$. A value of 3μF is chosen.

A damped filter made with a resistor R_d in series with a capacitor C_d as shown in **Fig.6**, all connected in parallel with the filter's capacitor C_i. The purpose of resistor R_d is to reduce the output peak impedance of the filter at the cutoff frequency. The choice of C_d ,that leads to the minimum peak output impedance, for a given value of R_d can be expressed as following:

For C_d =20μF, then

$n = \frac{C_d}{C_i}$ => n=6.7 , the optimum damping resistance value R_d is equal to, **Erickson, 1999**:

$$R_{d_{opt}} = \sqrt{\frac{L_i}{C_i}} \cdot \sqrt{\frac{(2+n)(4+3n)}{2n^2(4+n)}} \approx 6\Omega \tag{22}$$

The filter high-frequency attenuation is not affected by the choice of C_d and the high-frequency asymptote is identical to that of the original undamped filter, **Erickson, 1999**. The energy storage inductor L is used to store the energy from the source during the active mode of the chopper operation, then transfer it to the load during the discharging mode. During one switching period T_s, inductor must be satisfied such that, **Li, et al., 2011**:

$$L \geq \frac{V_o (1 - D_{max})^2 \eta}{2 I_o k_c f_s} \tag{23}$$

With maximum value of duty ratio evaluated from, **Li, et al., 2011**:

$$\mathrm{D_{max}} = \frac{1}{1+\left(\frac{\mathrm{V_{imin}}}{\mathrm{V_o}}\right)} \quad (24)$$

$$= 0.578$$

And for typical values of output current = 1.4A, **Kamel, 2013**, η (convertor efficiency)=90%, k_c (ripple coefficient of inductor current) ≤0.5, **Li, et al., 2011**. The boundary L_{min} =1mH, for $L \geq L_{min}$., the convertor operates in continuos conduction mode(CCM). A value of 1mH is chosen. Output filter capacitor is used to reduce the output voltage ripple and harmonics. During one T_s , to limit the output voltage ripple ,output filter capacitor C_o must satisfy the following, **Li, et al., 2011**:

$$\mathrm{C}_o \geq \frac{\mathrm{I_o D}_{max}}{k_v V_o f_s}$$

With k_v (ripple coefficient of output voltage) ≤0.1, **Li, et al., 2011,** the boundary C_{omin} equals to 1.43µF, for C_o more than or equal to C_{omin} ($C_o \geq C_{omin}$); A value of 10 µF is chosen in the simulation to ensure output voltage without harmonics.

5. SIMULATION RESULTS

To show the feasibility of the proposed analysis method and control strategy, the simulation model of the proposed voltage regulator is implemented using Matlab/Simulink software. The induction motor whose its specifications shown in **Table 1** and its measured parameters listed in **Table 2** are used in the simulation, **Kamel, 2013**. THD values for current input to the chopper and motor input current and voltage are listed in **Table 3** (the values are obtained by using Matlab/Simulink).

5.1 Results for Sudden Change of Supply Voltage from (220V rms) Value to (160V rms) Value

At normal operation the supply produces rms voltage equals to the rated voltage (220V); if suddenly a disturbance occurred in the system leads to drop the voltage to (160V) rms value, then the voltage controller will detect it and the AC chopper will regulate the dropped voltage to be the same as the rated voltage needed by the motor.

5.1.1 **No-load condition**

Fig. 7 shows the simulated waveform of the supply voltage and load voltage at no-load and at source main frequency of 50HZ. In this figure it is shown that after (0.5sec), the load voltage dropped then increased gradually within (60msec) to reach its rated (rms value =220V) with low THD equals to (0.35%) even if the supply continues on the low level (160V). **Fig. 8** shows the simulated waveform of the supply current which has rms value equals to (1.76A), with (THD = 4.09%). **Fig. 9** shows the simulated waveform of the motor input current, the disturbance occured at (0.5sec), current decreased temporarily (since voltage across load terminals dropped temporarily) then gradually increased to reach (rms value = 0.56A). Motor speed is shown in **Fig. 10**, at no-load motor runs at (1500rpm), after the disturbance occurred, the motor speed dropped temporarily corresponding to the drop in load voltage, then the motor returns to run at no-load speed.

5.1.2 **Full-load condition**

The simulated waveforms of the supply voltage and load voltage at full-load are the same as those shown in **Fig.7**, since same drop in supply voltage occurred at (0.5sec). The load terminals still receive rated (220V) rms value, with low THD equals to (0.51%) even if the supply continues on the low level (160V). **Fig. 11** shows the simulated waveform of the supply current which has (rms value =2.40A) at steady state operation; then if the disturbane occurred at (0.5sec) and input voltage dropped to (160V) rms value, then current drawn from the supply equals to (2.68A) rms value, with (THD =4.47%). **Fig. 12** shows the simulated waveform of motor input current, the disturbance occured at (0.5sec), current decreased temporarily (since voltage across load terminals dropped temporarily), then increased gradually to reach rms value close to (1.40A). Motor speed is shown in **Fig. 13**, full-load motor runs at (1240rpm) after the disturbance occurred at (0.5sec) motor speed dropes temprorily then increased (corresponding to the load voltage variation) until motor returns to run at speed of (1240rpm).

5.2 Results for Sudden Increase of Supply Voltage from (220V rms) Value to (280V rms) Value

At normal operation the supply produces rms voltage equals to the rated voltage (220V); if suddenly a disturbance occurred in the system leads to increase the voltage to (280V) rms value, then the voltage controller will detect it and the AC chopper will regulate the voltage to be same the rated voltage needed by the motor.

5.2**.1 No-load condition**

Fig.14 shows the simulated waveform of the supply voltage and load voltage at no-load and at source main frequency of 50HZ. In this figure it is shown that after (0.5sec), the load voltage increased then dropped gradually within (40msec) to reach its rated rms value (220V) with THD (measured from Matlab/Simulink) equals to (0.40%) even if the supply continues on the high level (280V). **Fig.15** shows the simulated waveform of the supply current which has (rms value = 1.76A); at (0.5sec) the supply is disturbed and input voltage is increased to (280V) rms value, then current drawn from the supply equals to (2.12A) rms value. **Fig.16** shows the simulated waveform of the motor input current, the disturbance occured at (0.5sec), current increased temprorily then returned to the rms value (0.56A). Motor speed is shown in **Fig.17**, at no-load motor runs at (1500rpm), after disturbance occurred at (0.5sec) motor speed exceeds (1528rpm) temprorily, corresponding to the increase in load voltage.

5.2.2 **Full-load condition**

The simulated waveforms of supply voltage and load voltage at full-load are the same as those shown in **Fig.14** and at source main frequency of 50HZ. In this figure it is shown that after (0.5sec), the load voltage increased then decreased gradually to reach its rated rms value (220V) with low THD equals to (0.70%) even if the supply continues on the high level (280V). **Fig.18** shows the simulated waveform of the supply current which has rms value equals to (2.40A); at (0.5sec) the supply is disturbed and input voltage is increased to (280V) rms value, then current drawn from the supply equals to (2.68A) rms value, with (THD =4.24%). **Fig.19** shows the simulated waveform of the motor input current, at full-load motor draws rated (rms value =1.41A), when the disturbance occured at (0.5sec), current increased temprorily then returned to its rated value. Motor speed is shown in **Fig.20**, at full-load motor runs at (1240rpm), after disturbance occurred at (0.5sec) motor speed increased to (1320 rpm) temprorily, corresponding to the increase in load voltage, then the motor returns to run at its full-load speed again. Total harmonic distortion (THD) for input current, load current and voltage across the load terminals after disturbing input voltage at (0.5sec) are shown in **Table 3** (The values of (THD) are obtained by using Matlab/Simulink software); it is seen that the design of input filter ensures that (THDi < 5%), also load current and load voltage has a low value of (THD) since the output filter is well designed. **Table 4** illustrates how the input power (W) that is drawn from the source, varies with the line power factor which is measured by using Matlab/Simulink, depending on the equation: $p.f = P/(P^2+Q^2)^{1/2}$, where P and Q are active and reactive input powers respectively and $p.f$ is the input power factor.

6. CONCLUSION

The analysis of Buck-Boost AC chopper circuit, that can regulate output voltage higher or lower steplessly is presented. The input current is sinusoidal waveform with low harmonic components. The output voltage control system is designed using PI control method and the peak-voltage detector. The simulation results show that the voltage controller has a good dynamic performance when input voltage swells or sags occur, since the output voltage returns to its normal value of (220V) rms value after no more than three cycles (60msec). The results show low total harmonic distortion factor for input current, load current and load voltage. The AC chopper achieves an acceptable line power factor at full load with low input voltage, since *(D*) will be increased, according to Eq. (13), the motor needs a power (175W) to run at full load, but the power drawn from the source is higher, the difference represents the losses since in the construction of the motor the resistances of its windings are very high, appendix B, **Kamel, 2013**.

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NOMENCLATURE

C_b = snubber capacitor, μF.
C_d = damping capacitor, μF.
C_i = input capacitor, μF.
C_o = output capacitor, μF.
D = duty ratio.
f = main frequency, Hz.
f_c = corner frequency, Hz.
f_s = switching frequency, Hz.
i_1 = output capacitor current, A.
i_2 = load current, A.
i_{Ci} = input capacitor current, A.
i_h = harmonic current, A.
I_{hp}= peak harmonic current, A.
$i_i(t)$ = average input current during the switching period, A.
I_{ip} = peak converter input current, A.
$i_L(t)$= average inductor current during the switching period, A.
I_{Lhp}= peak harmonic current in input inductance, A.
$i_o(t)$ = average output current during the switching period, A.
i_s = source current, A.
k_c = ripple coefficient of inductor current.
k_d = detection gain of the output voltage.
k_p , k_i = proportional and integral gains respectively.
k_v = ripple coefficient of output voltage.
L = energy storage inductor, mH.
L_i = input inductor, mH.
R_d= damping resistance, Ω.
R_o= resistance of the load, Ω.
S_1= first unidirectional switch.
S_2= second unidirectional switch.
S_3= third unidirectional switch.
S_4= forth unidirectional switch.
T_s = one switching period, s.
$v_i(t)$= average input voltage, V.
$v_L(t)$= average inductor voltage, V.
$v_o(t)$= average output voltage, V.
v_{od} = detected output peak-voltag, V.
v_r = reference voltage, V.
v_s = source voltage, V.
v_t = switch-to-switch voltage across two bidirectional switches, V.
X_o =inductance of the load, Ω.
Z_o= output impedance, Ω.
η= conversion efficiency.
θ= input phase angle.

θ_o= output phase angle.
ξ= damping factor.
ω= angular frequency of the source voltage, rad/s.
ω_s= angular switching frequency, rad/s.

9 FIGURES AND TABLES

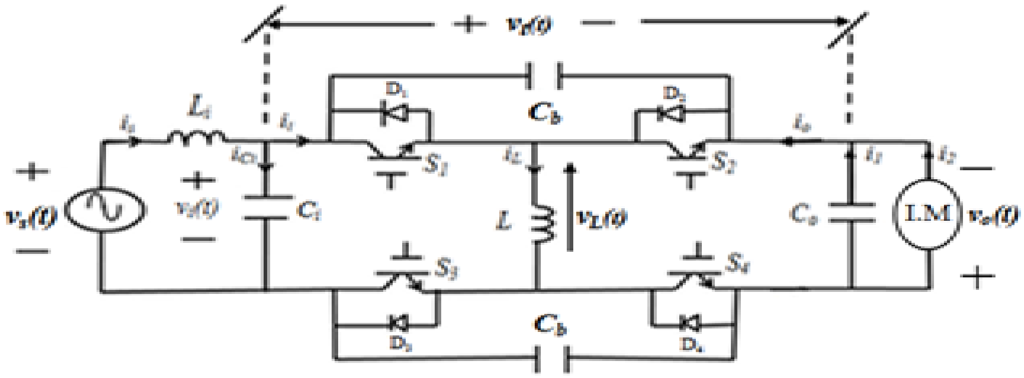

Figure1. Buck-boost AC chopper circuit configuration.

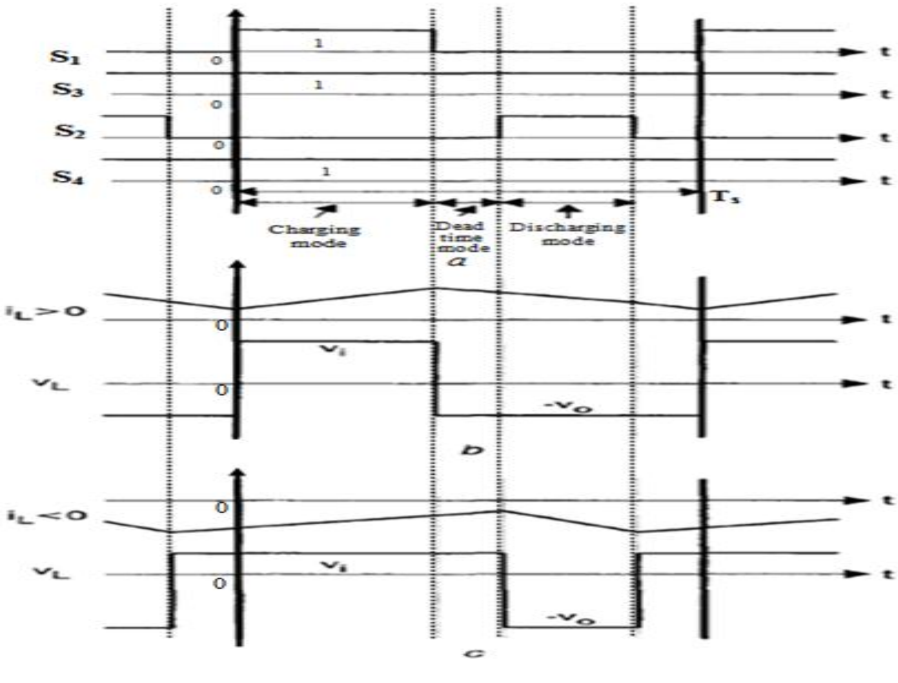

Figure 2. a.Gate signals, b. Inductor current and voltage waveforms for v_t>0, i_L>0 during one switching cycle,c. i_L and v_L for i_L<0.

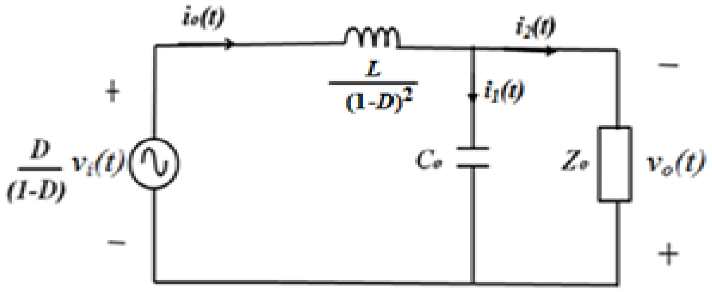

Figure 3. Steady state equivalent circuit of buck-boost AC chopper.

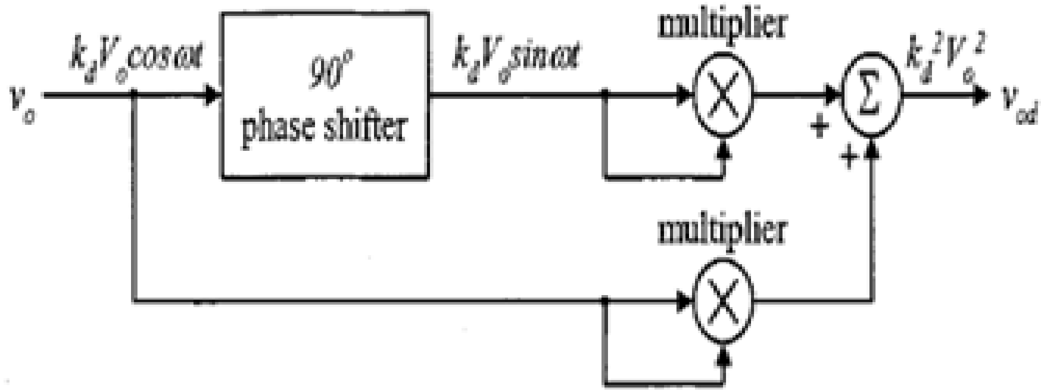

Figure 4. Fast peak voltage detector.

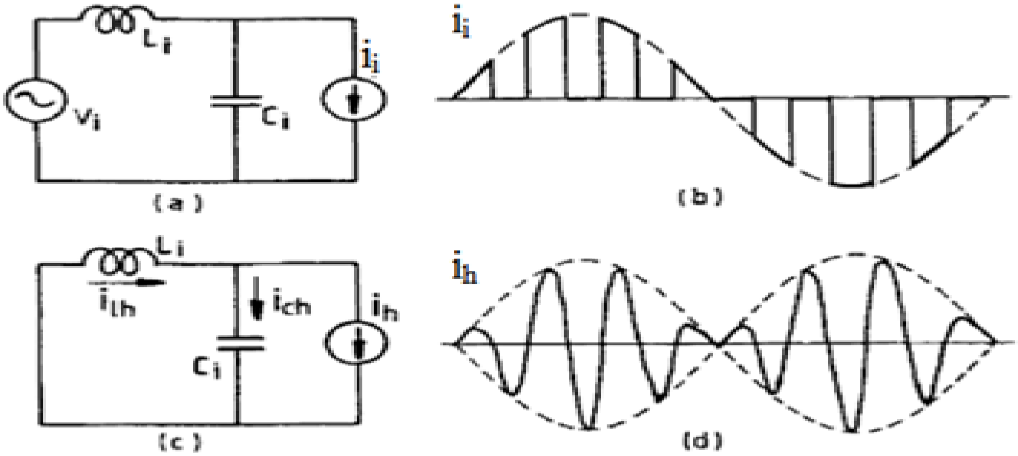

Figure 5. (a) Input filter:(b) input converter current; (c) harmonic equivalent circuit; (d) dominant harmonic current i_h

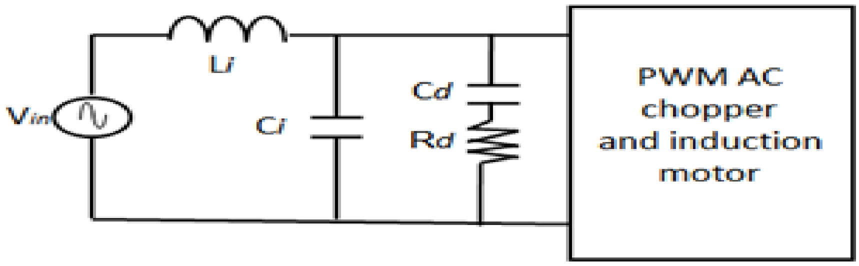

Figure 6. Structure of undamped and damped LC filter.

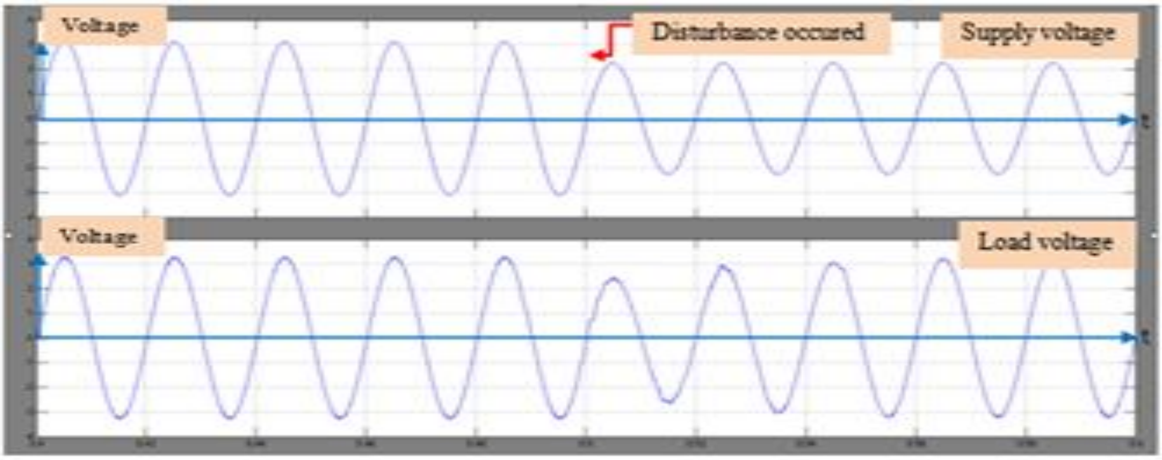

Figure 7. Steady state of supply voltage and load voltage waveforms at source main frequency HZ, (100V/div), time(20ms/div).

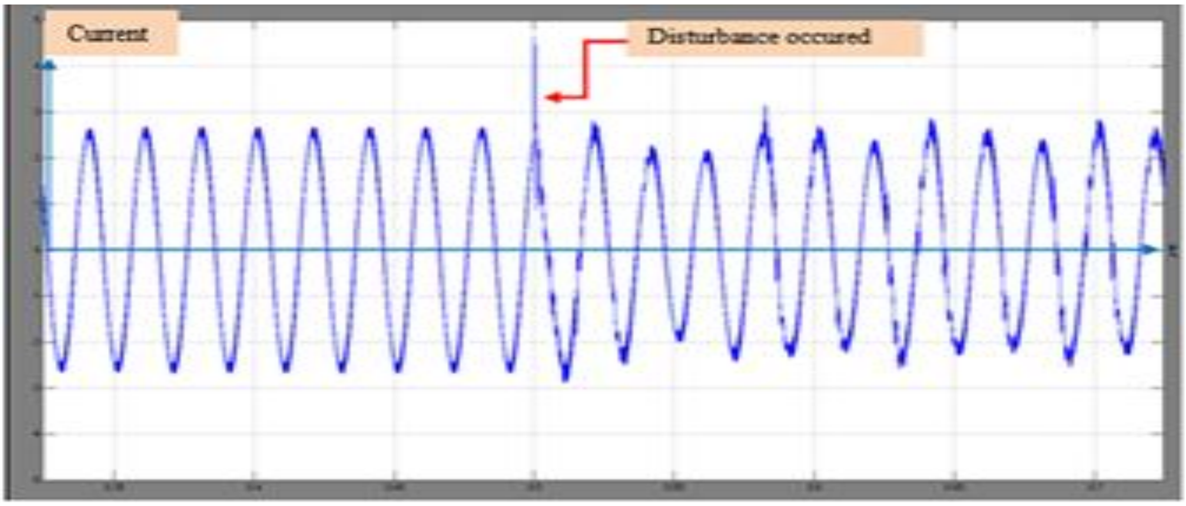

Figure 8. Steady state of supply current waveform, (1A/div), time(50ms/div).

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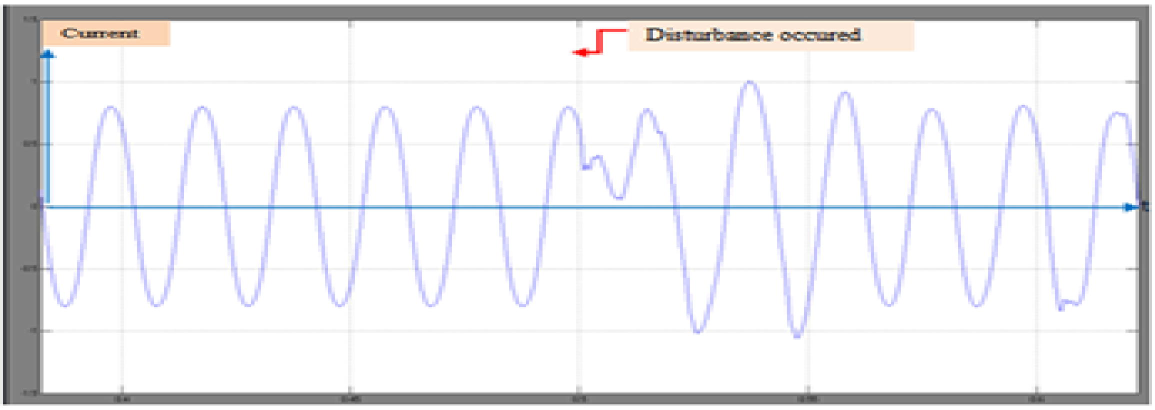

Figure 9. Steady state of motor input current waveform, (500mA/div), time(50ms/div).

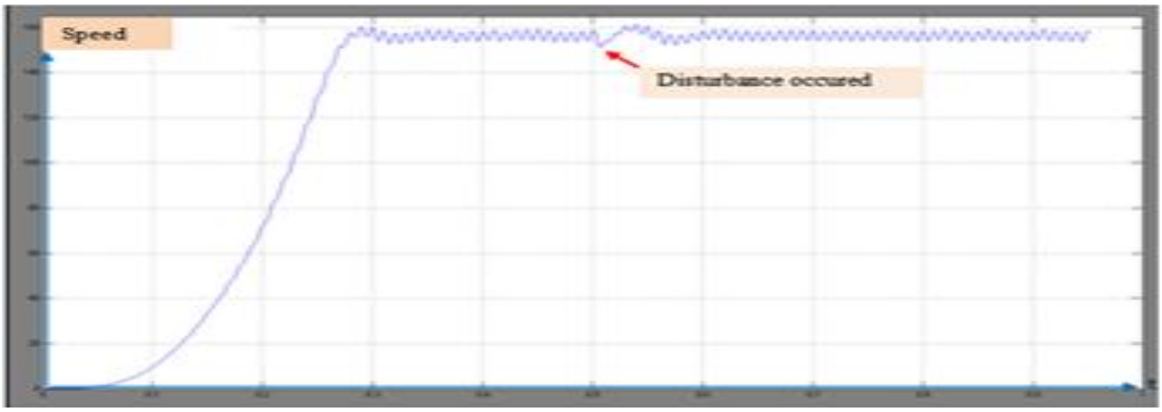

Figure 10. Steady state motor speed at source main frequency 50HZ, (20rad/s/div), time(s).

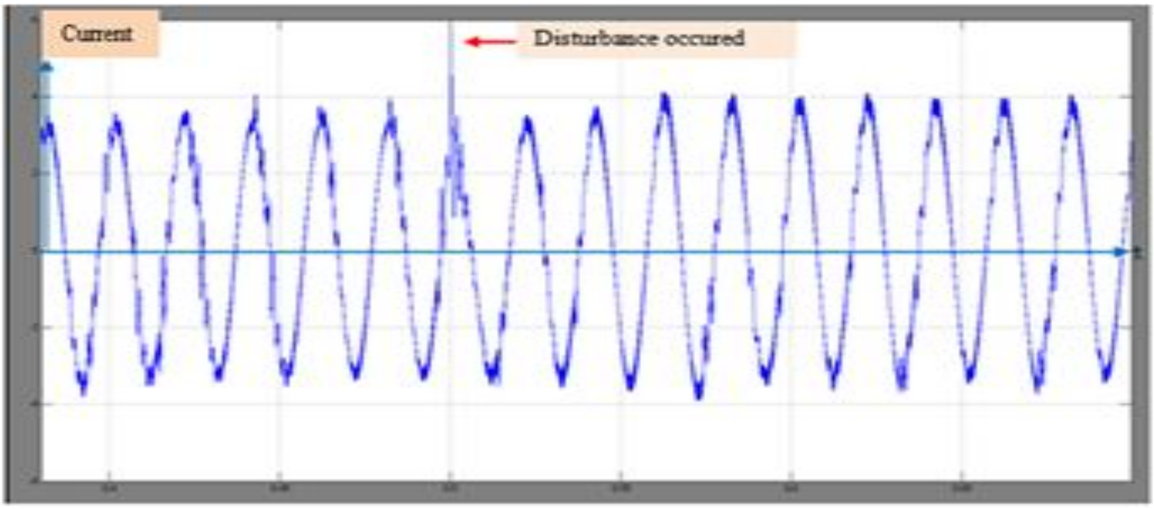

Figure 11. Steady state of supply current waveform, (2A/div), time(50ms/div).

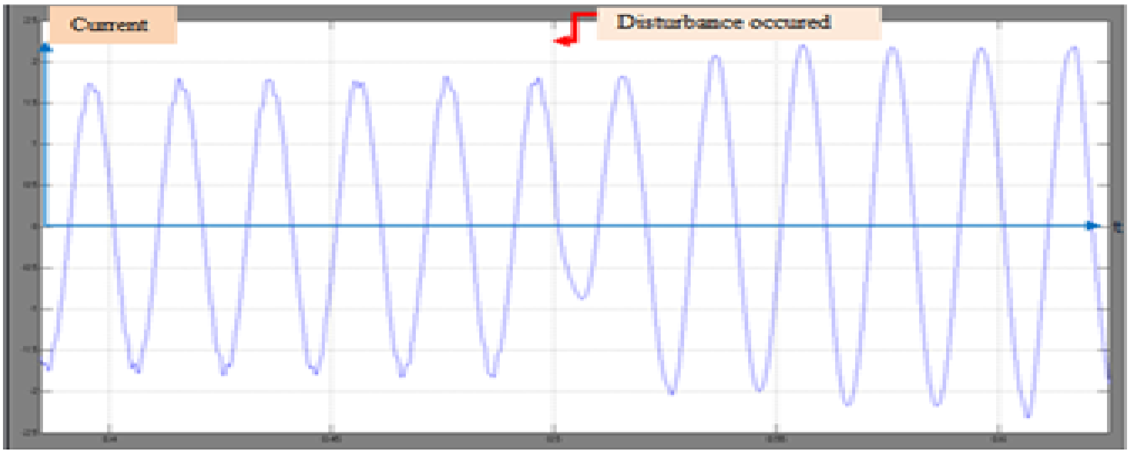

Figure 12. Steady state of motor input current waveform, (500mA/div), time(50ms/div).

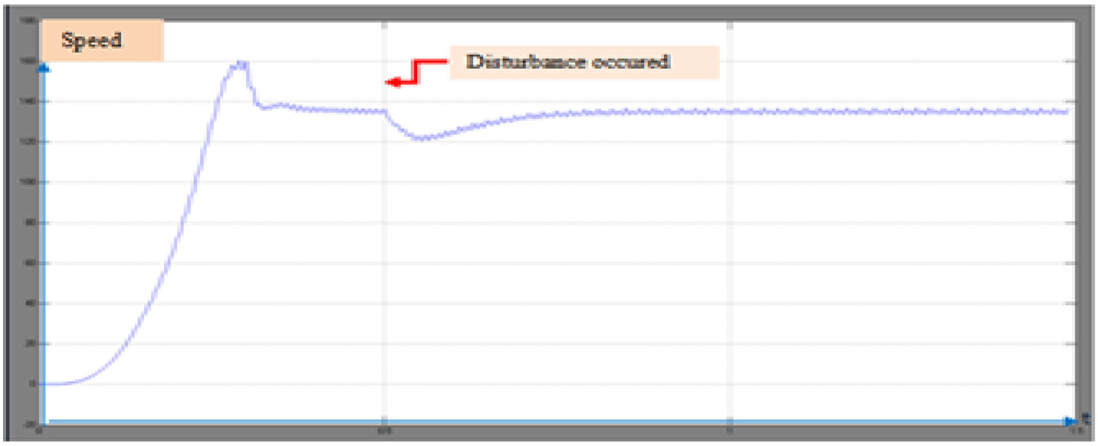

Figure 13. Steady state motor speed at source main frequency 50HZ, (20rad/s/div), time(s).

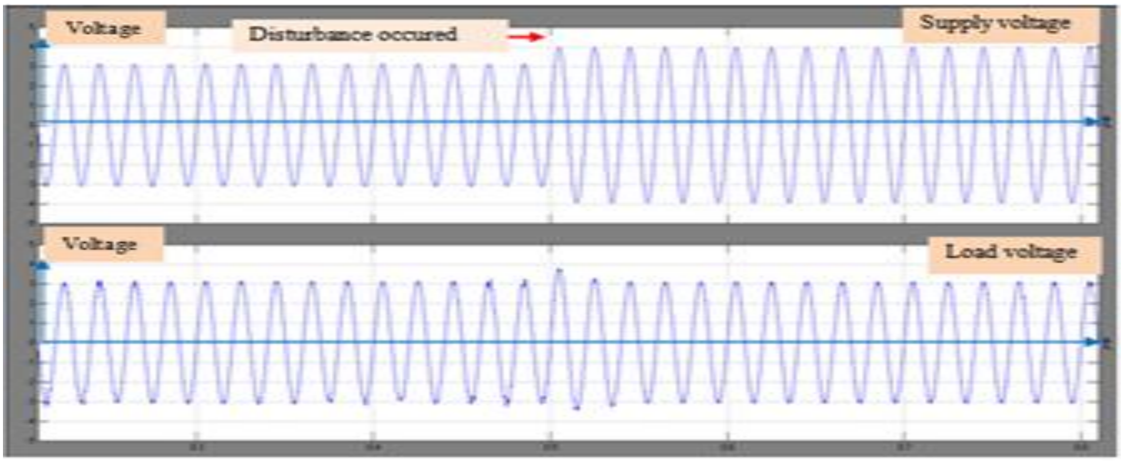

Figure 14. Steady state of supply voltage and load voltage waveforms at source main frequency 50HZ, (100V/div), time(100ms/div).

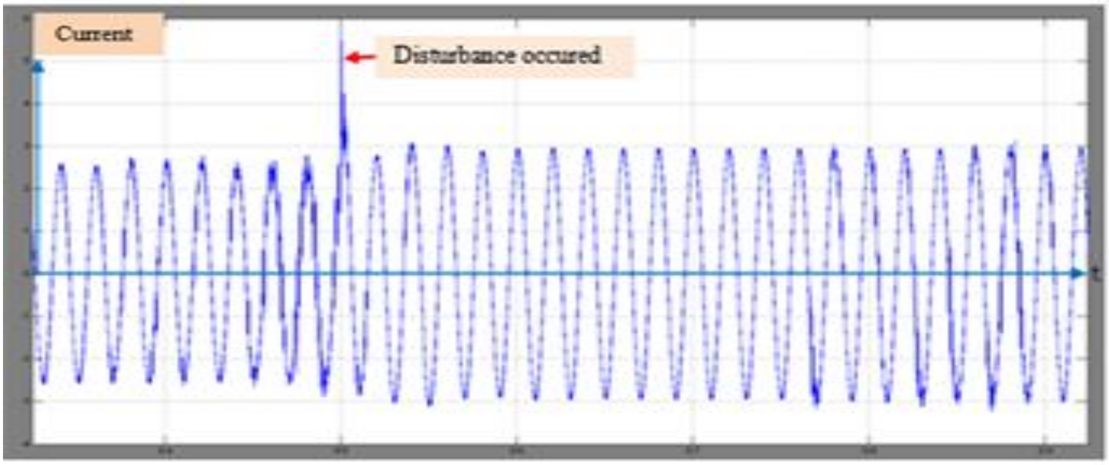

Figure 15. Steady state of supply current waveform, (1A/div), time(100ms/div).

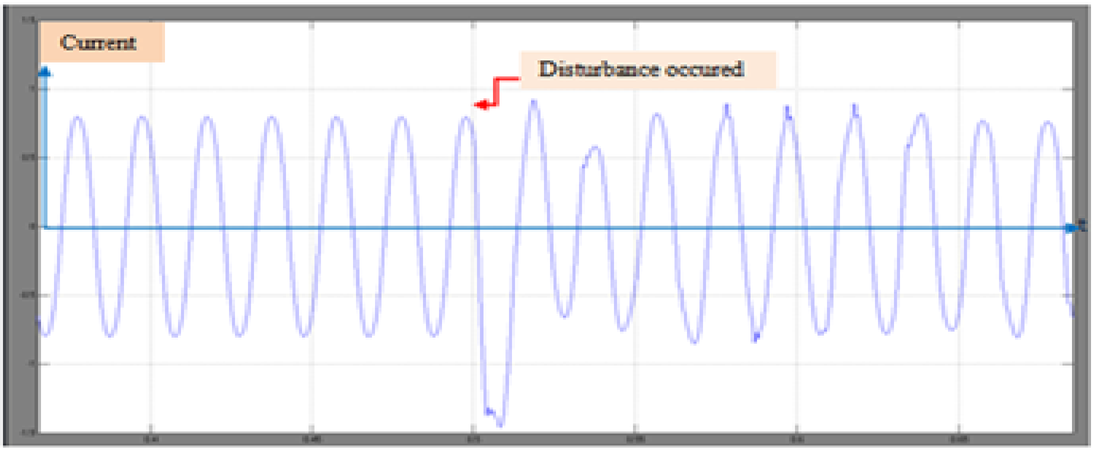

Figure 16. Steady state of motor input current waveform, (500mA/div), time(50ms/div).

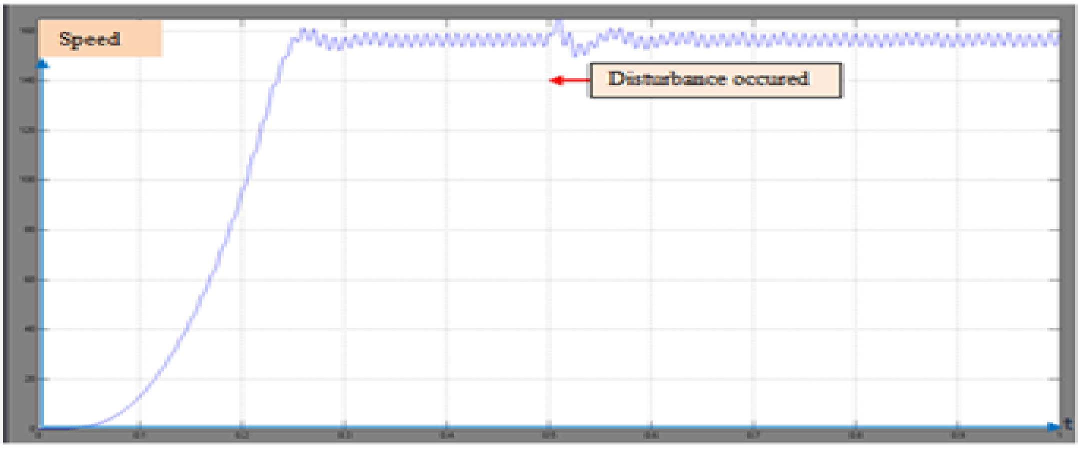

Figure 17. Steady state motor speed at source main frequency 50HZ, (20rad/sec/div), time(s).

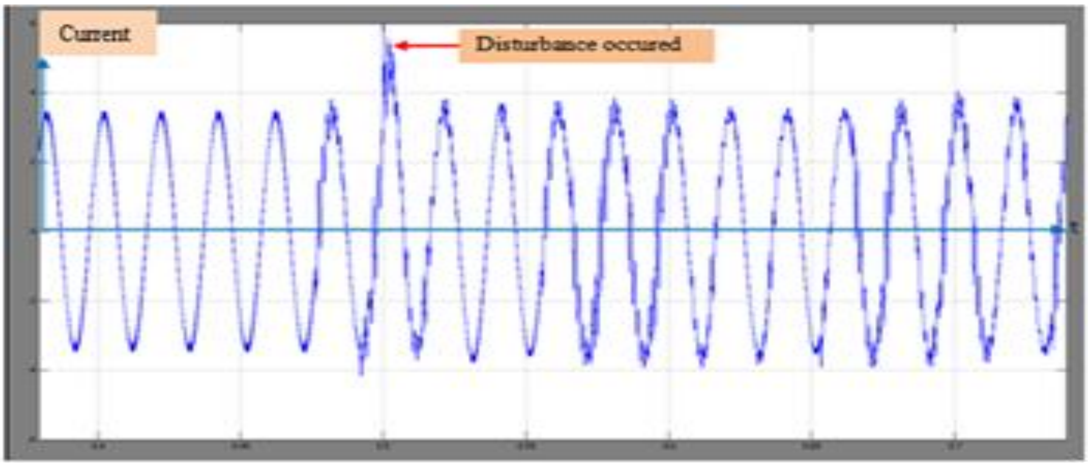

Figure 18. Steady state of supply current waveform, (1A/div), time(50ms/div).

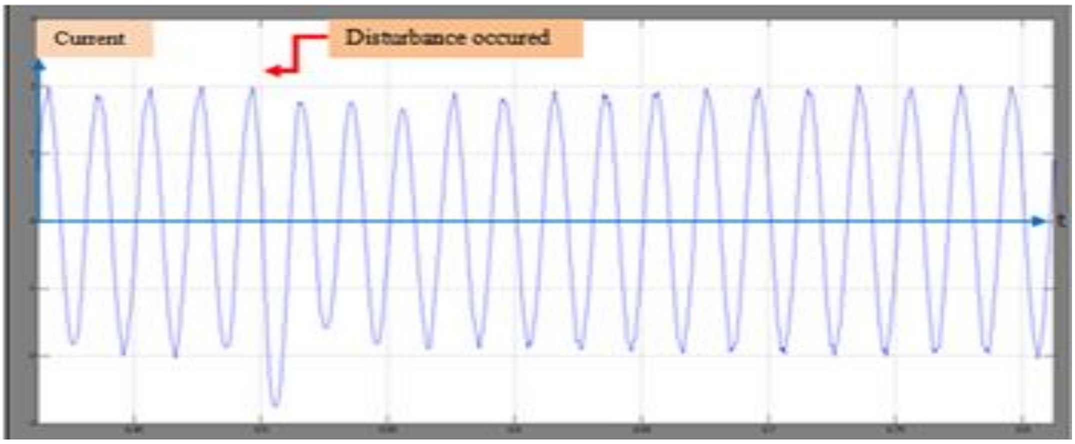

Figure 19. Steady state of motor input current waveform, (1A/div), time(50ms/div).

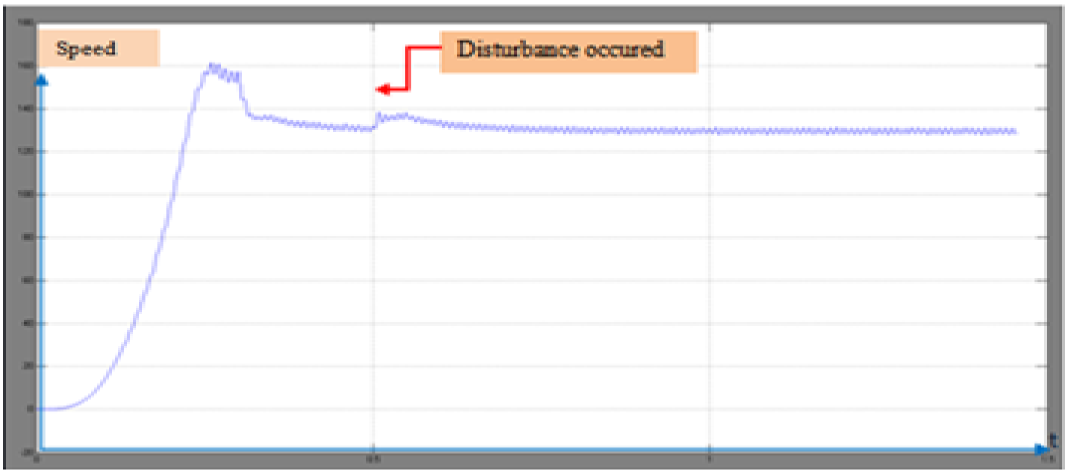

Figure 20. Steady state motor speed at source main frequency 50HZ, (20rad/sec/div), time(s).

Table 1. Name-plate data of the single phase capacitor run induction motor.

Induction motor	
Parameter	Value
Rated rms voltage	220 V
Rated frequency	50 HZ
Rated rms current	1.47 A
Number of poles	4
Capacitor	8 μ F ± 7%
Rated output	175 W
Rated speed	1140±40% rpm

Table 2. Motor parameters.

Parameter	value
Main winding stator resistance	43Ω
Main winding stator leakage reactance	34.05 Ω
Main winding rotor resistance	32.76 Ω
Main winding rotor leakage reactance	34.05 Ω
Main winding mutual reactance	466.60 Ω
Auxiliary winding stator resistance	23Ω
Auxiliary winding stator leakage reactance	40 Ω
turn ratio (aux/main)	1.1
Moment of inertia(J)	0.767x10^-3 Kg.m2
Friction coefficient	0.118x10 ^-3 N.m.s/rad

Table 3. Total harmonic distortion; for input current, load current and voltage across the load terminals when input voltage varied at (0.5sec).

Voltage variation	Load condition	THD_i	$THDi_L$	$THDv_o$
Rated rms input voltage dropped to(160V)	No-load	4.09%	3.80%	0.60%
	Half-full load	4.53%	1.97%	0.59%
	Full-load	4.47%	1.52%	0.51%
Rated rms input voltage increased to(280V)	No-load	3.29%	3.31%	0.40%
	Half-full load	3.44%	2.25%	0.46%
	Full-load	4.24%	1.32%	0.70%

Table 4. Input power (W) values according to line power factor variation at different supply voltage.

Input voltage (V)	Source Current (A)	Load applied on the motor (N.m)	Input power factor	Input capacity (VA)	Input power (W)
(160V)	1.76	0	0.355	281.6	99.9
	2.68	1.3	0.810	428.8	347.3
(220V)	1.76	0	0.250	387.2	96.8
	2.47	1.3	0.650	543.4	353.2
(280V)	2.12	0	0.170	593.6	100.9
	2.47	1.3	0.500	691.6	345.8