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Dynamic Analysis for Simulating the Effect of Power Quality on Sensitive Electronic Equipment

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Abstract: **With an increasing usage of sensitive electronic equipment power quality has become a major concern now. One critical aspect of power quality studies is the ability to perform automatic power quality data analysis. The impact of power quality on the operation of sensitive equipment has been illustrated through simulations in MATLAB SIMULINK. Such study is essential to predict the performance of modern loads and also to be able to explain why a specific load fails during a power quality event. The findings are reported in detail in this paper. The paper proposes a neural network solution to the indirect vector control of three phase induction motor including a real-time trained neural controller for the IM angular velocity, which permitted the speed up reaction to the variable load. The basic equations and elements of the indirect field oriented control scheme are given. The control scheme is realized by one recurrent and two feed-forward neural networks. The first one is learned in real-time by the dynamic BP method and the two FFNNs are learned off-line by the Levenberg-Marquardt algorithm with data taken by PIcontrol simulations. The final set up MSE of the LM algorithm is of 10-10. The graphical results of modeling show a better performance of the adaptive NN control system with respect to the PI controlled system realizing the same computational control scheme with variable load.**

Keywords: Power Quality, Wavelet Transform, Fourier Transform, Fuzzy Logic, Field oriented control, indirect vector control.

1. Introduction

Any variation in voltage, current, or frequency which may lead to an equipment failure or malfunction is potentially a Power Quality problem [1]. The classification and identification of power disturbances are governed by certain standards. The major cause of the problem is the increase in non-linear loads, which distort current and voltage waveforms. Modern electronic equipment is much more sensitive to such disturbances than traditional loads (lighting and motors) [2, 3]. The result is that the processes are interrupted, productivity is halted, and millions of dollars are lost.

To avoid this, the disturbances need to be first categorized so that relevant mitigation steps can be taken [4]. A technique based on fuzzy logic to categorize power quality events is described in this paper. The impact of power quality on the operation of sensitive equipment has been illustrated through simulations in MATLAB SIMULINK. The findings from this study will help to predict the performance of modern loads and also to be able to explain why a specific load failed during a power quality event.

Voltage Variations are caused by fault conditions and the energisation of large loads, where high starting currents are involved. The faults can cause a 'drop', 'rise' and 'supply void' in the supply voltage, and are also known as sag, swell and interruptions respectively. Voltage sag is normally caused by system faults, energisation of heavy loads and starting of large motors. Swells are usually associated with system fault conditions, but they are not as common as voltage sags.

The Neural Networks (NN) applications for identification and control of electrical drives became very popular in the last decade. In [1] an adaptive neuro-fuzzy system is applied for a stepping motor drive control. In [2] a multilayer perceptron-based-neural- control is applied for a DC motor drive. In [3] a recurrent neural network is applied for identification and adaptive control of a DC motor drive mechanical system. In the last decade a great boost is made in the area of induction motor drive control. The Induction Motor (IM), particularly the cage type, is most commonly used in adjustable speed AC drive systems [4]. The control of AC machines is considerably more complex than that of DC machines. The complexity arises because of the variable frequency power supply, AC signals processing, and complex dynamics of the AC machine [4, 5]. In the vector or Field-Oriented Control (FOC) methods, an AC machine is controlled as a separately excited DC machine, where the active (torque) and the reactive (field) current components are orthogonal and mutually decoupled so they could be controlled independently [4,7]. There exist two methods for PWM current controlled inverter – direct and indirect vector control, [4]. This paper will consider the indirect control method, where the slip angle, the direct and quadrature axes stator current set point components in stationary rotation frame are computed from the torque and rotor flux set points and used for vector control. There are several papers of NN application for AC motor drive indirect vector control. In [5] a Feedforward NN (FFNN) and Back propagation (BP) learning are used for angular velocity estimation and control of an IM, using only the stator current measurements. The authors of [6] presented a method of NN velocity estimation and IM control based on the flux, voltage and currents models. In [5] a neural controller is implemented based on a TMS320C30 microprocessor in order to emulate an indirect Field Oriented Control (FOC) of an IM drive. In an adaptive velocity controller is presented by a reference model, based on neural networks. In a model referenced robust method of NN velocity control is proposed that is based on neural identifier and a neural PI controller. In a NN based adaptive control of an IM is proposed. The NN learning algorithm is derived using the Lyapunov theorem of stability. In [5] the authors proposed an IM velocity control scheme, containing a conventional PI controller, a dynamic compensator, a neural IM parameter identifier and a NN load torque estimator. The NN identifier is used to estimate the IM parameters and so to tune the dynamic compensator gains and the output signal of the NN load torque estimator is used for feed forward control. In [7] it is proposed to use a NN in order to design a self tuning PI velocity controller incorporated in an IM indirect vector control scheme. The paper [8] proposed to use a NN as an adaptive feed forward IM velocity controller. In] a FFNN-based estimator of the feedback signals is used for induction motor drive FOC system. The authors of proposed two NN-based methods for FOC of induction motors. The first one used a NN flux observer in a direct FOC. The second one used a NN for flux and torque decoupling in an indirect FOC. The results and particular solutions obtained in the referenced papers show that the application of NN offers a fast and improved alternative of the classical FOC schemes. The present paper proposes a total neural solution of an indirect IM velocity vector control problem which assures fast response and adaptation to a variable load.

2. Models of the induction machine

2.1. A phase (a, b, c) model

The Induction Motor (IM) equations [6, 7], for stator and rotor voltages in vector matrix form are given as

$$v_{abcs} = r_s i_{abcs} + p\lambda_{abcs} \quad (1) \qquad v_{abcr} = r_r i_{abcr} + p\lambda_{abcr} \quad (2)$$

$$r_s = r_s I_3 ; \; r_r = r_r I_3 \quad (3)$$

Where,

$$v_{abcs} = (v_{as}, v_{bs}, v_{cs})^T ; v_{abcr} = (v_{ar}, v_{br}, v_{cr})^T ;$$
$$i_{abcs} = (i_{as}, i_{bs}, i_{cs})^T ; i_{abcr} = (i_{ar}, i_{br}, i_{cr})^T ;$$
$$\lambda_{abcs} = (\lambda_{as}, \lambda_{bs}, \lambda_{cs})^T ; \lambda_{abcr} = (\lambda_{ar}, \lambda_{br}, \lambda_{cr})^T \quad (4)$$

are voltage, current, and flux, stator and rotor, three dimensional (a, b, c) vectors, with given up phase components; rs and rr are stator and rotor winding resistance diagonal matrices, with given up equal elements rs and rr, respectively; I3 is an identity matrix with dimension three, and p. is a Laplacian differential operator. The vector-matrix block-form representation of the flux leakage is given by the equation

$$\begin{bmatrix} \lambda_{abcs} \\ \lambda^{?}_{abcr} \end{bmatrix} = \begin{bmatrix} L^{abc}_{ss} & L^{?abc}_{sr} \\ (L^{?abc}_{sr})^T & L^{?abc}_{rr} \end{bmatrix} \begin{bmatrix} i_{abcs} \\ i^{?}_{abcr} \end{bmatrix}$$

Where the stator, rotor and mutual block-inductance 3x3 matrices are:

$$L^{abc}_{ss} = \begin{bmatrix} L_{ls} + L_{ss} & -L_{ss}/2 & -L_{ss}/2 \\ -L_{ss}/2 & L_{ls} + L_{ss} & -L_{ss}/2 \\ -L_{ss}/2 & -L_{ss}/2 & L_{ls} + L_{ss} \end{bmatrix} \quad (6)$$

$$L^{?abc}_{rr} = \begin{bmatrix} L^{?}_{lr} + L_{ss} & -L_{ss}/2 & -L_{ss}/2 \\ -L_{ss}/2 & L^{?}_{lr} + L_{ss} & -L_{ss}/2 \\ -L_{ss}/2 & -L_{ss}/2 & L^{?}_{lr} + L_{ss} \end{bmatrix} \quad (7)$$

$$L^{?abc}_{sr} = \left[L^{?abc}_{rs} \right]^T = L_{ss} \begin{bmatrix} \cos\theta_r & \cos[\theta_r + (2\pi/3)] & \cos[\theta_r - (2\pi/3)] \\ \cos[\theta_r - (2\pi/3)] & \cos\theta_r & \cos[\theta_r + (2\pi/3)] \\ \cos[\theta_r + (2\pi/3)] & \cos[\theta_r - (2\pi/3)] & \cos\theta_r \end{bmatrix}$$

The matrix elements here are: Lls , Llr – stator and rotor leakage inductances; Lss, Lrr – stator and rotor winding inductances. Using the winding turns stator/rotor ratio n, the relative leakage inductance L'ls , could be written as

$$L^{?}_{lr} = n^2 L_{lr} \; ; \; \theta_r = \int_0^t \omega_r(\xi) d\xi + \theta_r(0)$$

where θr and ωr are the angular rotor position and the angular rotor velocity, respectively. Now, the voltage equations (1) and (2) could be expressed with respect to the stator in the final (a, b, c) form

$$\begin{bmatrix} v_{abcs} \\ v^{?}_{abcr} \end{bmatrix} = \begin{bmatrix} r_s + pL^{abc}_{ss} & pL^{?abc}_{sr} \\ (pL^{?abc}_{sr})^T & r^{?}_r + pL^{?abc}_{rr} \end{bmatrix} \begin{bmatrix} i_{abcs} \\ i^{?}_{abcr} \end{bmatrix}$$

Where the relative rotor voltage, current, flux and resistance values are:

$$v^{?}_{abcr} = n v_{abcr} ; \; i^{?}_{abcr} = (1/n) i_{abcr}$$
$$\lambda^{?}_{abcr} = n\lambda_{abcr} ; \; r^{?}_r = n^2 r_r .$$

2.2. A (q, d, 0) model

The (a, b, c) model is very complicated for control, so it could be simplified using a transformation to the (q, d, 0) form. The AC motor equations for the stator and rotor voltages in vector-matrix form are given as follows:

$$v_{qd0s} = r_s i_{qd0s} + \Omega\lambda_{qd0s} + p\lambda_{qd0s} \quad (11)$$
$$v'_{qd0r} = r_r i'_{qd0r} + \Delta\Omega\lambda'_{qd0r} + p\lambda'_{qd0r} \quad (12)$$
$$v_{qd0s} = (v_{qs}, v_{ds}, v_{0s})^T ; \quad v'_{qd0r} = (v'_{qr}, v'_{dr}, v'_{0r})^T$$
$$i_{qd0s} = (i_{qs}, i_{ds}, i_{0s})^T ; \quad i'_{qd0r} = (i'_{qr}, i'_{dr}, i'_{0r})^T$$
$$\lambda_{qd0s} = (\lambda_{qs}, \lambda_{ds}, \lambda_{0s})^T ; \quad \lambda'_{qd0r} = (\lambda'_{qr}, \lambda'_{dr}, \lambda'_{0r})^T$$

are: voltage, current, and flux, stator and rotor, three-dimensional (q, d, 0) vectors, with given up components; rs and rr are stator and rotor resistance diagonal matrices, given by (3); Ω, and Δ Ω are diagonal angular velocity matrices, given by

$$\Omega = \begin{bmatrix} \omega & 0 & 0 \\ 0 & -\omega & 0 \\ 0 & 0 & 0 \end{bmatrix} ; \quad \Delta\Omega = \begin{bmatrix} \omega - \omega_r & 0 & 0 \\ 0 & -(\omega - \omega_r) & 0 \\ 0 & 0 & 0 \end{bmatrix}$$

The vector-matrix block-form representation of the flux leakage is given by the equation

$$\begin{bmatrix} \lambda_{qd0s} \\ \lambda^{?}_{qd0r} \end{bmatrix} = \begin{bmatrix} L^{qd0}_{ss} & L^{?qd0}_{sr} \\ (L^{?qd0}_{sr})^T & L^{?qd0}_{rr} \end{bmatrix} \begin{bmatrix} i_{qd0s} \\ i^{?}_{qd0r} \end{bmatrix}$$

where the stator, rotor and mutual block-inductance 3x3 matrices are:

$$L_{ss}^{qd0}=\begin{bmatrix} L_{ls}+L_{ss} & 0 & 0 \\ 0 & L_{ls}+L_{ss} & 0 \\ 0 & 0 & L_{ls} \end{bmatrix}$$

$$L_{rr}^{?qd0}=\begin{bmatrix} L_{lr}^{?}+L_m & 0 & 0 \\ 0 & L_{lr}^{?}+L_m & 0 \\ 0 & 0 & L_{lr}^{?} \end{bmatrix}$$

$$L_{sr}^{?qd0}=\left[L_{rs}^{?qd0}\right]^{\mathrm{T}}=L_m I_3$$

Where Lm represents the mutual inductance. The (q, d, 0) model could be written in the stationary and synchronous frames taking the angular velocity equal to: ω = 0 and ω = ωe, where ωe corresponds to the angular velocity of the stator field. Now we could write the scalar electromagnetic torque equation which could be expressed in the following four forms:

$$T_{em}=\frac{3}{2}\frac{P}{2\omega_r}[\omega\lambda_{d-q,s}^{\mathrm{T}}i_{q-d,s}+$$
$$+(\omega-\omega_r)\lambda_{d-q,r}^{'\mathrm{T}}i_{q-d,r}^{'}]$$
$$T_{em}=\frac{3}{2}\frac{P}{2}\lambda_{d-q,s}^{\mathrm{T}}i_{q-d,s}$$
$$T_{em}=\frac{3}{2}\frac{P}{2}\lambda_{q-d,r}^{'\mathrm{T}}i_{d-q,r}^{'}$$
$$T_{em}=\frac{3}{2}\frac{P}{2}i_{d-q,r}^{'\mathrm{T}}i_{q-d,s}$$

where P is the number of poles and

$$\lambda_{d-q,s}=\left(\lambda_{ds},-\lambda_{qs}\right)^{\mathrm{T}};\quad \lambda_{d-q,r}^{'}=\left(\lambda_{dr}^{'},-\lambda_{qr}^{'}\right)^{\mathrm{T}}$$
$$i_{q-d,s}=\left(i_{qs},i_{ds}\right)^{\mathrm{T}};\quad i_{q-d,r}^{'}=\left(i_{qr}^{'},i_{dr}^{'}\right)^{\mathrm{T}}$$
$$i_{d-q,r}^{'}=\left(i_{dr}^{'},-i_{qr}^{'}\right)^{\mathrm{T}};\quad \lambda_{q-d,r}^{'}=\left(\lambda_{qr}^{'},\lambda_{dr}^{'}\right)^{\mathrm{T}}$$

If we know the output power Po of the IM, we could write the following relation for the torque with respect to the rotor angular velocity

$$T_{em}=\frac{P}{2}(P_o/\omega_r)$$

2.3. Field orientation conditions

The flux and torque equations decoupling must transform the stator flux, current and voltage vectors from (a, b, c) reference frame to (q–d, s) reference frame and then to stationary and synchronous reference frame. Fig. 1 illustrates the current and voltage vector representations in stator and rotor synchronous frames. Fig. 1 illustrates also the magnetic field orientation, where the rotor flux vector is equal to the d-component of the flux vector, represented in a synchronous reference frame (λ'e dr=λr), which is aligned with the d-component of the current in this frame. For more clarity, the current and flux orientation in the synchronous reference frame are shown on Fig. 2. So, the field orientation conditions are the following [7]:

$$\lambda_{qr}^{?e}=0;\ p\lambda_{qr}^{?e}=0;\ \lambda_r=\lambda_{dr}^{'e}$$

Taking into account that the rotor windings are shortcut (the rotor voltage is zero) and the field orientation conditions (24), the first two components of the equation (12), obtain the form

$$0=r_r^{'}i_{qr}^{'e}+(\omega_e-\omega_r)\lambda_{dr}^{'e}$$
$$0=r_r^{'}i_{dr}^{'e}+p\lambda_{dr}^{'e}$$

for the q-component of the rotor flux, it is obtained:

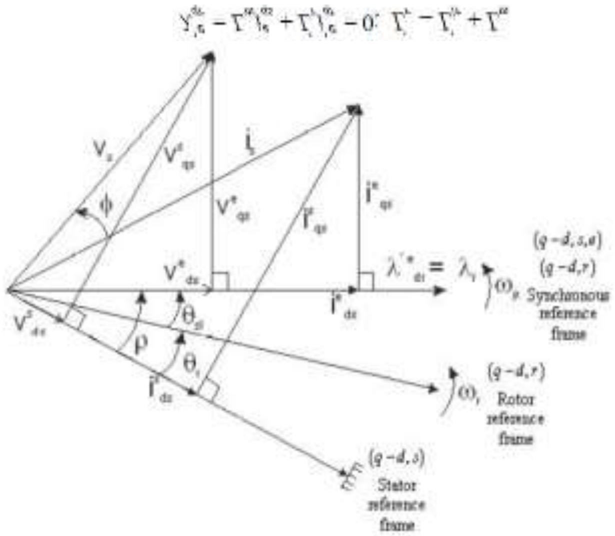

Figure 1: current and voltage vector representations in stator, rotor and synchronous reference frames

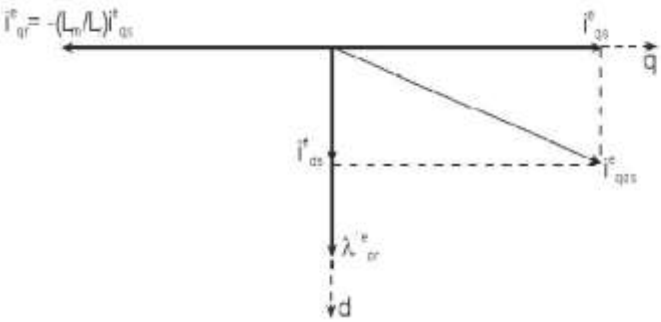

Figure 2: current and flux orientation in the synchronous reference frame

It is easy to obtain

$$i_{qr}^{'e}=-(L_m/L_r^{'})i_{qs}^{e}$$

Taking into account the condition , the torque equation could be written in the form

$$T_{em}=\frac{3}{2}\frac{P}{2}\lambda_{dr}^{'e}i_{qr}^{'e}$$
$$T_{em}=\frac{3}{2}\frac{P}{2}\frac{L_m}{L_r^{'}}\lambda_{dr}^{'e}i_{qs}^{e}$$

This equation shows that if the flux of the rotor is maintained constant, the torque could be controlled by the q- component of the stator current in synchronous reference frame. From the second equation of (25) it is easy to obtain the slipping angular velocity as

$$\omega_e-\omega_r=-r_r^{'}(i_{qr}^{'e}/\lambda_{dr}^{'e})_{30}$$
$$\omega_e-\omega_r=(r_r^{'}L_m/L_r^{'})(i_{qs}^{e}/\lambda_{dr}^{'e})_{31}$$

The final equations (29), (31) give the necessary basis for a direct decoupled field oriented (vector) control of the AC motor drive, where following Fig. 2, the q-component of the stator current produces torque and the d-component of the stator current produces flux.

2.4. Coordinate transformations

First of all we need to perform a coordinate transformation of stator variables from (a, b, c) to (q–d, s)

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reference frames and its inverse. For sake of simplicity we shall show only the stator currents transformation – the other vectors transformations are similar to that, which is

$$\begin{bmatrix} i^s_{qs} \\ i^s_{ds} \end{bmatrix} = \begin{bmatrix} 1 & -1/2 & -1/2 \\ 0 & -\sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} i_{as} \\ i_{bs} \\ i_{cs} \end{bmatrix}; \begin{bmatrix} i_{as} \\ i_{bs} \\ i_{cs} \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ -\frac{1}{2} & -\frac{\sqrt{3}}{2} \\ -\frac{1}{2} & \frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} i^s_{qs} \\ i^s_{ds} \end{bmatrix}$$

The (q–d, s) to (q–d, s, e) transformation of stator currents in synchronous reference frame and its inverse (see Fig. 1) are given by

$$\begin{bmatrix} i^e_{qs} \\ i^e_{ds} \end{bmatrix} = \begin{bmatrix} \cos\rho & -\sin\rho \\ \sin\rho & \cos\rho \end{bmatrix} \begin{bmatrix} i^s_{qs} \\ i^s_{ds} \end{bmatrix}; \quad \begin{bmatrix} i^s_{qs} \\ i^s_{ds} \end{bmatrix} = \begin{bmatrix} \cos\rho & \sin\rho \\ -\sin\rho & \cos\rho \end{bmatrix} \begin{bmatrix} i^e_{qs} \\ i^e_{ds} \end{bmatrix}$$

The combined stator current transformation from (a, b, c) to (q–d, s, e) synchronous reference frame and its inverse are obtained combining equations (32) and (33), as

$$\begin{bmatrix} i^e_{qs} \\ i^e_{ds} \end{bmatrix} = \begin{bmatrix} \cos\rho & f_1 & f_2 \\ \sin\rho & f_3 & f_4 \end{bmatrix} \begin{bmatrix} i_{as} \\ i_{bs} \\ i_{cs} \end{bmatrix}; \quad \begin{bmatrix} i_{as} \\ i_{bs} \\ i_{cs} \end{bmatrix} = \begin{bmatrix} \cos\rho & \sin\rho \\ f_2 & f_4 \\ f_1 & f_3 \end{bmatrix} \begin{bmatrix} i^e_{qs} \\ i^e_{ds} \end{bmatrix}$$

$$\begin{aligned} f_1 &= [-(1/2)\cos\rho - (\sqrt{3}/2)\sin\rho], \\ f_2 &= [(1/2)\cos\rho + (\sqrt{3}/2)\sin\rho], \\ f_3 &= [(1/2)\sin\rho + (\sqrt{3}/2)\cos\rho], \\ f_4 &= [-(1/2)\sin\rho - (\sqrt{3}/2)\cos\rho]. \end{aligned}$$

2.5. Stator current set point estimation

The indirect control is based on equation (31). If this equation holds, this is a necessary and sufficient condition to produce an adequate field orientation. This assure that the d-flux rotor component in synchronous reference frame λ'e dr will be aligned with the d-current stator component in synchronous reference frame ie ds (see Figs.1 and 2). Furthermore, this condition could be propagated to the set-point variables. The equation (31) could be expressed with respect to set point variables, so to obtain

$$\omega_e - \omega_r = (r'_r L_m / L'_r)(i^{e*}_{qs} / \lambda'^{e*}_{dr}) \quad 36$$

In the same manner, from equation (29), written for the set-point variables, we could obtain a relationship for the q-current stator set-point component in synchronous reference frame ie* qs , expressed with respect to torque and flux set-points, as it follows

$$i^{e*}_{qs} = (2/3)(2/P)(L'_r / L_m)(T^*_{em} / \lambda'^{e*}_{dr}); \lambda'^{e*}_{dr} = \lambda'^*_r \quad 37$$

The computation of the d-current stator set-point component in synchronous reference frame ie* ds, required some more mathematical manipulations. From the rotor part of the equation (15) we could extract the equation for the d-flux rotor component which is

$$\lambda'^e_{dr} = L'_{lr} i'^e_{dr} + L_m \left(i^e_{ds} + i'^e_{dr} \right) \quad 38$$

$$i'^e_{dr} = (\lambda'^e_{dr} - L_m i^e_{ds}) / L'_r \quad 39$$

$$i'^e_{dr} = -p\lambda'^e_{dr} / r'_r \quad 40$$

Equating the right parts of (40) and (39), and expressing the result with respect to the set-point variables, we could obtain the necessary d-current stator set-point component in synchronous reference frame e* ds i as follows

$$i^{e*}_{ds} = \left(r'_r + L'_r p \right)(\lambda'^{e*}_{dr} / r'_r L_m) \quad 41$$

If we accept that the rotor flux set- point (see (37)) is constant and its derivative is zero, the equation (41) is simplified as follows

$$\lambda'^{e*}_{dr} = L_m i^{e*}_{ds}; \quad i^{e*}_{ds} = \lambda'^*_r / L_m \quad 42$$

$$\omega_{sl} = \omega_e - \omega_r = (1/\tau_r)(i^{e*}_{qs} / i^{e*}_{ds}); \ \tau_r = (L'_r / r'_r); \ \rho = \int_0^t (\omega_{sl} + \omega_r) ds.$$

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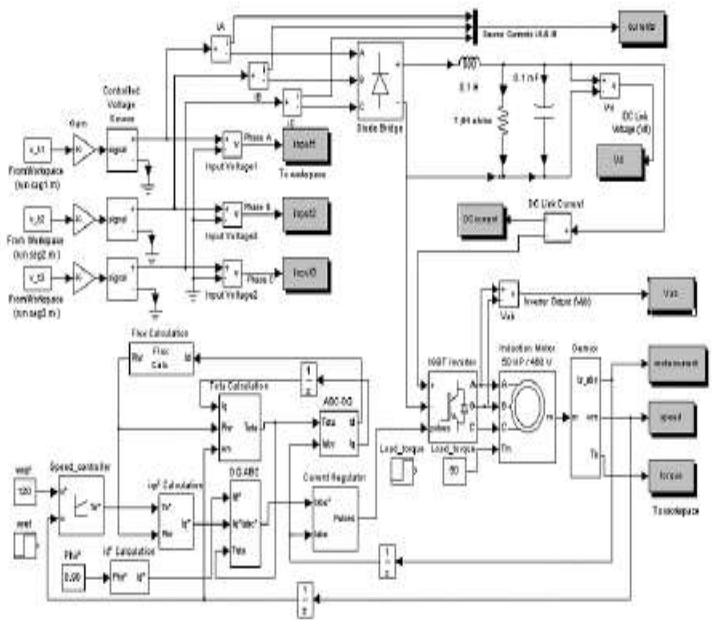
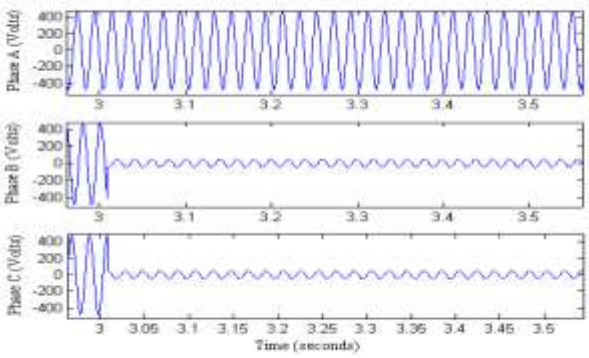


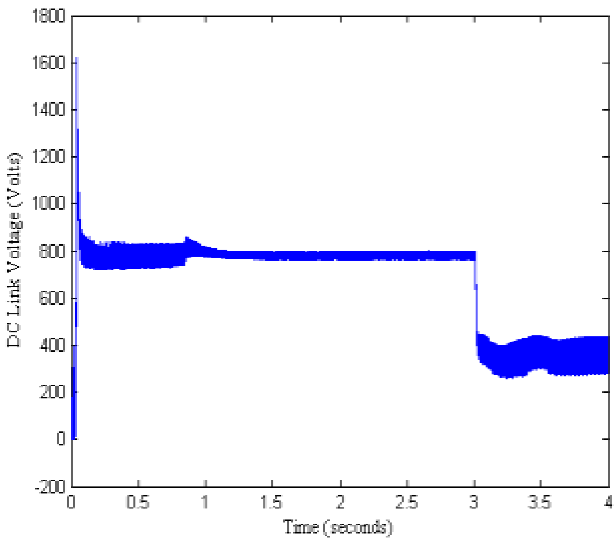
Fig. above illustrate disturbance inputs, the fall in DC link voltage and change in rotor speed for Case C corresponding to the sag event that occurs at time t= 3 seconds when Phase A and Phase B experience a lineto- ground fault. The fall in DC link voltage, and the rotor speed are observed for the period of the event. When normal supply resumes, the DC link voltage stabilizes at 780 Volts and the rotor speed at 120 radians per second.

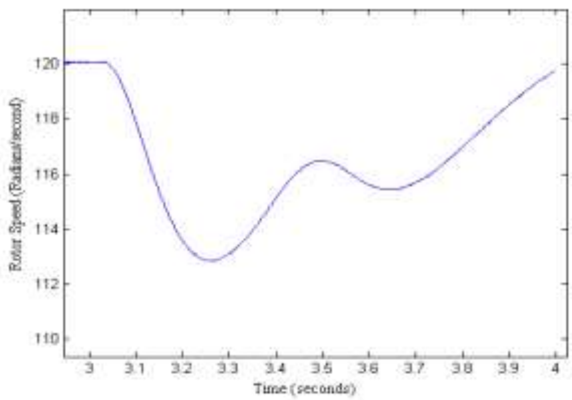
Source currents are positive as well as negative as they are coming from an AC source but since they pass through diodes they appear as square waveform. The variable response of the motor drive to the various inputs is evident from the above experimental results. No significant changes were observed for sag inputs with phase shift [3]. Phase angle jump has no effect unless the power electronics in the system use the phase angle information to determine firing instants. Our system did not have such a configuration. For unbalanced sag inputs, an unbalance in source/line currents was observed as shown in Fig. 11.

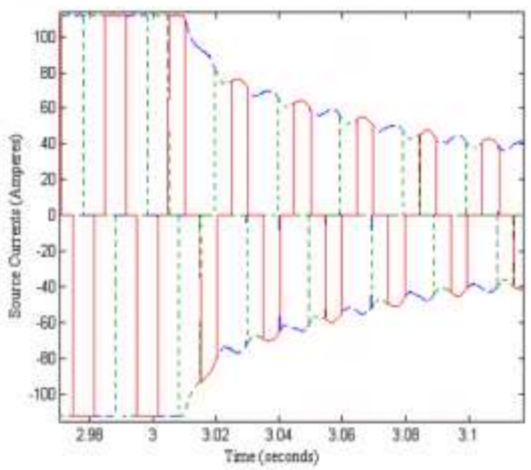
Also, the ability of the drive to ride-through a voltage sag event is dependent upon the energy storage capacity of the DC link capacitor, the speed and inertia of the load, the power consumed by the load, and the trip point settings of the drive. The control system of the drive has a great impact on the behavior of the drive during sag and after recovery. The trip point settings can be adjusted to greatly improve many nuisance trips resulting from minor sags which may not affect the speed of the motor.

3. Results:









4. Conclusion:

This paper presents a practically efficient tool for power quality data analysis and categorization as well as some initial findings of power quality impact on sensitive equipment. Behavior of a Vector controlled Variable Frequency Induction Motor Drive in the presence of sag events has been simulated as our initial investigation of impact of power quality on sensitive equipment. More work is currently in progress to validate the proposed tool, which will be reported in near future. The paper proposes a neural network solution to the indirect vector control of a three phase induction motor including a real-time trained neural controller for the IM angular velocity which permitted the speed up reaction to the variable load. The basic equations and elements of the indirect field oriented control scheme are given. The control scheme is realized by one recurrent and two feed-forward neural networks. The first one is learned in real-time by the dynamic BP method and the two FFNNs are learned off-line by the Levenberg-Marquardt algorithm with data taken by PI-control simulations. The final set up MSE of the LM algorithm is of 10-10. The graphical results of modeling shows a better performance of the adaptive NN control system with respect to the PI controlled system realizing the same computational control scheme with variable load.

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