Three Phase Induction Motor Torque Ripple Minimization Based On a Novel Nonlinear Dynamic Inverse Controller

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Abstract -A Nonlinear Dynamic Inverse (NDI) controller is proposed in this work to minimize the ripple torque based on space vector pulse width modulation (SVPWM) which is used in high power induction motors. The nonlinear dynamic inverse controller canceled a non-desirable response of the induction motor and enhancing the performance. This cancellation attempt by a carful nonlinear algebraic equations. First, a mathematical model of induction motor and linearization process were made decoupling between two inputs have achieved. Then the desired new dynamic are derived from implementing the proposed NDIC technique that reserves some benefits such as fast torque control, high torque at low speed, and fast speed response. Also, the proposed method greatly reduced the torque ripple which is the primary concerns of the classical hysteresis-based direct torque control (DTC) scheme and have an effect on the stator current distortion. The system stability with the designed (NDI-SVPWM) method is analyzed using the Lyapunov stability theory. Finally, the simulation results with MATLAB/Simulink achieved for a 50-HP IM drive. The results are verification proved that the proposed (NDI-SVPWM) system performs smaller torque ripple and faster torque response than the conventional SVM-based proportional plus integral (PI-DTC) method.

Keywords— Dynamic inverse; induction motor; ripple torque

1. INTRODUCTION

Decreasing the ripple torque in the induction motor has become a preoccupation of many researchers in recent years. It has many impacts on the effective performance of the induction motor (IM), increase efficiency, reduce losses and extend the life of its spare parts. Switching frequency which varies along operating conditions and the high torque ripple are two main problems, and they always come along with direct torque control (DTC) drives. As a result of induction motor (IM)'s features which are robustness, low price, reliability and free maintenance, it is used in industrial applications for a large scale. In general, when taking the total input cost the of energy consumption into consideration, the impact is too significant to be ignored. Thus, to reach the goal, the effective methods can be improving electric drives' efficiency through using the above-mentioned motor as well as controlling strategy to minimize energy loss. To stabilize state operation and to optimize an induction motor drive's efficiency, which is controlled by the field oriented method, varied schemes were developed.

A simple flux regulation for a direct torque control (DTC) of an induction motor (IM) to improve speed and torque estimations at low- and zero-speed regions was presented in [1] . In the presented system, the speed feedback for the closed-loop speed control was estimated using an extended Kalman filter, which requires heavy real-time computation. However, due to the simple structure of direct torque control - constant switching frequency controller (DTC-CSFC), small sampling time, hence large control bandwidth was possible. The performances of the speed sensorless hysteresis controller DTC-HC and DTC-CSFC were compared experimentally under different operating conditions. With the improved stator flux regulation, experimental results of the DTC-CSFC showed a significant improvement in speed and torque estimations at very low and zero-frequency operations.
A feedback which linearized direct torque control, and this control is with flux ripples for IPMSM drives and reduced torque was presented in [2]. In this paper the others succeed in reduction the torque and flux ripples to approximate (4-5) % with fast response to speed and torque varying conditions during the simulation process for uncertainties for some of the parameters. There is some difference between the simulation and experimental results.

A new predictable DTC method, via a voltage vector which has an optimal phase was illustrated in [3]. This approach can improve the classical DTC's dynamic reaction as well as decrease torque ripples and flux ripples. By using space-vector modulation, which has five segments, the voltage vector that obtained the fixed frequency of switching was synthesized.

To double fed induction machines, [4] have been considered designing a self-scheduled current controller. By using this objective, it was obtained that strong vital performance for all kinds of the automatic rotor speed, which was set in a detailed operating scope. Under consideration to the mechanical speed changing and being against dips of stator voltage, the controlled system's robustness and performance was demonstrated by the provided experimental results. Current-source converters (CSC) with five phases had been given a suitable scheme named space vector pulse width modulation (SVPWM) by [5]. This proposed scheme, which was verified utilizing a 1-kW prototype induction motor fed by CSC and both of them had five phases, efficiently gave out sinusoidal converter currents that were suited for energy conversion applications and motor drives.

An analytical model was presented by [6] to rapidly and accurately study the effect of such asymmetry of rotor poles on the torque ripple. The proposed approach was developed for symmetrical and asymmetrical geometries and validated against finite element models.

A model predictive torque control (MPTC) to be an adequate replacement for standard direct torque control (DTC) about drives of the induction motor (IM) was explained[7] . Via is portioning merely a part of control period to the active vector which is selected from regular MPTC, while rest time was portioned for a null vector, this paper proposed an optimized MPTC for reducing torque ripple. The active vector's duration was got and based on the theory of minimizing torque ripple. [8] also proposed another optimized MPTC controlling duty cycle, improving the duration and selecting vector at the same time with reducing both flux errors and torque

An optimal switching strategy to minimize torque ripple and switching frequency for Direct Torque Control of induction machines was illustrated [9, 10]. The most optimal voltage vectors will be used to improve DTC performances, i.e. reduced torque ripple and switching frequency. The identification of the vectors was based on operating conditions, specifically by examining the behaviors of torque error and switching frequency of error status produced from the hysteresis controllers. It can be shown that, the proposed optimal switching vectors according to the operating conditions can reduce the rate of change of torque and hence minimize torque ripple and switching frequency.

A comparative study of Direct Torque Control (DTC) and Predictive Torque Control (PTC) of three phase induction motor drive w.r.t flux ripple, torque ripple and also for dynamic response were presented by[11] and [12]. The control strategy combines of the classic PI controller to obtain good steady state response and a predictive controller scheme to achieve an excellent dynamic response. Simulation results were presented for both the schemes and the better features of PTC when compared with DTC are highlighted.

The combination of space-vector modulation and direct torque control was shown in this article by [13] to reduce the torque ripple contents in a medium-power adjustable speed drive system of the induction motor. The drive was supplied by a five level diode-clamped inverter. The results show improvement in the torque ripple contents compared with a conventional direct torque control scheme. The experimental data of the output voltage of the system for a diode-clamped multilevel inverter were also presented.

[14] and [15] gave a generalized two-vectors-based MPTC (GTV-MPTC) via separating the combination of the vector into double automatic voltage vectors. Via estimating vector's duration as well as a mix at the same time, under the defined worn function, minimizing torque error was got theoretically. Nevertheless, the computational burden was greatly heavier as well. Via a proper method chosen to decide the duration of the vector, random combinations of the vector could be avoided, which promotes optimized GTV-MPTC suited for real-time implementation. The results presented that, when compared to ex-MPTC no matter

there is controlling of duty cycle or not, optimized GTV-MPTC accomplishes a better performance while lowing frequency of sample for a wide speed scope.

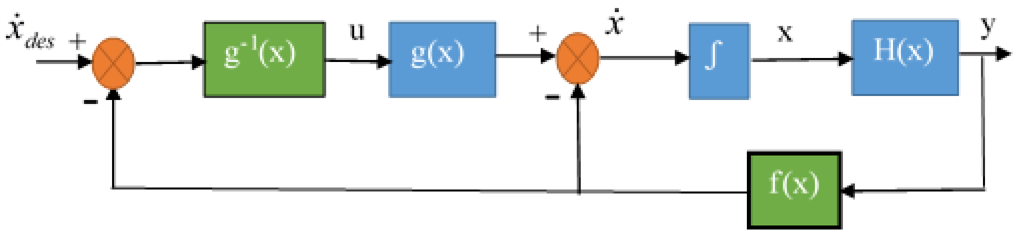

Figure 1: Dynamic inversion process.

Therefore, this paper proposes a new nonlinear dynamic inverse NDI techniques to minimize the ripple torque based on SVM. To apply the proposed NDI-SVPWM scheme, the decoupled dynamic model of an IM is first introduced by defining the two states (i.e., the stator flux and speed). Next, the nonlinear dynamic inverse is applied to the nonlinear IM model for obtaining an equivalent linearized model and then utilizing the linear control theory. The desired stator flux and rotor speed are adjusted with PID controller to get minimum allowed ripple torque with fast response. Consequently, the proposed method can significantly lessen the torque ripple which is the major weaknesses of the classical hysteresis-based DTC scheme. To confirm the performance of the proposed NDI-SVPWM scheme, simulation and investigations are carried out via MATLAB/Simulink of a 5-HP IM drive. Comparative results certainly indicate that the proposed NDI-SVPWM scheme realizes faster and small ripple torque than the conventional SVM-based PI-DTC approach under various conditions of speed and torque. The simulation results confirm that the proposed method reduces the torque ripple effectively while improving the dynamic response of the classical DTC method.

2. Nonlinear Dynamic Inverse Controller

Dynamic inversion is a controller synthesis technique by which existing deficient or undesirable dynamics are canceled out and replaced by desirable dynamics. This cancellation and replacement is accomplished by careful algebraic selection of the feedback function.

DI is a controller synthesis technique by which existing deficient or undesirable dynamics are canceled and replaced by designer-specified desirable dynamics. This cancellation and replacement are accomplished by careful algebraic selection of a feedback function [16] [17].

As previously suggested, the basic concept of dynamic inversion is quite simple. In general, the induction motor dynamics are expressed by [18]

$$\begin{aligned} \dot{x} &= F(x,u) \\ y &= H(x) \end{aligned} \tag{1}$$

where x is the state vector and u is the control vector; y is the output vector. For conventional the function F is linear in u. Equation (1) can also be re-written as

$$\dot{x} = f(x) + g(x)u \tag{2}$$

Where f is a nonlinear state dynamic function and g is a nonlinear control distribution function [19]. If we assume g(x) is invertible for all values of x, the control law is obtained by subtracting f(x) from both sides of equation (2) and then multiplying both sides by g-1(x).

$$u = g^{-1}(x)[\dot{x} - f(x)] \tag{3}$$

The next step is to command the induction motor to specified states instead of specifying the desired states directly, we specify the rate of the desired states, $\dot{x}$. By swapping $\dot{x}$ in the previous equation to $\dot{x}_{des}$, we get the final form of a dynamic inversion control law:

$$u = g^{-1}(x)[\dot{x}_{des} - f(x)] \tag{4}$$

Figure 1 shows the block diagram representation of the DI process. Even though the basic dynamic inversion process is simple, there are a few points to be emphasized. First, we assume g(x) is invertible for all values of x. However, this assumption is not always true. For example, g(x) is not generally invertible if there are more states than there are controls. Furthermore, even if g(x) is invertible (for example: g(x) is small), the control inputs, u, become large and this growth are a concern because of flux saturation. Since the dynamics of the induction motor, as well as sensor noise in the feedback loop [18], are also neglected during this primitive controller development to illustrate the process, a “perfect” inversion is not possible.

Dynamic Inversion is also essentially a special case of model-following. Similar to other model-following controllers, a DI controller requires exact knowledge of the model dynamics to achieve a good performance. To overcome these difficulties, a DI controller is normally used as an inner loop controller in combination with an outer loop controller, which is designed using other control design techniques.

3. APPLIED NONLINEAR DYNAMIC INVERSE TO INDUCTION MOTOR

Induction motor does not always behave like linear systems. In some work regimes, or in some (failure) scenarios, they behave in a nonlinear way. To control them, you therefore also need a nonlinear controller. And this controller should be as robust as possible, for the model never exactly matches reality. Nonlinear dynamic inversion (NDI) is one of the few control techniques in the literature that can be extended directly to nonlinear systems. The induction motor model with respect to a fixed stator reference frame (α, β) in the following form [20]:

$$\dot{\omega} = f(\omega) + \mathrm{B}(\omega).\,u_{s\alpha\beta} \tag{5}$$

with output: $y = C(\omega)$, where $\omega = [\Omega, \phi_{r\alpha}, \phi_{r\beta}, I_{s\alpha}, I_{s\beta}]^T$, and $u_{s\alpha\beta} = [u_{s\alpha}, u_{s\beta}]^T$ where ω is the state vector and $u_{s\alpha\beta}$ is the control vector.

$$f(\omega) = \begin{bmatrix} k(\phi_{r\alpha}I_{s\beta} - \phi_{r\beta}I_{s\alpha}) - \frac{T_l}{J} \\ -k_1\phi_{r\alpha} - p\Omega\phi_{r\beta} + k_1 m_{sr} I_{s\alpha} \\ p\Omega\phi_{r\alpha} - k_1\phi_{r\beta} + k_1 m_{sr} I_{s\beta} \\ k_1k_2\phi_{r\alpha} + pk_2\Omega\phi_{r\beta} - \epsilon I_{s\alpha} \\ -pk_2\Omega\phi_{r\alpha} + k_1k_2\phi_{r\beta} - \epsilon I_{s\beta} \end{bmatrix}, B(\omega) = [B_\alpha, B_\beta] = \begin{bmatrix} 0 & 0 \\ 0 & 0 \\ 0 & 0 \\ \frac{1}{\sigma L_s} & 0 \\ 0 & \frac{1}{\sigma L_s} \end{bmatrix} \tag{6}$$

$$y = \begin{bmatrix} C_1(\omega) \\ C_2(\omega) \end{bmatrix}, \quad \text{where } C_1(\omega) = \Omega \quad \text{and} \quad C_2(\omega) = \Omega_{r\alpha}^2 + \Omega_{r\beta}^2$$

where TL is the load torque, $k_1 = \frac{R_r}{L_r}$, $k_2 = \frac{m_{sr}}{\sigma L_s L_r}$, $k = \frac{pm_{sr}}{JL_r}$, $\epsilon = \frac{m_{sr}^2 R_r}{\sigma L_s L_r^2} + \frac{R_s}{\sigma L_s}$ $\quad \sigma = 1 - \left(\frac{m_{sr}{}^2}{L_s L_r}\right)$ is the new parameters of induction motor.

NDI is highly model dependent and requires an accurate system model in order to perform path-following with minimal error. It also assumes that the state derivative $\dot{\omega}$ depends on the input u. If this is not the case, finding an explicit inversion is not possible. These particular limitations can be alleviated by using approximate dynamic inversion [21]. For instance, let a nonlinear system be as follows:

$$\begin{aligned} \dot{x}_1 &= x_2 \\ \dot{x}_2 &= a(\omega) + b(\omega)u_s \end{aligned} \tag{7}$$

Then, in order to linearize this system, a nonlinear feedback is

$$u_s = \frac{1}{b(\omega)}(V_{des} - a(\omega)) \tag{7a}$$

So the linearized system is

$$\begin{aligned} \dot{x}_1 &= x_2 \\ \dot{x}_2 &= V_{des} \end{aligned} \tag{8}$$

The choice of the transformation x= f(ω) is the keystone of the method, as well as the inversion of the matrix b(ω). For IM, the nominal system is used in the control algorithm. The outputs are Ω and$(\phi_{r\alpha}^2 + \phi_{r\beta}^2)$. Let

$$\begin{gathered} x_1 = C_1(\Omega) = \Omega \quad , \quad x_2 = \dot{x}_1 = k(\phi_{r\alpha}I_{s\beta} - \phi_{r\beta}I_{s\alpha}), \; x_3 = C_2(\Omega) = \phi_{r\alpha}^2 + \phi_{r\beta}^2, \\ x_4 = \dot{x}_3 = -2k_{1n}(\phi_{r\alpha}^2 + \phi_{r\beta}^2) + 2k_{1n}m_{srn}(\phi_{r\alpha}I_{s\alpha} + \phi_{r\beta}I_{s\beta}) \end{gathered} \tag{9}$$

$$x_5 = C_3(\omega) = arctan\left(\frac{\phi_{r\beta}}{\phi_{r\alpha}}\right)$$

where k_{1n} ,m_{srn}, are the magnitude of k_1, msr taken in nominal value.

To set up the new state space representation of the system in the new coordinates (x_1, x_2, x_3, x_4, x_5), we introduce the notation Lfh , which is used for the Lie derivative of the state function h:$L_{fh} = \sum_{i=1}^{n} \frac{\partial h}{\partial x_i} f_i(x)$ with n the rank of the nonlinear system (f(x)=[f1(x),…..,fn(x)]T) and $L_f^i h = L_f(L_f^{i-1} h)$.

$$\dot{x}_1 = x_2 \quad , \dot{x}_2 = L_f^2 h_1 - \frac{k}{\sigma L_s}\phi_{r\beta} * u_{s\alpha} + \frac{k}{\sigma L_s}\phi_{r\alpha} * u_{s\beta}$$
$$\dot{x}_3 = x_4 \quad , \dot{x}_4 = L_f^2 h_2 + \frac{2k_1 m_{sr}}{\sigma L_s}\phi_{r\alpha} * u_{s\alpha} + \frac{2k_1 m_{sr}}{\sigma L_s}\phi_{r\beta} * u_{s\beta}$$
$$\dot{x}_5 = p\Omega + \frac{k_{1n} m_{srn}}{(\phi_{r\alpha}^2 + \phi_{r\beta}^2)}(\phi_{r\alpha} I_{s\beta} - \phi_{r\beta} I_{s\alpha}) \tag{10}$$

where $L_f^2 h_1$ and $L_f^2 h_2$, are given by:

$$L_f^2 h_1 = -kk_{1n} p\Omega(\phi_{r\alpha}^2 + \phi_{r\beta}^2) - k(k_{1n} + \epsilon_n)(\phi_{r\alpha} I_{s\beta} - \phi_{r\beta} I_{s\alpha}) - kp\Omega(\phi_{r\alpha} I_{s\alpha} + \phi_{r\beta} I_{s\beta}) \tag{11}$$

$$L_f^2 h_2 = (4k_{1n}^2 + 2k_{1n}^2 k_{2n} m_{srn})(\phi_{r\alpha}^2 + \phi_{r\beta}^2) + 2k_{1n} m_{srn} p\Omega(\phi_{r\alpha} I_{s\beta} - \phi_{r\beta} I_{s\alpha})$$
$$-(6k_{1n}^2 m_{srn} + 2k_{1n}\epsilon_n m_{srn})(\phi_{r\alpha} I_{s\alpha} + \phi_{r\beta} I_{s\beta}) + 2k_{1n}^2 m_{srn}^2 (I_{s\alpha}^2 + I_{s\beta}^2) \tag{12}$$

The dynamic inverse controller will be effective when sufficient condition must be provided. Matrix b(ω) is not singular matrix and has an inverse and $\phi_r^2 = (\phi_{r\alpha}^2 + \phi_{r\beta}^2) \neq 0$ Startup condition for induction motor. This nonlinear dynamic inverse can be interpreted as a nonlinear term compensation. It is like performing an inverse of the nonlinear model of the induction motor. Note that this dynamic inverse compensates terms like $\Omega(\phi_{r\alpha} I_{s\beta} - \phi_{r\beta} I_{s\alpha})$ and etc. This makes the feedback sensitive with respect to the high speed and particularly when the speed is estimated. From equation (6) and (7) can get the new linearized system:

$$\dot{x}_1 = x_2 \quad , \dot{x}_2 = V_{des1}, \quad \dot{x}_3 = x_4 \quad , \dot{x}_4 = V_{des2} \tag{13}$$

Now, the cancellation of undesired system is attempt but we must modify the state variable $(\dot{x}_2, \dot{x}_4)$ by new desired roots ($V_{des1} = \dot{x}_{2des}, V_{des2} = \dot{x}_{4des}$) these desired roots must be chosen carefully in algebraic control law to get the desired.There are many types of desired dynamics such as ride quality, flying quality, PI, PID and also can used the distinct poles. The inverse of the b(ω) is:

$$b^{-1}(\omega) = \frac{\sigma L_s}{\phi_{r\alpha}^2 + \phi_{r\beta}^2}\begin{bmatrix} \frac{-\phi_{r\beta}}{k} & \frac{\phi_{r\alpha}}{2k_1 m_{sr}} \\ \frac{\phi_{r\alpha}}{k} & \frac{\phi_{r\beta}}{2k_1 m_{sr}} \end{bmatrix} \tag{14}$$

So the Nonlinear dynamic Inverse NDI controller is:

$$\begin{bmatrix} u_{s\alpha} \\ u_{s\beta} \end{bmatrix} = \frac{\sigma L_s}{\phi_{r\alpha}^2 + \phi_{r\beta}^2}\begin{bmatrix} \frac{-\phi_{r\beta}}{k} & \frac{\phi_{r\alpha}}{2k_1 m_{sr}} \\ \frac{\phi_{r\alpha}}{k} & \frac{\phi_{r\beta}}{2k_1 m_{sr}} \end{bmatrix}\left(\begin{bmatrix} \dot{x}_{2des} - L_f^2 h_1 \\ \dot{x}_{4des} - L_f^2 h_2 \end{bmatrix}\right) \tag{15}$$

The following steps necessary to a chive the nonlinear dynamic inverse control for nonlinear system:
Make a transformation to a new state by feedback linearization for the nonlinear model of an induction motor by careful algebraic calculations.

- Make decoupling for inputs.
- Choose a desired suitable path to change and replace the undesired path roots.
- Design a robust outer loop controller to get a stable system with desired requirements.

Figure 2 shows a complete block diagram of the nonlinear dynamic inverse controller to 3-phase induction motor with PWM and inverter .

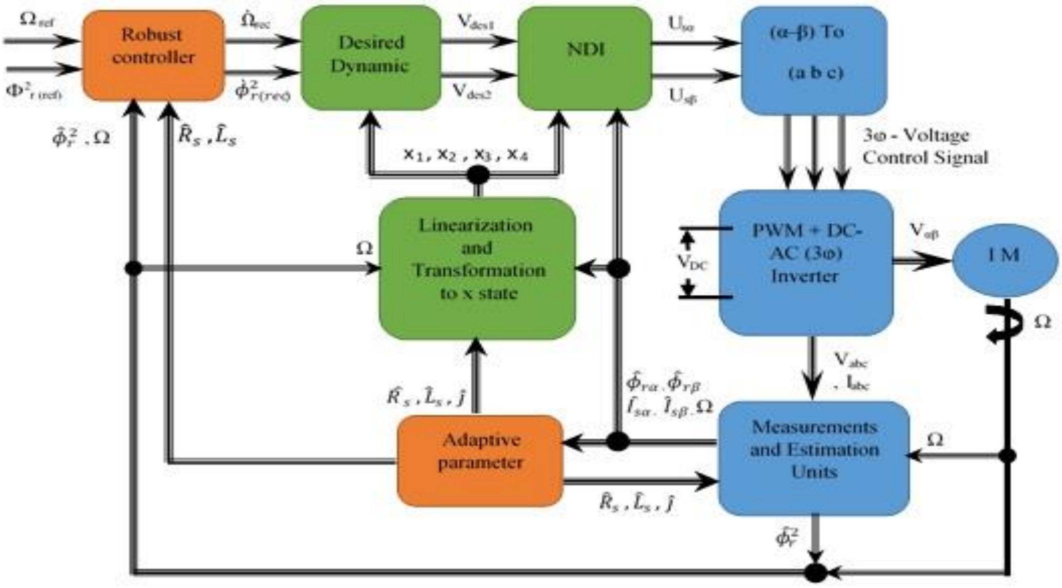

Figure 2: A block diagram of the dynamic inverse controller for an induction motor.

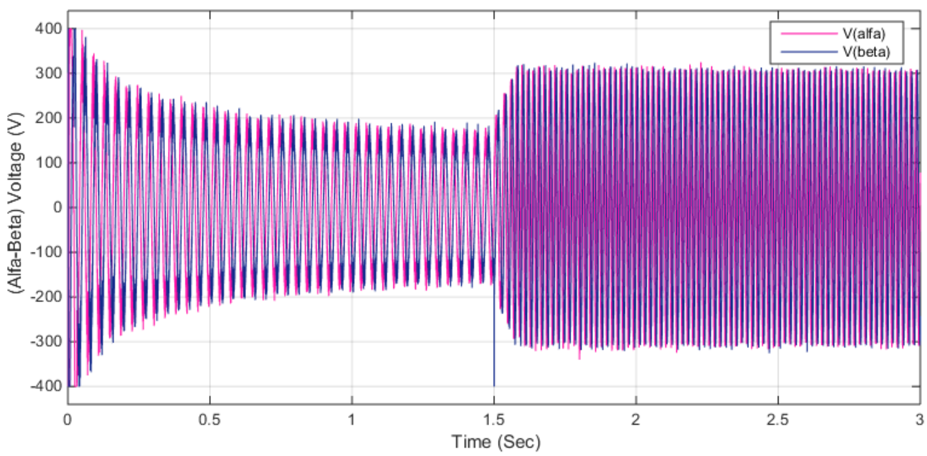

Figure 3: (α-β) control switching voltage.

4. RESULTS AND DISCUSSION

The simulation results for the 50hp induction motor illustrated in figures below with input desired speed is (80-160) Rad/sec and 0.8 web flux with load torque changed from (50-100) N.m. The parameters of the induction motor are listed in Table 1.

After fed the induction motor the desired speed and flux with the load the NDI controller already cancels all the undesirable dynamic response in the inner loop and replaced by the desired new dynamic response as in equation (15).

Figure 3 illustrates the (α-β) control voltage that is used as input to the SVPWM to switching the inverter that is fed by 530 V DC. This voltage is proportional to desired speed.

Figure 4 explained the line output voltage of the inverter and changed the amplitude and frequency according to desired speed that is used as input to the induction motor.

Figure 5 shows the smooth rotor speed behavior when applied the NDI controller in the model of induction motor with the fast and stable response at less than 0.1 Sec at rotor speed 80 Rad/Sec. Adding, when the desired speed changing from (80-160) Rad/Sec the rotor speed will follow the speed you want through 0.1 Sec, this delay due the inertia of induction motor. Ripple in speed appears in the same zooming figure through the operating time.

Figure 6 explained the developed torques in the induction motor with (SVPWM), (NDI-SVPWM) and the required load torque. The ripple torque is evident in SVPWM about (30-50)% at low speed and torque and (15-25)% at high speed and high torque, whereas the ripple torque is strongly subsidence to (1-3)% at low speed and to (1-2)% at high speed.

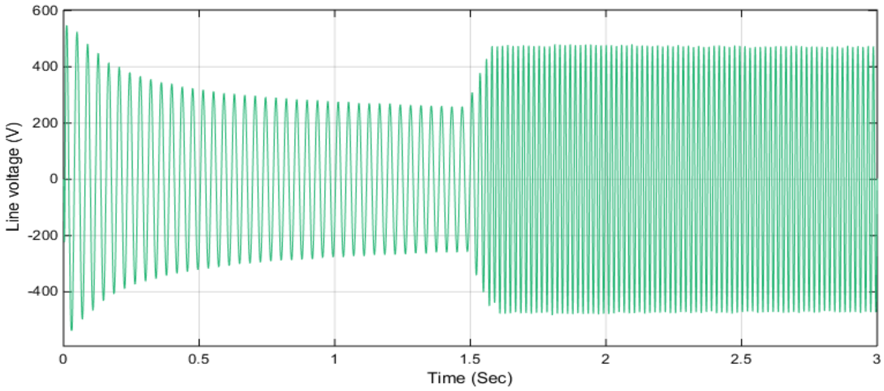

Figure 4: Input line voltage to the induction motor.

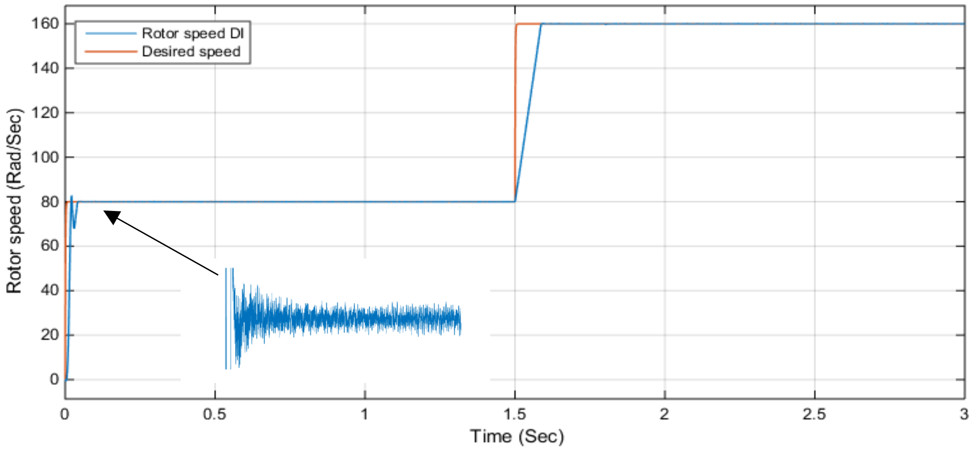

Figure 5: Rotor speed and desired speed.

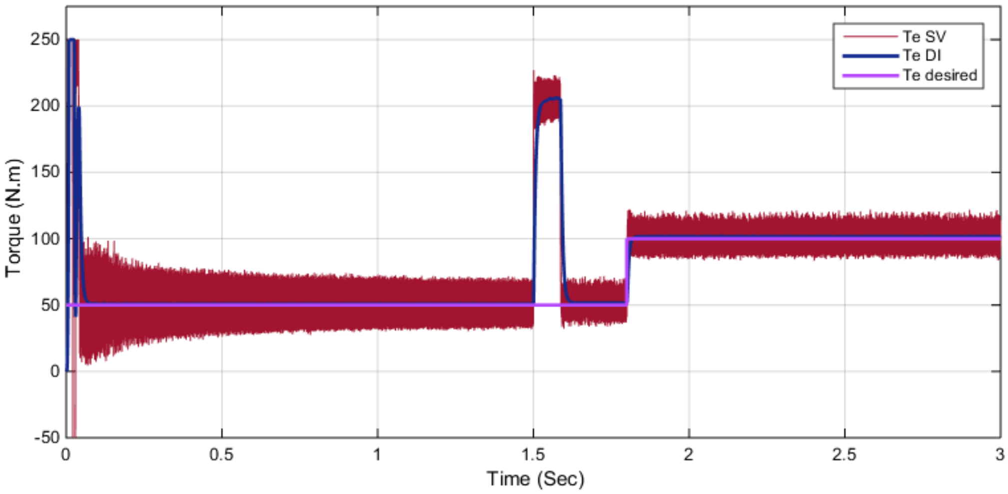

Figure 6: The developed torque for SVPWM, NDI-SVPWM and desired torque.

Figure 7 shows the line current of the induction motor at varying speed and torque. The amplitude is proportional to the magnitude of the load torques and rotor speed as expected. Also the current is suddenly jump to high value when the speed increased from (80 to 160) Rad/Sec and then back to steady state at 0.1 sec. Finally, the current increase normally when the torque is increased from (50 to 100) N.m.

Figure 8 illustrates the (d-q) current in estimation unit changing with speed and developed torque.

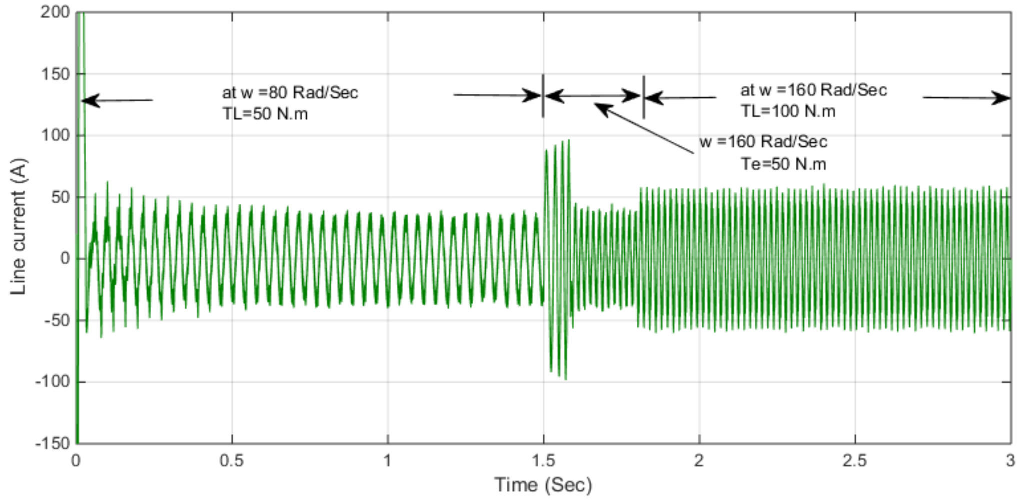

Figure 7: Line current with NDI controller at varying speed and torque.

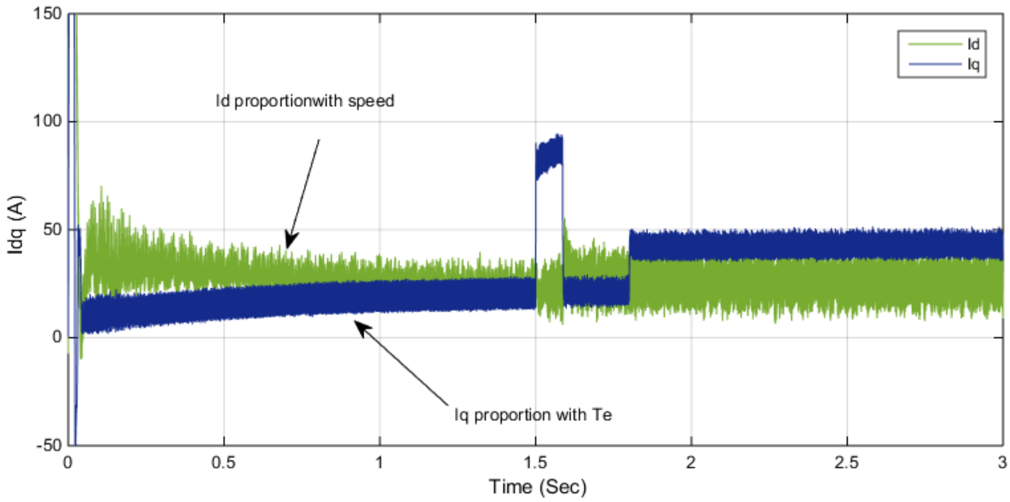

Figure 8: Estimated (d-q) current of IM

Table 1: Three phase induction motor parameters.

Symbol	parameter	value
V	Rated voltage	480 volt
P	Power	50 hp
f	Frequency	60 Hz
F	Friction factor	0.01 (N.m.s)
Rs	Stator resistance	0.087 Ω
Rr	Rotor resistance	0.228 Ω
Ls	Stator inductance	0.8e-3 H
Lm	Mutual inductance	34.7e-3 H
Lr	Rotor inductance	0.8e-3 H
J	Inertia	01.662 (kg.m^2)
p	Pole pairs	2

5. CONCLUSION

Simulation results show the proposed method is very active to minimize the ripple torque of the induction motor with the deferent condition in speed and torque. NDI-SVPWM strategy accomplishes faster and low ripple torque than the conventional space vector modulation based on PI-DTC and VSM methods under different conditions speed and torque. Both the simulation and experimental results confirm that the proposed method reduces the torque ripple effectively while improving the dynamic response of the classical DTC method. While, choosing the desired dynamics and a suitable controller for the model of the induction motor to follow the desired dynamics attempt by group of PID controllers with torque and speed regulator.

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