Sensorless Vector Control of Induction Motor

M.Tech. Dissertation

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requirements for the award of the degree

of

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in

Power Systems

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CERTIFICATE

I hereby certify that the work which is being presented in the M.Tech. Dissertation entitled **“Sensorless Vector Control of Induction Motor”,** in partial fulfillment of the requirements for the award of the **Degree of Master of Technology in Power Systems** and submitted to the Department of Electrical Engineering of Delhi Technological University is an authentic record of my own work carried out under thesupervision of **Prof. Mukhtiar Singh (Professor), EE Department**.

The matter presented in this report has not been submitted by me for the award of any other **Degree/Diploma** elsewhere.

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Abstract

In this work Scalar Control, Vector Control and Sensorless Vector Controlled Induction Motor drive is explained. In which it is shown that sensorless control is superior in case of control. Scalar control is simple and a low-cost technique but it gives poor transient response and it is unable to control two variables torque and magnetic flux that’s why it cannot be used in high performance application. Since vector control is complex and of high price but it is a high-performance control system which is valid in both steady state and transient conditions. It has fast dynamic response and speed regulation is good. This can be better performed in sensorless vector control in which speed sensor is not used which clearly lowers the cost and the adaptive method used here is for better estimation of motor parameters

Contents

1 Introduction 6

2 Literature Review 10

3 Scalar Control of Induction Motor 25

3.1 Γ Equivalent Circuit of An Induction Motor 25

3.2 Principles of the Constant Volts/Hertz Control 26

3.3 Scalar Speed Control System 31

3.4 Mechanical Characteristics of an Induction Motor 31

3.5 Γ' Equivalent Circuit of an Induction Motor 32

3.6 Principles of Torque Control 34

3.7 Scalar Torque Control System 35

4 Vector Control of Induction Motor 37

4.1 Space Vectors and Transformations 38

4.2 Principles of Vector Control 40

4.3 Direct Vector Control 42

4.3.1 Direct Vector Control Sensing Induced EMF and Current 45

4.4 Indirect Vector Control 47

5 Sensorless Vector Control 52

5.1 For Low-Performance Application 53

5.2 For High-Performance Drives 54

5.3 Model Reference Adaptive Systems 54

5.3.1 Popov's Hyperstability Theorem 56

5.3.2 Speed tuning Signal $\in_{\omega} = \mathbf{Im}(\lambda_R^s \hat{\lambda}_{Rref}^s)$ 57

5.3.3 Speed tuning Signal $\in_e = \mathbf{Im}(e\hat{e}_{ref})$ 58

5.3.4 Speed tuning Signal $\in_{\Delta e} = \mathbf{Im}(\Delta e i_{sref})$ 59

5.3.5 Speed Tuning Signal $\in_{\Delta e'} = \mathbf{Im}(\Delta \bar{E} \delta i_{sref})$ 60

6 Result 62

List of Symbols

E_d, E_q d & q components of Back EMF from reference model

E'_d, E'_q d & q components of Back EMF from adaptive model

f Supply Frequency, Hz

I_s Stator Current, A

I_R^*, I'_R Phasor of Rotor Current and transformed version A/ph

I_M, I'_M Phasor of Magnetizing Current and transformed version A/ph

I_{sref} Reference magnitude of Stator Current, A

$I_{s\theta ref}$ Reference Flux Producing Current, A

I_{sTref} Reference Torque Producing Current, A

$\vec{I}$ Stator Current Space Vector

$\vec{I}_{Rr}$ Rotor Current Space Vector in Rotor Reference Frame

I_{Rs} Rotor Current in Stator Reference Frame

I_f, I_T Field and Torque Producing current

I_{ss} Stator current in stator reference frame

L_M, L'_M Mutual Inductance, Transformed Mutual Inductance, H

L_R Rotor Inductance, H

L_s Stator Inductance, H

L_l Leakage Inductance, H

P_M Mechanical Power, W

P Number of Poles

R_R, R_R' Rotor Resistance, Transformed Rotor Resistance Ω / ph

P_E Electrical Power, W

R_s Stator Resistance, ohm

s Slip

T Developed Torque, N-m

V_s Stator voltage, V

$\vec{V}$ Stator Voltage Space Vector

V_{sd}, V_{sq} d, q component of stator voltage in stationary reference frame

ω_R Rotor Angular Speed

ω_M Speed of P-pole Motor, rad/sec

ω Supply Frequency, Radian, rad/sec

ω_o Speed of a 2-pole motor, rad/sec

ω_R^* Reference Speed, rad/sec

ω_{Rref} Reference Rotor Speed, rad/sec

ω_R Rotor Speed, rad/sec

λ_s Magnitude of Stator flux

λ_{Rref} Magnitude of rotor flux reference

$\vec{\lambda}_s$ Space Vector Stator Flux Linkage

$\lambda_{sd}, \lambda_{sq}$ Stator flux linkage in d and q reference frame

$\lambda_{Rd}, \lambda_{Rq}$ Rotor flux linkage in d and q reference frame

λ_{Rs} Rotor Flux linkage in stator reference frame

τ, τ' Rotor Time Constant, Transformed Rotor Time Constant

λ Flux

θ Angle with the reference axis.

θ_f Angle Between Field Reference Frame and stator reference frame

List Of Figure

Figure 1: *Block Diagram Of Electric Drives* -- 8

Figure 2 : Γ Steady State Equivalent Circuit of An Induction Motor---------------------- 26

Figure 3: *Torque vs Slip Speed Characteristics of IM with Constant Stator Flux* ---------- 29

Figure 4: *V vs f of Constant V / f Control* --- 30

Figure 5: *Mechanical Characteristics of IM in Constant V / f control* ----------------- 32

Figure 6: *Dynamic Γ′ Equivalent Circuit Of Induction Motor* ---------------------------- 33

Figure 7: *Block Diagram Of Scalar Control (Torque)* -- 35

Figure 8: *Direct Vector Control* -- 43

Figure 9: *Phasor Diagram of Indirect Vector Control* --------------------------------------- 47

Figure 10: *Basic MRAS Scheme* -- 55

Figure 11: *Equivalent Non – Linear Feedback System* --------------------------------------- 56

Figure 12: Rotor and reference Rotor Speed of Vector Controlled Induction Motor Drive-- 63

Figure 13:Electromagnetic torque and Reference torque of vector controlled Induction Motor Drive -- 63

Figure 14:Stator Current and DC Bus Voltage of Vector controlled IM Drive --------- 63

Figure 15: Error Signal--- 63

Figure 16:Rotor Speed, Reference Speed & Estimated Speed of sensorless Vector Controlled IM Drive-- 63

Figure 17:Stator Current and DC Bus Voltage of Sensorless Vector Controlled IM Drive --- 63

Figure 18:Electromagnetic Torque and Reference Torque of Sensorless controlled IM Drive-- 63

Figure 19:Speed Error Signal--- 63

Figure 20: Constant V/f control Ref. Speed 500rpm Load Torque[0 15 5] at [0 1 2]sec --- 63

Figure 21: V/f Control reference speed 500rpm Load Torque 10 Nm------------------- 63

1 Introduction

In the past, wherever variable-speed operation was required DC motors were used. Because of the fact that flux and torque can be controlled effortlessly by way of the field and armature current. Particularly, the separately excited DC motor is used in for fast reaction and 4-quad. operation with good operation near 0 speed. However, there are disadvantages too, because of the life of the commutator and the brushes. that is, they require regular upkeep; these DC machines cannot be utilized in abrasive environments and they have limitations on commutator adequacy under excessive speed, excessive-voltage operational situations. These can be overcome with the aid of the AC motors, which have easy and rough shape, are robust and immune to heavy overloading. Compared to DC machines in size these AC can machines can be designed with high ratings with low weight and inertia.

Variable-speed AC drives were used to perform relatively trouble-free roles in applications which avoid using DC machines, both due to commutator limits. Due to cost, high switching frequency and lower cost of AC machines are decisive component in Multi-motor systems. However, due to the advancement in power electronics, trending low cost high efficient, and single motor AC drives compete with low cost DC drives. The cage induction motor has a particular cost benefit over all the AC drives. It is easy and rugged and is one of the most inexpensive machines to be had in any respect of ratings. VFD can compete with highly efficient 4-quadrant DC drives using AC motors and highly efficient static converters.

These days, cutting-edge power electronics and drives are used in electrical in addition to mechanical enterprise. The power converter circuits used with drives, providing either DC or AC outputs, and working from both type of power supply. Right here we can spotlight the most essential components which are common to every kind of drive converters. As there are many distinctive kinds of converters, all besides very low-power ones are primarily based on a few shapes of digital switching. The need to adopt a switching strategy is shown in the Wrist example, in which outcomes are explored in a few intensities. To achieve high-efficiency power conversion, switching is essential, but that the ensuing waveforms are unavoidably less than perfect from the point of view of the motor. The controlled rectifier offers adjustable voltage for the armature, presenting speed control. On every occasion, motor or generator is used, we assume that the rotation of those machines is totally managed most effective by way of the carried-out voltage and frequency of the source. However, the rotation speed of an electrical system can be managed exactly also via imposing the idea of power. The structures which control the motion of the electrical machines, are called electric drives. an average power gadget is assembled with an electric motor (can be numerous) and a sophisticated manipulate system that controls the rotation of the motor shaft. Now days, this manipulate can be executed easily with the assist of software. So, the controlling turns into an increasing number of correct and this idea of pressure also presents the ease of use. This drive device is widely utilized in massive quantity of commercial and home applications like factories, transportation structures, textile mills, fanatics, pumps, automobiles, robots etc. Drives are hired as top movers for diesel or petrol engines, gas or steam generators, hydraulic motors and electric cars.

Electromechanical device for changing electric energy into mechanical power called Electric Drive. It is a device designed to convert electric energy into mechanical energy using electrical control. The machine employed for motion control is known as an electrical drive.

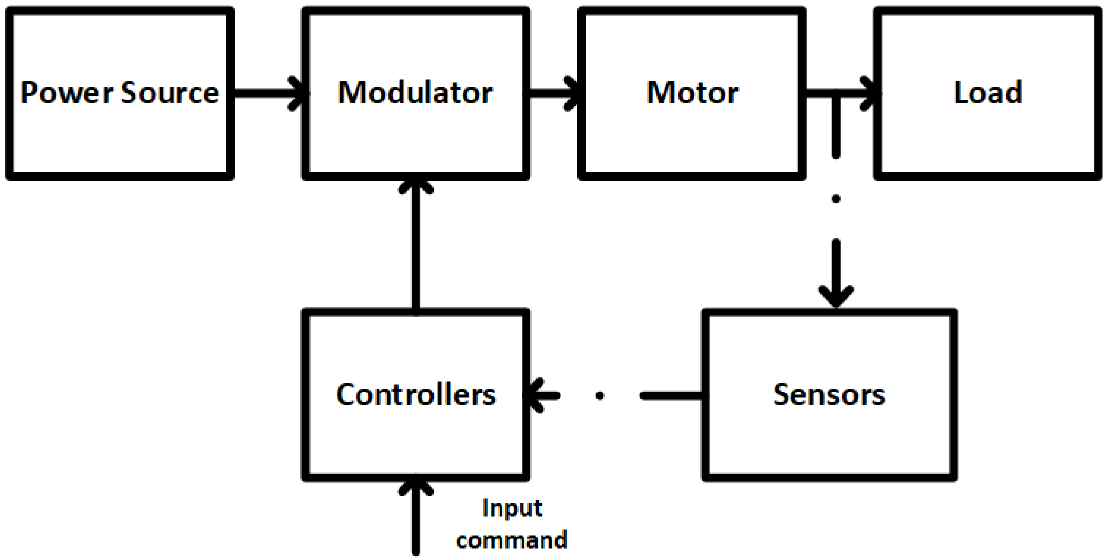

Figure 1: Block Diagram Of Electric Drives

The Power source here could be AC or DC depending on what kind of drives are we talking about. For low power drives, single phase source is applied in AC else 3-phase AC source. 400V supplied motors are low or medium powered motor while 3.3kV to 11kV motors are in category of high powered motors. It powers the drive system and the converter interface with the energy source and it ensures motor with the adaptable frequency current and voltages. The control unit monitors the whole operation of the entire system and it ensures efficiency, stability. It controls the power converter according to the load requirements. The load is chosen by the character of the operation required and Energy source is chosen on the availability. The converter here is to convert the signal waveform so that it can be used by the motor it could be Rectifiers, Inverter, Chopper, Cycloconverters. The motor in the system is selected in view of cost, power and the operation demand. The sensors here could be of Speed sensor using tachometer, torque sensor using magnetoelastic sensor used in race cars, automobile, and aircraft, position sensor using rotary encoder, current and voltage sensor using simple circuits, temperature sensor using thermistor.

Classification of electrical drives:

1. Based on Number of components
   i. Individual Drive: Single motor is used here for various parts.
   ii. Group Drive: Group of motors are connected to one shaft controlled by one motor.
   iii. Multi Motor Drive: The multimotor drive consist of several individual motors which serve to one of many motions or mechanism in some production unit.

Reference Frame

Transformations of voltage, flux linkages and current are done in generalized way. These reference frame are chosen arbitrarily but in some cases, precise reference frames are chosen such as stationary, rotor or synchronous. In 1920 R.H. Park introduced d – q model, in which the variables are transformed with synchronously rotating reference frame fixed to rotor called Park's Transformation. He then, showed that variable inductance can be eliminated caused due to relative motion of circuit and with variable magnetic reluctance.As,

$$\begin{bmatrix} V_{as} \\ V_{bs} \\ V_{cs} \end{bmatrix} = \begin{pmatrix} \cos\theta & \sin\theta & 1 \\ \cos(\theta - 120^0) & \sin(\theta - 120^0) & 1 \\ \cos(\theta + 120^0) & \sin(\theta + 120^0) & 1 \end{pmatrix} \begin{bmatrix} V_{qs} \\ V_{ds} \\ V_{os} \end{bmatrix}$$

Here, V_{os} is zero sequence component, set $\theta = 0$.

We get,

$$V_{qs} = \frac{2}{3} V_{as} - \frac{1}{3} V_{bs} - \frac{1}{3} V_{cs} = V_{as}$$

$$V_{ds} = \frac{-1}{\sqrt{3}} V_{bs} + \frac{1}{\sqrt{3}} V_{cs}$$

2 Literature Review

1. Kioskeridis and N. Margaris, "Loss minimization in scalar-controlled induction motor drives with search controllers," in *IEEE Transactions on Power Electronics*, vol. 11, no. 2, pp. 213-220, Mar 1996. doi: 10.1109/63.486168

In This paper, it is shown that losses can be minimized in scalar control of IM Drives if the input stator current is control variable instead of input power. Stator current is more sensitive to air gap flux variation than input power, adding to the benefit of stator current sensitivity is that it is independent of speed and efficiency. This can be proved by the fact that the minimum of input power is not distinct and the oscillation of air gap flux produces undesirable torque disturbances. It can be overcome by stator current variable.

2. Chun-Chieh Wang and Chih-Hsing Fang, "Sensorless scalar-controlled induction motor drives with modified flux observer," in *IEEE Transactions on Energy Conversion*, vol.18, no. 2, pp. 181-186, June 2003. doi: 10.1109/TEC.2002.805181

It presents simple sensorless scalar control of IM and to overcome the reduction in the voltages of the stator resistance an auto-boost controller was designed and also to maintain constant stator flux amplitude. To overcome the pure integration problem a high pass filter was used in flux observer, and cutoff frequency is proportional to voltage frequency.

Hence the open loop scalar control and proposed sensorless scalar control works witch good speed accuracy in no load and in rated load speed error of 127r/min was detected in open loop but in proposed control speed accuracy remained intact due to the fact that auto-boost controller maintains constant flux, and in open loop at rated load stator flux decreases.

3. H. Luo, Q. Wang, X. Deng and S. Wan, "A Novel V/f Scalar Controlled Induction Motor Drives with Compensation Based on Decoupled Stator Current," *2006 IEEE International Conference on Industrial Technology*, Mumbai, 2006, pp. 1989-1994. doi: 10.1109/ICIT.2006.372556

This method presents low frequency operation of IMD.V/f control is used here for the control scheme using current sensors only. Rotor flux vector is estimated through mathematical modelling of IM to decouple the stator current. Slip frequency and stator voltage drop is compensated based on torque and flux component of stator current in the rotor flux oriented reference frame.

4. A. Smith, S. Gadoue, M. Armstrong and J. Finch, "Improved method for the scalar control of induction motor drives," in *IET Electric Power Applications*, vol. 7, no. 6, pp. 487-498, July 2013. doi: 10.1049/iet-epa.2012.0384

It shows encoder less scalar control of induction motor which can be used where precise control of IM is not required. This method does not require flux estimation and speed sensor. It is better than the open loop V/f control and stator resistance compensation schemes.

This Scalar control scheme provides better response in steady and transient state than any other scalar control methods. It overcomes the problem arises at zero speed, with slip compensation it is better.

5. P. M. Menghal and A. J. Laxmi, "Scalar control of an induction motor using artificial intelligent controller," *2014 International Conference on Power, Automation and Communication(INPAC)*, Amravati, 2014, pp.60-65. doi: 10.1109/INPAC.2014.6981136

It presents Adaptive neuro fuzzy controller which overcomes the problem arises with parameters associated with the membership function and the intuition of the experts.

It shows comparison between PI, Fuzzy controller, ANN, and Adaptive neuro fuzzy controller.

This control is superior as it overcomes parameter variations, load disturbances. The addition of learning algorithm to supply voltage makes it superior than any other control. It decreases rise time with the frequency of sine wave which is changing according to the error from given speed.

6. J. T. Boys and S. J. Walton, "Dynamic flux controlled AC drive," in *IEE Proceedings B - Electric Power Applications*, vol. 137, no. 4, pp. 259-264, July 1990. doi: 10.1049/ip-b.1990.0031

This dynamic method uses Phase Decoupling method to improve transient response. The dynamic flux controller is a frequency decoupled scalar-controlled drive. An input signal frequency is combined with decoupling signal frequency to obtain a signal of the frequency at which motor would operate.

It is faster than other control. It suppresses poles associated with the electrical dynamics of the motor which are irrelevant to the steady state analysis of the machine. The only limitation is accuracy and resolution with which phase can be adjusted.

7. J. T. Boys and S. J. Walton, "Scalar control: an alternative AC drive philosophy," in *IEE Proceedings B - Electric Power Applications*, vol. 135, no. 3, pp. 151-158, May 1988. doi: 10.1049/ip-b:19880017

In this control air gap, flux varies proportionally to load which causes machine current to vary. The controller produces constant slip frequency requiring no external transducers. Torque estimation is done through microprocessors.

It maintains slip speed constant so that speed doesn't vary with load. Critically Motor torque estimation didn't depend upon machine parameters can be calculated from operating point. Boost voltage problem eliminated by varying flux. It is good for speed controlled drives not for position controlled drives.

8. A. Pugachev, "Induction motor temperature influence on scalar control systems efficiency," *2016 2nd International Conference on Industrial Engineering, Applications and Manufacturing (ICIEAM)*, Chelyabinsk, 2016, pp. 1-5. doi: 10.1109/ICIEAM.2016.7911501

It shows efficiency calculation and comparative study of electric drives, how efficiency is affected by the windings temperature and other parameters too. Constant slip modes and minimum stator current are evaluated too.

It concludes that practicality and productivity of application of scalar system with both minimum stator current and constant slip in a small range of winding temperature. As the temperature varies it affects system which can be overcome by the adaptation of slip else the productivity of the system decreases and stator current increases. The advantages of the stator current minimum are, it manages to find the minimum in windings temperature and load torque. Also, hardly any changes in stator voltage presuming any mode in full range of torque.

9. B. K. Bose, "Scalar Decoupled Control of Induction Motor," in IEEE Transactions on Industry Applications, vol. IA-20, no.1, pp.216-225, Jan. 1984. doi: 10.1109/TIA.1984.4504396

A feed forward decoupling network is used to intensify the active response of Induction motor. The adaptive decoupler is implemented through microcontroller.

It shows that parameters are adaptive to different working region provided through lookup table. It has four quadrant operations from zero speed with full torque. Proved on both Voltage and current controlled inverters.

10. I. Boldea, A. Moldovan and L. Tutelea, "Scalar V/f and I-f control of AC motor drives: An overview," *2015 Intl Aegean Conference on Electrical Machines & Power Electronics (ACEMP), 2015 Intl Conference on Optimization of Electrical & Electronic Equipment (OPTIM) & 2015 Intl Symposium on Advanced Electromechanical Motion Systems (ELECTROMOTION)*, Side, 2015, pp. 8-17.doi: 10.1109/OPTIM.2015.7426739

It shows different properties of scalar control of AC motors.

The instabilities in open loop V/f drives can be overcome by compensating slip frequency and boost voltage regulator.

Till DC link current decreases, energy saving can be done by feed forward step variation of voltage. Instability of Open Loop V/f control of PMSMs can be overcome by the frequency modulation and the DC current high pass filtering.

11. Y. N. Dementyev, N. V. Kojain, A. D. Bragin and L. S. Udut, "Control system with sinusoidal PWM three-phase inverter with a frequency scalar control of induction motor," 2015 International Siberian Conference on Control and Communications (SIBCON), Omsk, 2015, pp. 1-6. doi: 10.1109/SIBCON.2015.7147008

It shows comparison between simple sinusoidal PWM control system and PWM control system with additional third harmonics signal and gain modulated control signal. It also shows the result at peak amplitude of control signal and recommends about supply voltage of IM Electric drive.

With addition of 3rd harmonic control signal, it causes fluctuation in speed, torque and motor current, therefore, it is advised to adjust the magnitude of 3rd harmonic as function of frequency.

To compensate heat stress and motor current at high load, motor voltage must be equal to nominal value.

12. N. V. Naik and S. P. Singh, "Improved dynamic performance of direct torque control at low speed over a scalar control," *2013 Annual IEEE India Conference (INDICON)*, Mumbai, 2013, pp. 1-5. doi: 10.1109/INDCON.2013.6725882

It shows the speed control of IM with Direct torque control and scalar control. The Dynamic response of scalar control is poor. SVPWM is used here for better voltage regulation and also it diminishes harmonics in current and voltage of the inverter. Dynamic response is improved in DTC with SVPWM.

Both DTC and Scalar control is compared here which shows Scalar control gives poor transient as well as steady state response. DTC drives with SVPWM gives lesser current distortion under no load, lesser speed regulation, lesser transient time with SVPWM technique.

13. K. Tungpimolrut, Fang-Zheng Peng and T. Fukao, "Robust vector control of induction motor without using stator and rotor circuit time constants," in *IEEE Transactions on Industry Applications*, vol. 30, no. 5, pp. 1241-1246, Sep/Oct 1994. doi: 10.1109/28.315235

The biggest problem in these indirect vector control is parameter variation. It has to be estimated very precisely which is shown here. This method proposes that using terminal voltages and currents we can calculate the regulated energy present in the magnetizing inductance. This could compensate the performance mortification due to parameter difference without usage of significant value of stator and rotor parameter.

It can compensate thermal variations in rotor time constant and stator resistance. It can be rapidly estimated. Therefore, it can be applied without knowing stator resistance and rotor time constant.

14. T. Ohtani, N. Takada, and K. Tanaka, "Vector control of induction motor without shaft encoder," in *IEEE Transactions on Industry Applications*, vol. 28, no. 1, pp. 157-164, Jan/Feb1992. doi: 10.1109/28.120225

It contains no shaft encoder and can control both torque and speed using vector control based on rotor flux speed control, done by rotor flux and torque producing current obtained from stator currents and voltages. The feature of this control deteriorates because of dependence of rotor flux on stator resistance and it is hard to estimate rotor flux at standstill.

Well, this method is free of motor parameters and sensorless, it is proved here that it is independent of stator and rotor resistances and at standstill, it is controllable too.

15. M. Koyama, M. Yano, I. Kamiyama and S. Yano, "Microprocessor-Based Vector Control System for Induction Motor Drives with Rotor Time Constant Identification Function," in *IEEE Transactions on Industry Applications*, vol. IA-22, no. 3, pp. 453-459, May 1986. doi: 10.1109/TIA.1986.4504742

A microprocessor-primarily based high-performance vector manipulate machine for induction motor drives is mentioned. Current control method which is utilized here can function stably even if the saturation of a source voltage occurs. To calculate rotor flux precisely rotor time constant has to be known, this method does not require any extra sensors. It is done by microprocessor.

The DC component control technique consolidated with the variable current and voltage limiter has demonstrated better than the feedback current strategy. Estimation of rotor time constant is also not affected with stator resistance.

16. T. Matsuo and T. A. Lipo, "A Rotor Parameter Identification Scheme for Vector-Controlled Induction Motor Drives," in *IEEE Transactions on Industry Applications*, vol. IA-21, no.3, pp.624-632, May1985.doi: 10.1109/TIA.1985.349719

A rotor parameter identity method for the motive of updating the control gains of an induction motor vector controller is defined. The technique makes use of supply nature of a current regulated PWM inverter through injecting a prescribed negative sequence current signal and then parted into d-q components. By means of injecting the sign at extensively separated frequencies, it is proven that the rotor resistance can be uniquely derived.

Rotor resistance here is calculated by supplying negative sequence current and observing negative sequence voltage.

17. K. Ide, Zhi-Guo Bai, Zi-Jiang Yang and T. Tsuji, "Vector approximation method with parameter adaptation and torque control of CSI-fed induction motor," in *IEEE Transactions on Industry Applications*, vol. 31, no. 4, pp. 830-840, Jul/Aug 1995. doi: 10.1109/28.395293

The idea of the vector approximation approach for CSI-fed induction motors includes vector control and modification of a stator current wave. It implies that space current vector is estimated by 2 feasible current vectors which have 6 directions and DC current control. Not only it decreases pulsating torque but also it controls torque, not done by PWM method. MRAS has been added to the system so that it can work at any constant flux level.

The torque control is done in both transient and steady state by vector estimation with DC current control. The MRAS estimates rotor resistance correctly.

18. Y. Ohdachi, Y. Kawase and M. Hirako, "Dynamic analysis of vector controlled induction motor using finite element method," in IEEE Transactions on Magnetics, vol. 31, no. 3, pp. 1904-1907, May 1995. doi: 10.1109/20.376411

To design IM, the estimation of an exact reaction time needs dynamic investigation utilizing limited component technique where motion of rotor is considered too. New viable technique, which takes the movement condition of the rotor into record and considers the physical movement of the rotor by a programmed subdivision of lattices, has been created. In this paper, a dynamic examination, which is a speed response for the vector controlled IM, is measured using recently created strategy.

19. Fang-Zheng Peng and T. Fukao, "Robust speed identification for speed-sensorless vector control of induction motors," in *IEEE Transactions on Industry Applications*, vol. 30, no.5, pp. 1234-1240, Sep/Oct 1994.doi: 10.1109/28.315234

It describes a new estimation scheme for motor speed from measured terminal voltages and currents for speed-sensorless vector control. It is based on estimation of reactive power of the motor. It is robust to the variation in machine parameter, thermal variations and also independent of integration.

It is different from conventional MRAS as it observes reactive power of Magnetizing inductances. It can equal the control which can be done by shaft position encoders.

20. Ting-Yu Chang and Ching-Tsai Pan, "A practical vector control algorithm for μ-based induction motor drives using a new space vector current controller," in *IEEE Transactions on Industrial Electronics*, vol. 41, no. 1, pp. 97-103, Feb 1994. doi: 10.1109/41.281614

It proposes microprocessor based vector control algorithm. The proposed technique controls the motor torque at once and linearly to achieve instantaneous torque reaction without oscillation. A space vector based current controller is proposed to serve as a brief response torque controller. The current error and its derivative is further applied for reducing the switching frequency. It is very simple to implement.

A larger hysteresis band is employed to check influence of the command signal and Back EMF’s.

21. G. Yang and T. H. Chin, "Adaptive-speed identification scheme for a vector-controlled speed sensorless inverter-induction motor drive," in *IEEE Transactions on Industry Applications*, vol. 29, no. 4, pp. 820-825, Jul/Aug 1993. doi: 10.1109/28.232001

This presents sensorless MRAS system for speed control of IM. Rotor speed is calculated with adaptive system and it is used as feedback signal. The problem arises at low speed, to overcome this problem stator résistance is estimated at the same time. This gives wide bandwidth for speed control and can be stable even at zero speed.

It is proved here that using identifying scheme we can converge rotor speed and flux to the real value. The MRAS control here can estimate stator resistance at any time as the current error includes the resistance value which decreases as value of resistance decreases. The demerit is that rotor resistance and its speed cannot be monitored at the same time with equations so that the error in the observer and the real one will affect accuracy.

22. T. Orlowska-Kowalska and M. Dybkowski, "Stator-Current-Based MRAS Estimator for a Wide Range Speed-Sensorless Induction-Motor Drive," in IEEE Transactions on Industrial Electronics, vol. 57, no. 4, pp. 1296-1308, April 2010.doi: 10.1109/TIE.2009.2031134

It evaluates the MRAS for rotor speed estimation. The effect of change on machine parameter, transfer function and stability of the system is evaluated. The permissible range of machine parameter change is calculated which shows smooth operation of sensorless vector controlled drive with flux and speed estimator.

The rotor flux current based model and stator current estimator gives wide range of speed operation. The application of the actual IM as a reference object in this MRAS estimator concept, in addition to the adjustable stator-current estimator and current-flux version, improves the robustness of the speed reconstruction to motor-parameter uncertainties.

23. M. J. Duran, J. L. Duran, F. Perez and J. Fernandez, "Induction-motor sensorless vector control with online parameter estimation and overcurrent protection," in *IEEE Transactions on Industrial Electronics*, vol. 53, no. 1, pp. 154-161, Feb. 2006. doi: 10.1109/TIE.2005.862302

It shows the complicacy in these drives of low speed operation, parameter variation, and current control. This method is equipped with temperature estimation to compensate steady state, for transient accuracy skin effect estimation is used and overcurrent protection is for current control. These estimation is done using lumped parameter models, this new scheme is capable of account for static friction.

The thermal model could be used for monitoring purpose. The skin effect estimation used here could be for any rotor bar shape and static friction consideration shows better performance in transient state in both conditions, against target speed or load torque changes. The stated protection suggests that it allows transient current over the rated value which improves drive performance. No saturation improves the drive performance.

24. S. A. Odhano, R. Bojoi, A. Boglietti, Ş. G. Roşu and G. Griva, "Maximum Efficiency per Torque Direct Flux Vector Control of Induction Motor Drives," in *IEEE Transactions on Industry Applications*, vol. 51, no. 6, pp. 4415-4424, Nov.-Dec. 2015. doi: 10.1109/TIA.2015.2448682

It presents efficient operation for a demand torque using direct regulation of stator flux according to torque using stator flux reference. With no load test and short-circuited test data stator flux map is calculated for maximum performance per torque. An iron loss model is explained here for the calibration of the machine loss model and additionally for on-line monitoring of the iron losses for the duration of motor operation.

The magnetic saturation is considered for MEPT operation and iron losses can be predicted with iron loss model using stator flux and frequency.

25. M. Mengoni, L. Zarri, A. Tani, G. Serra and D. Casadei, "Stator Flux Vector Control of Induction Motor Drive in the Field Weakening Region," in *IEEE Transactions on Power Electronics*, vol. 23, no. 2, pp. 941-949, March 2008. doi: 10.1109/TPEL.2007.915636

The control variables used in this scheme is Stator flux components. The proposed scheme permits the motor to make the most the most torque capability inside the whole speed range and shows a discounted dependence at the motor parameters. The control algorithm decreases the d component of the stator flux as quickly because the voltage similar to the maximum torque exceeds the available voltage, also gives easy transition into and out of the field weakening region.

It gives less dependence on machine parameters, no need of estimation of base speed and fast torque response.

26. H. Tajima, G. Guidi and H. Umida, "Consideration about problems and solutions of speed estimation method and parameter tuning for speed-sensorless vector control of induction motor drives," in *IEEE Transactions on Industry Applications*, vol. 38, no. 5, pp1282-1289, Sep/Oct2002.doi: 10.1109/TIA.2002.802893

An established model-based speed-sensorless field-situated control technique for a universally useful IM is considered. Enhanced renditions of both speed and stator resistance online estimators are given the point of broadening control capacities down to zero speed. Just electrical stator estimations are required, making the technique reasonable for broadly useful inverter applications. Online rotor resistance tuning is likewise included to compensate thermal effect.

Using online tuning of machine parameters and by modifying the control algorithm, current-voltage based sensorless drives operation can be enhanced. Zero speed can be controlled too.

27. K. Matsuse, S. Taniguchi, T. Yoshizumi and K. Namiki, "A speed-sensorless vector control of induction motor operating at high efficiency taking core loss into account," in *IEEE Transactions on Industry Applications*, vol. 37, no. 2, pp. 548-558, Mar/Apr 2001. doi: 10.1109/28.913721

Core loss is considered here in the rotor flux observer. It causes the rotor flux and current to obstruct each other which causes O/P torque with reference torque.

Stator core loss has a significant unfavorable impact in conventional field-oriented control. It affects the output torque, rotor flux, and current. with the intention to eliminate this impact, stator current has to be compensated. Then using reference torque, rotor flux is calculated which grasps maximum efficiency. To maintain this efficiency without deterioration of the dynamic reaction, fast control of rotor flux was proposed that employs deadbeat rotor flux control.

28. N. Kobayashi, F. P. Wijaya, K. Kondo and O. Yamazaki, "Induction Motor Speed-Sensorless Vector Control Using Mechanical Simulator and Disturbance Torque Compensation," in *IEEE Transactions on Industry Applications*, vol. 52, no. 3, pp. 2323-2331, May-June 2016. doi: 10.1109/TIA.2016.2524440

A mechanical simulator technique is proposed as a speed-sensorless vector control in an extremely-low speed region. In this process frequency error occurs when torque error occurs. It causes increment in slip frequency which decrement rotor flux and torque. To overcome this problem the current error is compensated by PI controller which compensates the disturbance torque in the mechanical simulator.

Disturbance Torque is estimated here with the help of torque current. Therefore, we can get desired torque even when the disturbance torque is not from the set value. Which proves working at very low speed. Useful in railway vehicles.

29. M. Comanescu and L. Xu, "Sliding-mode MRAS speed estimators for sensorless vector control of induction Machine," in *IEEE Transactions on Industrial Electronics*, vol. 53, no. 1, pp. 146-153, Feb. 2006.doi: 10.1109/TIE.2005.862303

It presents Sliding Mode Observer in MRAS system for speed calculation. This method uses the flux calculation from voltage model and make Sliding mode flux observer that permits speed calculation. Dynamics and stability is compared with the conventional MRAS observer. It is easy to tune. As in conventional MRAS scheme, it is based on calculations that do not consider RHS zeroes and underdamped poles.

This method only requires sliding mode gain calculation. The speed here is calculated by any of the algebraic calculations or low pass filtering methods. This method does not show any speed dip or damped responses caused by RHS zeroes.

30. S. M. Gadoue, D. Giaouris and J. W. Finch, "MRAS Sensorless Vector Control of an Induction Motor Using New Sliding-Mode and Fuzzy-Logic Adaptation Mechanisms," in *IEEE Transactions on Energy Conversion*, vol. 25, no. 2, pp. 394-402, June 2010. doi: 10.1109/TEC.2009.2036445

In this paper to replace the PI controller, two adaptation schemes are proposed based on rotor flux. The first one is on sliding mode theory and the other one is based on fuzzy logic. Speed adaptation is based on Lyapunov theory. Comparison is done between conventional scheme and the new ones in open and closed loop both.

Chapter 1 These new schemes show better transient performance and better disturbance torque rejection in both open and closed loop. Need of LPF of estimated speed in sliding mode controller makes it slower than Fuzzy logic strategy.

3 Scalar Control of Induction Motor

These variable speed drives with IM are low performance drives in which magnitude and frequency of stator voltage or current are adjusted. It helps in control of the constant speed or torque of the motor, and magnetic field remains constant. This form of control is called scalar control, the stator voltage or current which has to be controlled are assumed sinusoidal. Only magnitude and frequency are controlled nit the phasor likely in vector control. The vector control schemes are based on the field orientation principle which adjusts the magnitude and phasor of the motor quantity. In vector-controlled drive systems, torque and magnetic field of the motor are regulated, each under transient and steady state conditions. Vector control systems are way complex than scalar control systems, however, the superiority of the vector control in dynamic conditions is unquestionable.

Scalar control uses constant Volts/Hertz (V/f), due to the fact that value of stator voltage is regulated proportionally to the frequency. It is to preserve constant stator flux. V/f method controls the speed of the rotating magnetic field of the stator by ch1anging the supply frequency. The torque produced is dependent on the slip velocity, i.e., the difference in synchronous speed and the rotor. The control device is simple, on the grounds that simplest speed feedback is needed. Scalar control also consists of torque control in which magnitude and frequency of stator current is maintained in such a way that magnetic field remains constant.

3.1 Γ Equivalent Circuit of an IM

It is a per phase representation of steady state model of IM. The parameters are calculated from Blocked Rotor tests and no-load tests. No Load test is done for Mutual Inductance, L_m. Stator and Rotor Resistance ($R_s + R_r$) and stator and rotor leakage reactance can be calculated from Blocked Rotor test. Stator resistance can be calculated easily as it is easily accessible.

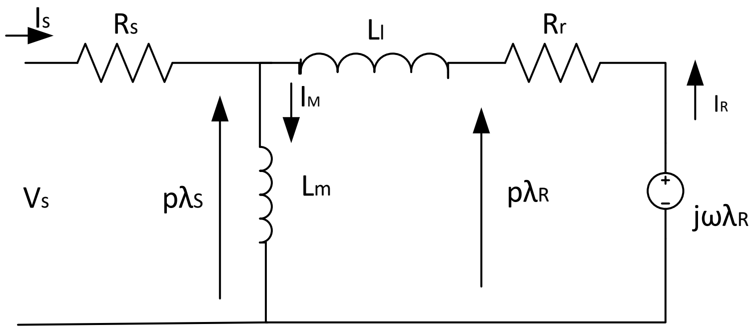

Figure 2 : Γ Steady State Equivalent Circuit of An Induction Motor

The Torque and Voltage equations of this circuit are:

$$T = 1.5 P L_M \operatorname{Im}(I_s I_R^*) \tag{3.1}$$

$$\begin{pmatrix} 0 \\ V_s \end{pmatrix} = \begin{pmatrix} j\omega L_m & \dfrac{R_R}{s} + j\omega L_R \\ R_s + j\omega L_m & j\omega L_m \end{pmatrix} \begin{pmatrix} I_s \\ I_R \end{pmatrix} \tag{3.2}$$

All the four parameters can be calculated from the tests and stator resistance R_s by direct measurement.

3.2 Principles of the Volts/Hertz Control

Under V/f control, waveform of stator voltage is considered sinusoidal with adjustable magnitude and frequency. Using steady-state Γ equivalent circuit we can analyze the IM operation.

$$P_E = 3(I^2 R)_R \frac{\omega}{\omega_r} \tag{3.3}$$

$$P_M = P_E - 3(I^2 R)_R \tag{3.4}$$

$$T = \frac{P_M}{\omega_M} \tag{3.5}$$

$$\omega_M = \frac{2}{P}(\omega - \omega_r) \tag{3.6}$$

Where I_R can be calculated from Γ equivalent circuit:

$$I_R = \left| \frac{-j\omega\lambda_s}{R_R \frac{\omega}{\omega_r} + j\omega L_l} \right|$$

$$= \frac{\lambda_s}{\sqrt{\frac{R_R^2}{\omega_r^2} + L_l^2}}$$

$$= \frac{\lambda_s^2}{R_R^2} \frac{\omega_r^2}{(\tau\omega_r)^2 + 1} \tag{3.7}$$

Where, $\tau = \frac{L_l}{R_R}$

Substituting (3.7) in (3.3),(3.4) we get,

$$P_E = 3\frac{\lambda_s^2}{R_R} \frac{\omega\omega_r}{(\tau\omega_r)^2 + 1}$$

$$P_M = 3\frac{\lambda_s^2}{R_R} \frac{\omega_r(\omega - \omega_r)}{(\tau\omega_r)^2 + 1}$$

Therefore,

$$\omega_0 = 0, T = 0, P_M = 0,$$

From above equations, we conclude:

i. If $s < 0$ then $T < 0$, $P_M < 0, P_E < 0$ it represents retardation in which mechanical energy is being consumed by the machine and Electrical Energy is delivered by the machine. i.e. machine is operating as a generator as its rotor is moving at super synchronous speed.

ii. If $s = 0$, then $\omega_0 = \omega, T = 0, P_M = 0 \,\&\, P_E = 0$, i.e. no-load condition at synchronous speed.

iii. If $0 < s < 1$, then $0 < \omega_0 < \omega, T > 0, P_M > 0,$ and $P_E > 0$ Motoring Region.

iv. If $s = 1$ then $\omega_0 = 0, T = 0, P_M = 0,$ in this condition rotor is still and starting torque is

$$T_{st} = 1.5P \frac{\lambda^2}{R_R} \frac{\omega}{\tau^2 \omega^2 + 1} \tag{3.8}$$

v. If $s > 1$, then $\omega_o < 0, T > 0, P_M < 0, and\ P_E > 0$ shows developed torque is working against motion of rotor, and both Electrical and Mechanical power is consumed,acting as brake.

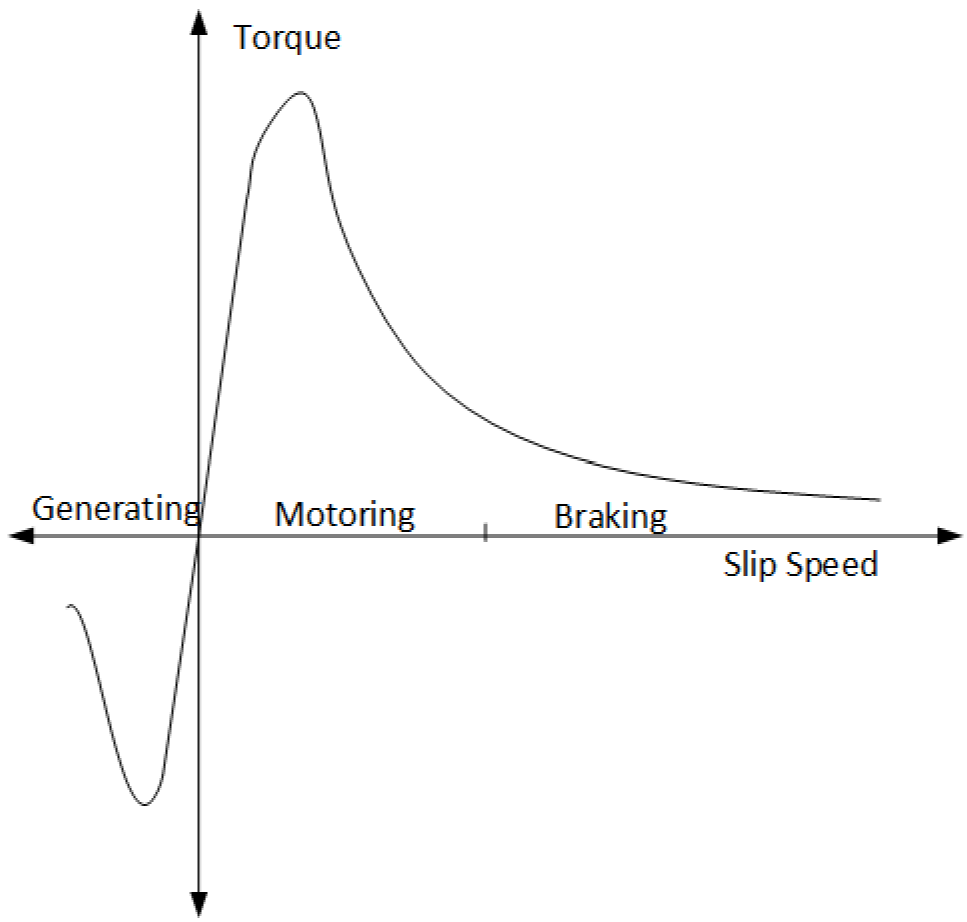

Figure 3: Torque vs Slip Speed Characteristics of IM with Constant Stator Flux

The machine analysed here is a motor so when machine delivers power, mechanical power is presumed positive and when it consumes power then electrical energy is presumed positive.

It states that magnitude V_s is controlled so that the magnitude (stator flux) λ_s is remained constant, irrespective of changes in frequency then the torque depends on slip speed only. And it follows Eq. (3.6) which shows motor speed is a linearly dependent on frequency.

$$\omega_M = \frac{2}{P}(2\pi f - \omega_r) \tag{3.9}$$

If stator voltage drop across R_s is negligible in comparison with V_s then

$$V_s \approx \omega\lambda_s \tag{3.10}$$

Which says

$$\lambda_s \propto \frac{V_s}{f} \tag{3.11}$$

Eq. (3.11) shows "Constant V/f control" i.e. if ratio V/f remains constant then λ_s remains constant which shows torque is independent of the supply frequency but it has limitations too the frequency when lowered the voltage is lowered too which makes the stator resistance voltage drop visible and cannot be neglected then for compensation voltage has to be increased.

And, when the frequency is increased above rated value then the insulation problem comes into effect because the voltage cannot be increased above rated value and the stator flux decreases with rise in frequency, and torque decreases.

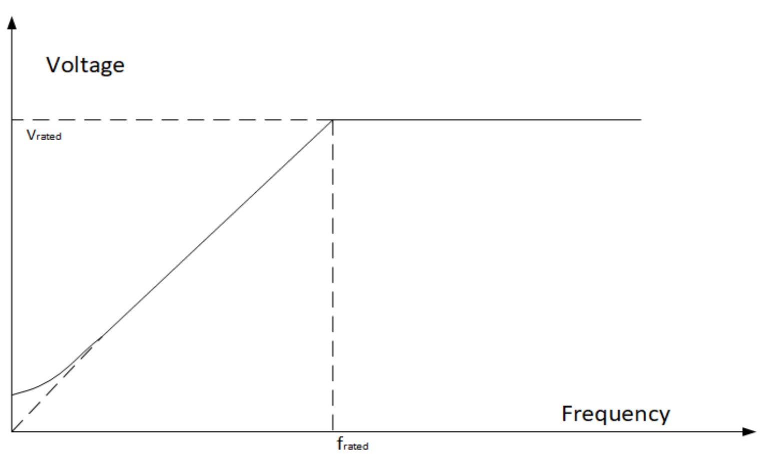

Figure 4: V vs f of Constant V / f Control

3.3 Speed Control System

In this control system, a speed sensor is fitted on the motor shaft which feedbacks motor speed signal. Which is then compared with the reference speed signal, ω_{Mref} which results in speed error $\delta\omega_{M}$ is then given to a controller which outputs reference slip signal ω_{slref} . The controller suppresses the speed to a value relative to the peak torque. After that motor is forced to work in the operating region i.e. within the peaks of the developed torque. A load change in an uncontrolled drive will result in a change in speed such that the motor torque levels with a new load torque and then system settles to a new stable operating point. Which shows that, it can provide fast response to change the reference speed if the torque limit is set in the vicinity of the peak torque.

Addition of ω_{slref} and ω_{M} gives reference synchronous speed signal ω_{synref} proportional to supply frequency ω (in radian). Similarly, a voltage controller generates a reference stator voltage signal which is to be supplied by the inverter.

3.4 Mechanical Characteristics of an Induction Motor

It is defined in context of "Constant V / f control". It is represented by a torque as a function of speed. It is shown in the Figure-5 with different supply frequency. As it is mentioned earlier that below rated frequency maximum torque remains constant and it starts decreasing with increase in frequency above rated value. Where ""Constant V / f control"" doesn't exist.

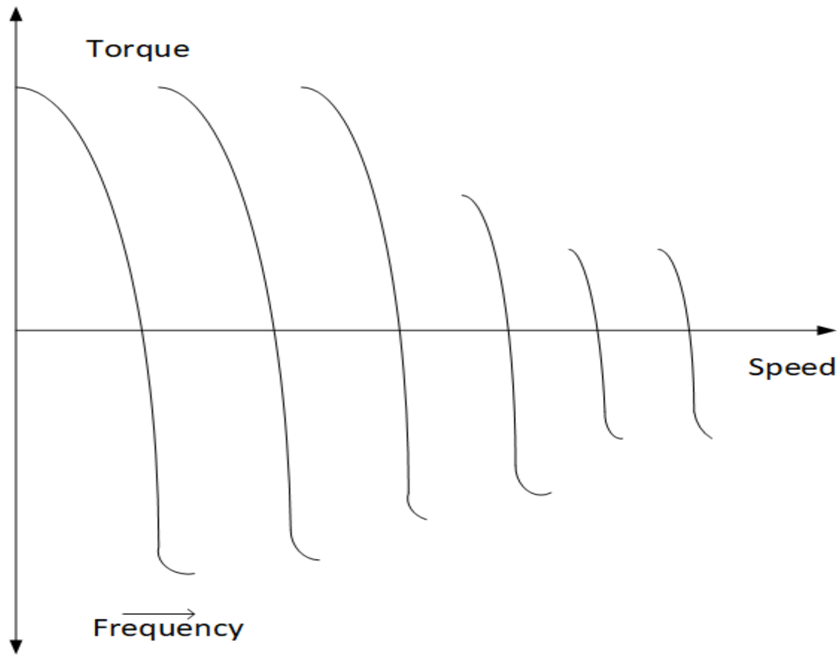

Figure 5: Mechanical Characteristics of IM in Constant V / f control

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3.5 Γ' Equivalent Circuit of an IM

This circuit is drawn here for the torque control system and τ' is introduced for $\Gamma \rightarrow \Gamma'$ transformation.

Using,

$$\tau' = \frac{L_M}{L_R} \tag{3.12}$$

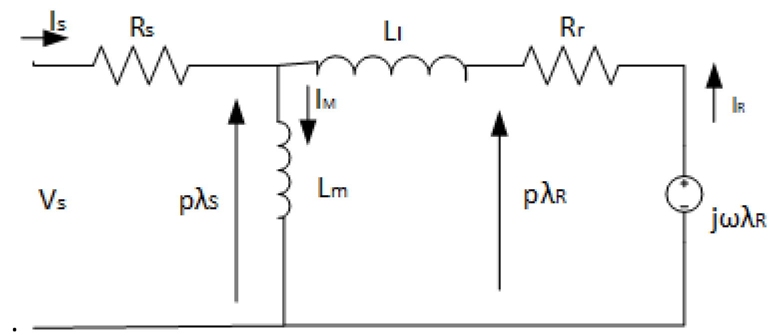

Figure 6: *Dynamic* Γ' *Equivalent Circuit Of Induction Motor*

Voltage and torque equation are as:

$$\begin{pmatrix} 0 \\ V_s \end{pmatrix} = \begin{pmatrix} j\omega_0 L'_M & j\omega_0 L'_M + \dfrac{R'_R}{s} \\ R_s + j\omega L'_s & j\omega_0 L'_M \end{pmatrix} \begin{pmatrix} I_s \\ I'_R \end{pmatrix} \qquad (3.13)$$

$$T = 1.5 P L'_M I_M (I_s {I^*}_R) \qquad (3.14)$$

3.6 Principles of Torque Control

$$T \;\alpha\; \frac{P_{Mech.}}{\omega_o}$$

Therefore,

$$T = \frac{P}{2}\frac{3I'^2_R R'_R \frac{\omega_o}{\omega_R}}{\omega_o}$$

$$T = 3\frac{P}{2}\frac{I'^2_R R'_R \omega_o}{\omega_R} \tag{3.15}$$

Where,

$$I'_R = \frac{I'_M \omega_R L'_M}{R'_R}$$

Above equation says that torque could be directly proportional to the rotor current if the magnitude of the magnetizing current is maintained constant. And Also, the magnitude of the rotor flux, λ_R is proportional to the magnetizing current, I'_M .The Stator current could be shown as the difference of Magnetizing current and Rotor current or say torque producing current $-I'_R$. When these stator current components are individually controlled then the motor becomes linear at constant stator current. For this, a current controller is needed which can synthesize flux and torque producing current from stator current. Input to this current controller would be reference magnitude stator current and frequency in radian.

$$I_{sref} = \sqrt{I^2_{s\theta ref} + I^2_{sTref}} \tag{3.16}$$

$$\omega_{Rref} = \frac{R_R}{L_R} \frac{I_{s\theta ref}}{I_{sref}} \tag{3.17}$$

3.7 Scalar Torque Control System

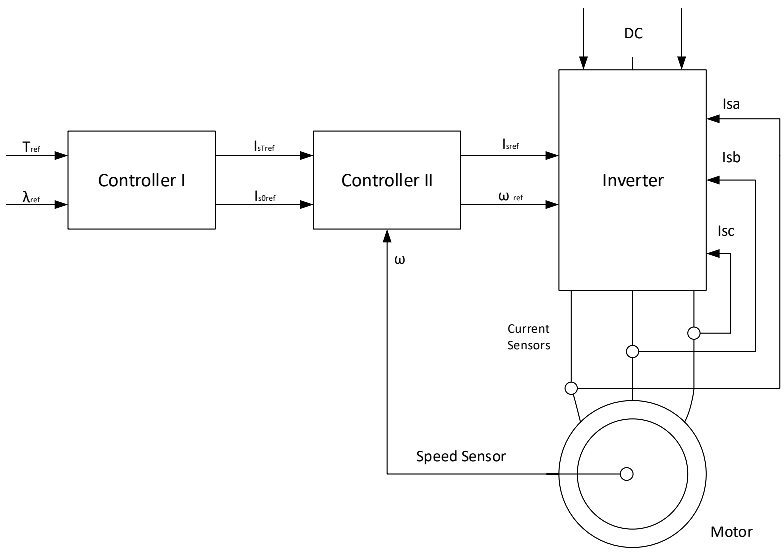

Figure 7: Block Diagram Of Scalar Control (Torque)

The first controller in the *Figure 7* computes reference torque and rotor flux signal to produce torque producing current and flux producing current respectively.

Torque producing current,
$$I_{sTref} = \frac{T}{1.5P\lambda_{Rref}} \tag{3.18}$$

Flux producing torque current:
$$I_{s\theta ref} = \frac{\lambda_{Rref}}{L_M} \tag{3.19}$$

And the second controller produces stator current reference and frequency reference from the output of the first controller i.e. I_{sTref} and $I_{s\theta ref}$ and ω_M signal too that is obtained from speed sensor.

Both control in practical is done by single microprocessor.

Here, the inverter must be current controlled which means there must be current sensors which feedbacks all the 3-phase stator current and send it for further calculation for reference current signal which ensures these current to be controlled in correspondence with reference signals.

The developed torque is limited due to the fact that increasing voltage will cause insulation breakdown and increasing current would cause overheating which leads to damage winding. For transient time only the torque above rated could be permitted.

Permissible Rotor flux and torque at various speed defines safe operating area of motor.

4 Vector Control of Induction Motor

Steady state performance of an inverter fed IM drives is comparable with Self Excited DC motor. In comparison to DC machines dynamic performance of IM with smooth controls is sluggish. This drawback has constrained the utility of induction motor drives in industries requiring excessive overall performance, such as tool drives. The vector control or FOC of AC machines is advanced to compete with the similar performance of a separately excited DC motor. The goal of vector control is to obtain advanced performance despite changes in speed and torque. In separately excited DC motor the field and armature current are completely independent of each other which applies on their fluxes too. Which clearly shows that these two circuits are not magnetically coupled to each other so that the control is easy in both the circuit. Thus, armature current and torque can be modified keeping field flux constant. In AC Machines, too current and flux are responsible to produce torque. But in induction motor, only the stator is fed electrically so that flux and torque producing current are not so easily separable. Basic principle behind vector control is to completely separate these two flux and torque producing current. This can be achieved by controlling stator currents magnitude, frequency and phase too using inverter control. As the name suggests “vector control” that is magnitude and phase both has to be controlled.

4.1 Space Vectors and Transformations

This vector control is all based on space vectors.

Stator current space vector can be shown as:

$$\vec{I}_s = I_{as} e^{j0} + I_{bs} e^{j(\frac{2\pi}{3})} + I_{cs} e^{j(\frac{4\pi}{3})} \quad (4.1)$$

Eq. (4.1) shows constant magnitude of stator current and is rotating at synchronous speed. The 'd-q-0' model can be shown as:

$$\begin{pmatrix} I_{ds} \\ I_{qs} \\ I_0 \end{pmatrix} = \begin{pmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & -\frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} \\ 1 & 1 & 1 \end{pmatrix} \begin{pmatrix} I_{as} \\ I_{bs} \\ I_{cs} \end{pmatrix} \quad (4.2)$$

Here we used balanced phase current:

$$I_0 = I_{as} + I_{bs} + I_{cs} = 0 \quad (4.3)$$

Similarly Eq. (4.1) (4.2) (4.3) can be written for stator voltages (v_s) too.

The resultant voltage v_s in rectangular coordinates can be expressed as:

$$\vec{V}_s = V_s \cos\theta + jV_s \sin\theta \quad (4.4)$$

θ is the angle with the ref. axis.

$$V_s = V_{ds} + jV_{qs} \qquad (4.5)$$

In eq. (4.5) d axis is taken as reference.

Stator flux linkage for space vector can be written as:

$$\vec{\lambda}_s = \frac{2}{3}(\lambda_{as} + \lambda_{bs} e^{j\frac{2\pi}{3}} + \lambda_{cs} e^{j\frac{4\pi}{3}}) \qquad (4.6)$$

In 'd-q' axis it can be shown as:

$$\vec{\lambda}_s = \lambda_{ds} + j\lambda_{qs} \qquad (4.7)$$

All the above equations derived in stator reference frame for stator currents, voltages & flux linkages. As we know in IM rotor windings are short circuited, so that voltage across it is 0 and current flows in both stator and rotor

In rotor reference frame:

$$\vec{I}_{rr} = I_{rrd} + jI_{rrq} \qquad (4.8)$$

If both the stator and rotor reference frame are aligned then i_r will be equal and if there is an angle θ between them then rotor current represented in stator ref. frame:

$$\begin{pmatrix} I_{rsd} \\ I_{rsq} \end{pmatrix} = \begin{pmatrix} \cos\theta & -\sin\theta \\ \sin\theta & \cos\theta \end{pmatrix} \begin{pmatrix} I_{rrd} \\ I_{rrq} \end{pmatrix} \qquad (4.9)$$

Stator and rotor windings are coupled with mutual flux, as both windings carry currents. This mutual flux linkage works as a channel between stator and rotor for power transfer and produces torque. It will be maximum L_M when both the stator and rotor reference axis coincides with each other. This mutual inductance changes with rotor angle θ can be described as $L_M \cos\theta$ in this case there will be mutual inductance between d-q axis and it will be decoupled.

4.2 Principles of Vector Control

Vector control is done by controlling the torque and field producing component of stator current individually through some synchronized change in the source voltage magnitude, frequency, and phase. The IM with vector control is similar to a dc device in overall performance. This field oriented control states that the stator current must be decoupled in torque and flux producing component. The direct axis component is the flux producing component aligned with rotor flux vector and the quadrature axis component is the torque producing current. Synchronously rotating d-q reference frame is used because in this frame steady state AC quantities behaves like DC quantities. The torque of IM is proportional to $i_{ds}\, and\, i_{qs}$, the field and torque producing current respectively. The DC like feature is possible if i_{qs} is responsible for the production of torque and i_{qs} is for production of flux and aligned with flux direction. For vector control, we can choose any of the reference frame from Rotor flux, Stator Flux, and Air Gap Flux. Two of these flux stator and field flux rotates at synchronous speed but rotor flux rotates with slips peed.

This control can be implemented with respect to any of the phasors of these fluxes by aligning d-axis with the same flux. Out of these flux linkage which is attached to the Total Rotor flux linkage phasor is the preferred one. It’s a rotating reference frame synchronized with flux linkage of rotor. Here the reference frame is a field frame which justifies the “Field Oriented Control”.

Inverter which has vector controller have two command inputs I_{dsref} and I_{qsref} . These are d and q axis ref. components of stator current in a synchronously rotating ref. frame. The 3 phase stator currents $I_{as}, I_{bs}, and\, I_{cs}$ are converted into the d-q components I_{dsref} and I_{qsref} and then these currents are converted into field reference frame which is having an angle θ_f with respect to stationary reference frame. As rotor flux linkage vector rotates this field angle varies too. After transformation of these currents they are then compared with reference current to produce error signal which is then amplified for controlling flux and torque. Using inverse transformation these currents are then converted into stator frame. As soon as the direct and quadrature axis components are known it can be transformed using 2/3 transformation to obtain a-b-c components which are then compared with actual motor current to produce switching current of the inverter. This decoupled control results similar to the DC motors dynamic response. It is very likely to use the current I_{sd} with any of the axes with rotor, stator or air gap flux for vector control. Rotor flux axis was chosen for the fact that there is natural decoupling. Two types of vector control are present:

i) Direct Vector Control (Feedback)
ii) Indirect Vector Control (Feedforward)

Sensors or flux model are used for field calculation in direct vector control. In indirect vector control speed of flux linkage space vector is calculated with respect to rotor speed and then integrated to know Rotor’s position. It is then summed up with angle moved by rotor to obtain rotor angle. The flux position obtained here is by indirect means that is why it is called Indirect Vector Control.

4.3 Direct Vector Control

In this technique, rotor flux vector is calculated directly with the help of direct flux measurement. The sensors that are used here for flux measurement are disadvantages too. Complexity of these sensors and costs are the major drawback of this control system. Output of these sensors are then used in feedback loop, alternatively; the flux models can also be used instead of flux sensor for which stator voltages and can be used as feedback signal and it outputs rotor flux angle.

Fig.8 shows direct vector control system. ω_{Rref} is the reference speed. Rotor speed ω_R is compared with this reference speed and error is then fed to a PI controller. Input reference of the torque loop is same as the output of the speed controller. Torque error is then computed with comparison of the reference and calculated torque. Torque producing stator component current, I_{sTref} is computed from this error. Similarly, Reference rotor flux linkage λ_{Rref} is obtained from rotor speed. It ensures constant and smooth operation of motor. To equal the Torque component current rotor flux linkages are reduced and therefore the electromagnetic torque is reduced too, and due to the field weakening effect speed is increased & constant power is received. From the comparison of rotor flux linkages, reference flux producing current I_{fref}. Addition of phasors of I_{fref} *and* I_{Tref} gives I_{sref} and angle between them gives θ_{Tref}. Knowing of stator current reference (I_{sref}), torque angle reference (θ_{Tref}), stator phase current reference ($I_{aref}, I_{bref}, and\ I_{cref}$). With all these known components, it is possible with PWM inverters to comply with those ref. currents. Consequently, in this control, the feedback variables $T_E, \theta_{fref}, \lambda_R$ are required. These variables can be calculated from the flux and torque block. From the measured components input to the processor could be:

i. Stator currents or L-L Voltages

ii. Induced EMF (From Sensors) and Stator Currents

It computes $T_E, \theta_{fref}, and\ \lambda_R$.

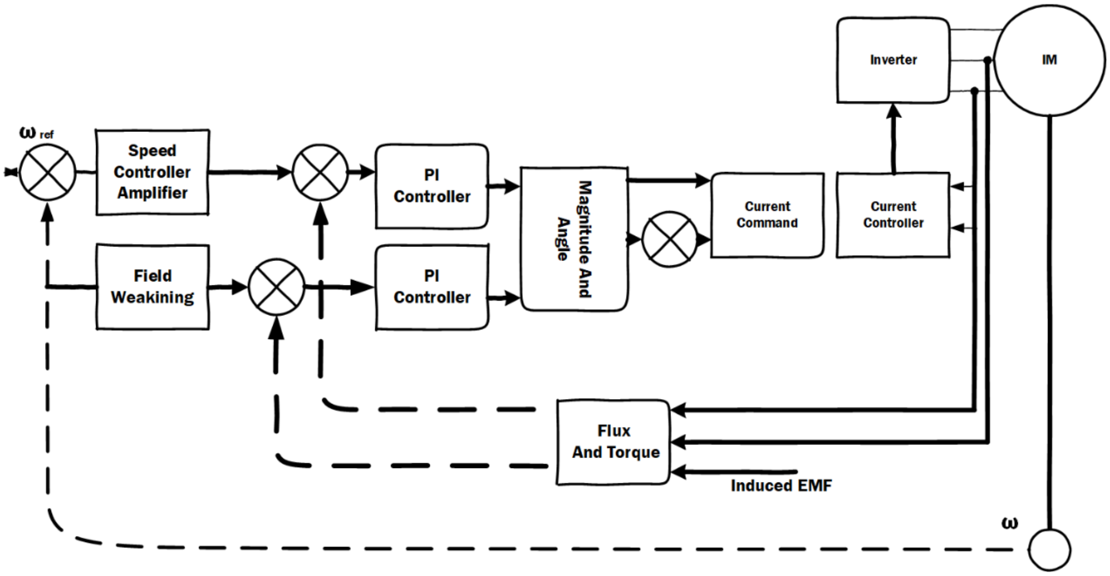

Figure 8: Direct Vector Control

4.4 Control Using Line Currents and Voltages

$T_E, \theta_{fref}, and\ \lambda_R$ these three are calculated using any of the rotor or stator flux linkages which can be computed with terminal voltages and currents. Here we are using rotor flux linkage method.

From 2 to 3 phase transformations:

$$I_{qs} = \frac{1}{\sqrt{3}}(-I_{Bs} + I_{Cs}) \tag{4.10}$$

$$I_{ds} = I_{As} \tag{4.11}$$

Similarly, for Voltages:

$$V_{qs} = \frac{1}{\sqrt{3}}(-V_{Bs} + V_{Cs}) \tag{4.12}$$

$$V_{ds} = \frac{2}{3}V_{As} - \frac{1}{3}V_{Bs} + \frac{1}{3}V_{Cs} \tag{4.13}$$

Rotor Currents Equations:

$$V_{ds} = R_s I_{ds} + j\omega_s L_s I_{ds} + j\omega_s L_M I_{dr} \tag{4.14}$$

$$V_{qs} = R_s I_{qs} + j\omega_s L_s I_{qs} + j\omega_s L_M I_{qr} \tag{4.15}$$

$T_E, \theta_{fref}, and\ \lambda_R$ can be calculated from known stator and rotor currents.

$$T_E = 1.5\frac{P}{2}L_M(I_{qs}I_{dr} - I_{ds}I_{qr}) \tag{4.16}$$

And

$$\lambda_{dr} = L_R I_{qr} + L_M I_{qs} \tag{4.17}$$

$$\lambda_{qr} = L_R I_{dr} + L_M I_{ds} \tag{4.18}$$

$$\tan\theta = \frac{\lambda_{dr}}{\lambda_{qr}} \tag{4.19}$$

From these equations, we can see that motor control totally depends upon Motor Parameters Inductance and Resistance which finally depends upon Magnetic Saturation Level.

4.5 Control using Induced EMF and Current

To overcome the variations in stator resistance we could use induced EMF's. Sensors are placed to obtain this flux.

d and q axis, stator flux linkages are:

$$\lambda_{ds} = L_S I_{ds} + L_M I_{dr} \tag{4.20}$$

$$\lambda_{qs} = L_S I_{qs} + L_M I_{qr} \tag{4.21}$$

$$\tan\theta = \frac{\lambda_{qs}}{\lambda_{ds}} \tag{4.22}$$

And Torque,

$$T_E = 1.5\frac{P}{2}(I_{qs}\lambda_{ds} - I_{ds}\lambda_{qs}) \tag{4.23}$$

In case of rotor flux linkages, rotor currents are:

$$I_{dr} = \frac{1}{L_M}\int (E_{ds}dt - L_s I_{ds}) \tag{4.24}$$

$$T_E = 1.5\frac{P}{2}L_M(I_{qs}I_{dr} - I_{ds}I_{qr}) \tag{4.25}$$

$$I_{qr} = \frac{1}{L_M}\int (E_{qs}dt - L_s I_{qs}) \tag{4.26}$$

Rotor flux linkages can be computed using sensed stator currents and rotor currents:

$$\lambda_{dr} = L_r I_{dr} + L_M I_{ds} \tag{4.27}$$

$$\lambda_{qr} = L_r I_{qr} + L_M I_{qs} \tag{4.28}$$

$$\tan\theta = \frac{\lambda_{qr}}{\lambda_{dr}} \tag{4.29}$$

$$T_E = 1.5\frac{P}{2} L_M (I_{qs} I_{dr} - I_{ds} I_{qr}) \tag{4.30}$$

At low speed, the induced EMF is very small which is a drawback of using induced EMF. So, the calculations of rotor flux linkages and rotor currents results in huge error. And these sensor results in complexity of the system.

4.6 Indirect Vector Control

The position of rotor field is obtained using integration of relative speed of rotor flux with respect to rotor. It is then summed with rotor angle movement to get θ_R . This method is much known in Industrial process. Phasor diagram as in Fig. 9. It explains Indirect vector control. Rotor field reference is having an angle θ_F with stator reference frame. It is time variable as rotor flux linkage vector λ_R rotates. Rotor and Rotor field speed, ω_R and ω_F with respect to rotor. The synchronous speed

$$\omega_s = \int (\omega_R + \omega_F) dt$$

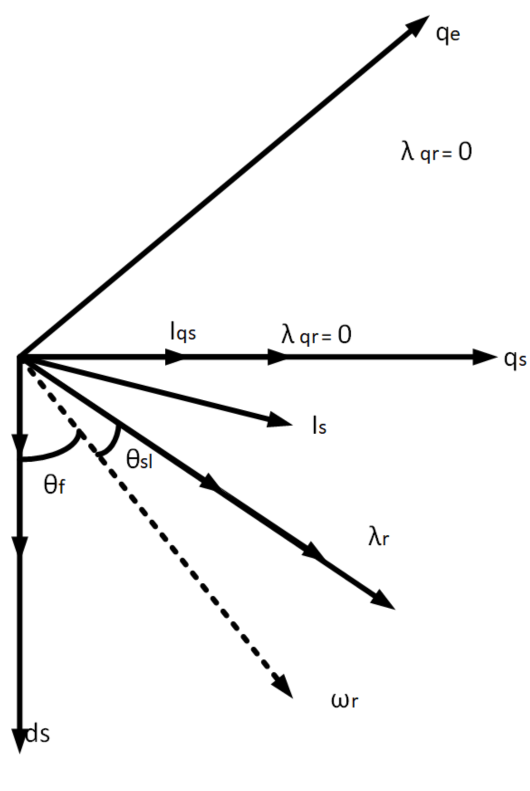

Figure 9: Phasor Diagram of Indirect Vector Control

For indirect vector control, IM is indicated in synchronously rotating reference frame and is supplied with CSI so that stator current can be modified directly. From d-q model we can derive control equations.

In synchronous reference frame:

$$\frac{d\lambda_{dR}}{dt} + R_R I_{dR} - \omega_F \lambda_{qR} = 0 \tag{4.31}$$

$$\frac{d\lambda_{qR}}{dt} + R_R I_{qR} - \omega_F \lambda_{dR} = 0 \tag{4.32}$$

Here,

$$\lambda_{dR} = L_R I_{dR} + L_M I_{dS} \tag{4.33}$$

$$\lambda_{qR} = L_R I_{qR} + L_M I_{qS} \tag{4.34}$$

Substituting, Rotor Currents $I_{dR}\, and\, I_{qR}$ from Eq. (4.33),(4.34) in Eq. (4.31),(4.32) we get:

$$\frac{d\lambda_{dR}}{dt} + \frac{R_R}{L_R}\lambda_{dR} - \frac{L_M}{L_R}R_R I_{dS} - \omega_F \lambda_{qR} = 0 \tag{4.35}$$

$$\frac{d\lambda_{qR}}{dt} + \frac{R_R}{L_R}\lambda_{qR} - \frac{L_M}{L_R}R_R I_{qS} - \omega_F \lambda_{dR} = 0 \tag{4.36}$$

The rotor flux linkage is inferred to be in phase with d-axis so that $\lambda_{qR} = 0$ and $\frac{d\lambda_{qR}}{dt} = 0$

Therefore, total rotor flux

After substituting these we get Field producing component ($I_f \text{ or } I_{ds}$) and torque producing component ($I_T \text{ or } I_{qs}$).

$$I_f = \frac{1}{L_M}(\lambda_R + \frac{L_R}{R_R}\frac{d\lambda_R}{dt}) \tag{4.37}$$

$$I_T = \omega_F \frac{L_R}{L_M}\frac{\lambda_R}{R_R} \tag{4.38}$$

And Torque,

$$T_E = 1.5\frac{P}{2}\frac{L_M}{L_R}\lambda_R I_{qs} \tag{4.39}$$

From Eq.(4.39) it is proved that torque is dependent on quadrature axis component of stator current and is controllable with q-axis stator current component.

As we have seen above torque can be controlled if $\theta_F \text{ and } I_S$ is known. The speed error is calculated from ref. speed, ω_{Rref} and actual rotor speed, ω_R which is obtained from speed sensor. The ref. torque T_{Eref} can be calculated as a function of speed error. Similarly, λ_{rref} can be obtained from rotor speed, it is kept constant at rated for speed in range of 0 to 1 p.u., and for speeds above 1 p.u. it is varied as:

$$\lambda_{rref} = \frac{\omega_{Rrated}}{|\omega_R|}\lambda_{rrated} \tag{4.40}$$

Eq. (4.40) which clearly shows that above-rated speed flux weakening will happen to keep constant power. The command values of $\omega_F, I_T, and\, I_f$ is produced by vector controller as:

$$I_{Tref} = \frac{T_{Eref}}{K_T \lambda_{rref}} \tag{4.41}$$

$$I_{fref} = \frac{1}{L_M}(1 + \frac{L_R}{R_R})\frac{d\lambda_{rref}}{dt} \tag{4.42}$$

$$\omega_{fref} = \frac{L_M R_R}{L_R}\frac{I_{Tref}}{\lambda_{rref}} \tag{4.43}$$

$$\theta_{fref} = \int \omega_{fref} dt \tag{4.44}$$

From $I_{Tref}\, and\, I_{fref}$ we can calculate $I_{qsref}\, and\, I_{dref}$ and using 2-3 phase transformations we can get $I_{Asref}, I_{Bsref}, and\, I_{Csref}$. These 3-phase current are command to the inverter. Which further supplies to motor.

In Indirect vector control $\omega_{fref}\, and\, \theta_{fref}$, both are dependent on machine parameters. That's why it must be equal to the actual values in every operating condition. For the fact, the temperature varies rotor resistance and the stator current varies leakage inductance. This detuning of slip speed which causes coupling effect is the major disadvantage of indirect vector control.

d-q axis will move in opposite direction, if the estimated rotor resistance is lower than the actual rotor resistance which causes the actual frequency to be higher than ref. slip frequency. As torque increases the corresponding q-axis current increases, current component along with d-axis increases too and it affects flux (Increases). In this way, both flux and torque command will deviate from the actual value. Equivalently, torque & flux component will be lower than the actual value if the estimated rotor resistance is more than actual ones.

5 Sensorless Vector Control

Other than voltage and current sensors, position and speed sensors are used in vector controlled drives. The position and speed sensors causes motor size and cost to increase and reduces noise immunity and reliability. These drives are popular for the definite control of speed and torque. In this type of drives, the speed is calculated from supplied voltage, current, and frequency. These drives are popular nowadays and not in use where fluent operation at lower speed is not required.

Common feature of Sensorless drives is that they depend on machine parameters, Temperature, Frequency, Saturation Levels. To overcome parameter variation various adaptation scheme is proposed. The ideal Sensorless drives have an accuracy of 0.5% between speed information and control from Zero to Rated Speed for all operating conditions and is independent of saturation level and parameter variation. Conventional approach doesn’t suit for stable, Low-speed operation in Sensorless IM drive. The estimation of speed is difficult due to parameter variation and noise. At zero frequency, it becomes unobservable. However, various effects such as rotor slot harmonic, Saliency etc can be used for estimation of speed. And also, it became possible to get the position of rotor at low speed even at Zero frequency in squirrel cage IM using variation in rotor angle of the Machine inductances.

5.1 For Low-Performance Application

Slip and speed from steady state equivalent circuit can be estimated with low-cost sensing device, which uses stator current, and voltages of the induction motor. The slip is calculated as:

$$s = \frac{R_R^{'}}{3} \frac{\omega T_E}{|V_M|^2} \tag{5.1}$$

Here, the slip estimation is using monitored values of Electromagnetic torque and Magnetising voltage (Magnitude) which is the difference between terminal voltage and sum of stator losses and stator leakage voltage drop, and the torque can be computed with the help of stator current and voltage. This EMT can be calculated using air gap power too.

$$T_E = \frac{P_{AG}}{\omega} \tag{5.2}$$

This P_{AG}, Air Gap Power can be computed by subtracting DC Link Power and Power losses of inverter, choke, stator.

All these can be done with monitored current and voltages. Any inaccuracy in measurement can affect in estimation of slip. At low slip values above mentioned losses has a significant value.

This monitoring method cannot be used in dynamic conditions and slip estimation is done only when it is near to its rated value, i.e. for low values only. Thus, there is limit in range of speed, and also the magnetizing voltage which contains derivative of the stator current which can cause serious problem if it contains noise.

That’s why this technique can only be used in low-performance device.

5.2 For High-Performance Drives

In this part, various methods are pointed out which are used in high-performance drives for calculation of slip, rotor angle, speed, flux:

i. Observers (Kalman, Leuenberger);
ii. Using Artificial Intelligence (Neural Network, Fuzzy Logic Based, etc);
iii. Estimators Using Saliency effects;
iv. Estimators using spatial saturation (3rd Harmonic Voltage)
v. Estimators using monitored Current and Voltages;
vi. Model Reference Adaptive System (MRAS)

We are here to talk about MRAS.

5.3 Model Reference Adaptive Systems

In other estimators like in open loop where Stator current and voltage are monitored, this process highly depends upon machine specifications. To increase the accuracy closed loop observers are used. State variables, rotor flux linkage ($\lambda_{dR}, \lambda_{qR}$) and Back EMF ($E_d, E_q$) components are calculated using sensed stator current and voltages in reference model and compared with adaptive model's state variables. This error is then fed to adaptive mechanism which outputs estimated rotor speed and it regulate adaptive model so that error is minimized.

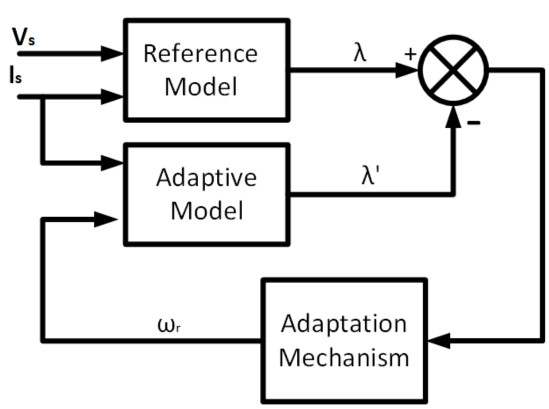

Figure 10: Basic MRAS Scheme

The adaptive mechanism is done using Popov's hyperstability criteria. It makes this system fast and stable and it produces speed tuning signal using state errors. Estimated rotor speed can be produced using this speed tuning signal with PI type controller. A total of four scheme has been discussed which use the speed tuning signals ($\in$), $\in_{\omega} = \text{Im}(\lambda_R^S \widehat{\lambda}_{Rref}^S)$, $\in_e = \text{Im}(e\widehat{e}_{ref})$, $\in_{\Delta e} = \text{Im}(\Delta e i_{sref})$ *and* $\in_{\Delta e'} = \text{Im}(\Delta \overline{\text{E}} \delta i_{sref})$. Also, AI based MRAS estimators are also there and some estimators which do not even contain any adaptive model. The main drawback comes when integrators are used. Measured stator current and voltages are used to find all these but in VSI fed inverter drive there is no need to measure stator voltages. Using switching states and DC link voltages it can be done. Best speed estimator in all of these is AI based MRAS.

5.3.1 Popov's Hyperstability Theorem

This theorem proves that there is a need of PI controller and it also shows what type of speed tuning signal has to be used. In particular, MRAS estimator consists of feedforward time-invariant system and feedback nonlinear time varying system.

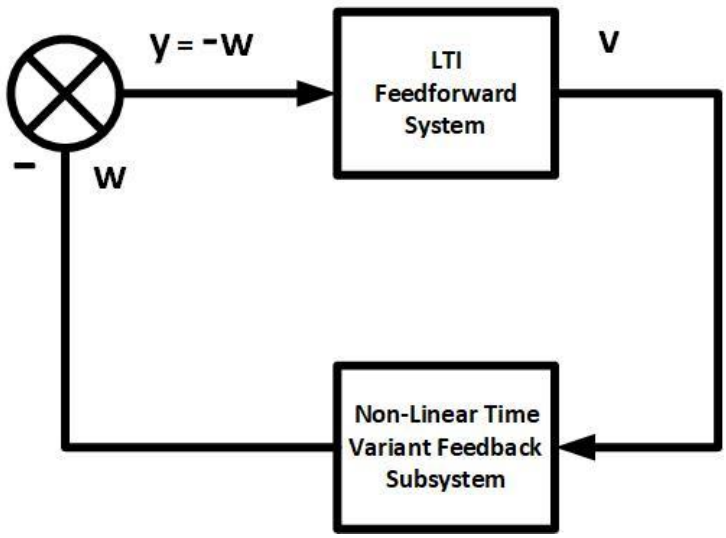

Figure 11: Equivalent Non – Linear Feedback System

When stator voltages and current (y) is given to this LTI system it outputs speed tuning signal (v). Nonlinear time variant system outputs y = -w. Using this theorem, we can choose the adaptation mechanism which says transfer function must be positive real matrix of this LTI system and the feedback system which is non-linear time varying must satisfy Popov's Integral theory which says $\int v^T y dt \geq 0$ in $[0, t_1], \forall t_1 \geq 0$. For adaptation mechanism, Transfer function of LTI feed forward system has to be known.

It gives stable nonlinear feedback system.

5.3.2 Speed tuning Signal $\in_{\omega} = \mathbf{Im}(\lambda_R^s \widehat{\lambda}_{Rref}^s)$

Rotor flux linkage in stationary reference frame using stator voltage of IM.

$$\lambda_{dR} = \frac{L_R}{L_M}\left[\int (V_{sd} - R_s i_{sd})dt - L_s' i_{sd}\right] \tag{5.3}$$

$$\lambda_{qR} = \frac{L_R}{L_M}\left[\int (V_{sq} - R_s i_{sq})dt - L_s' i_{sq}\right] \tag{5.4}$$

Eq. (5.3),(5.4) represent stator voltage model (Reference model, do not contain rotor speed). These equations can be modified using stator voltage equation in SRF and eliminating rotor current using;

$$0 = R_R I_{Rs} + \frac{d\lambda_{Rs}}{dt} - j\omega_R \lambda_{Rs} \tag{5.5}$$

$$I_{Rs} = \frac{(\lambda_{Rs} - L_M I_{ss})}{L_R}$$
$$\lambda_{Rd} = L_M I_{MRd}$$
$$\lambda_{Rq} = L_M I_{MRq}$$

We get

$$\lambda_{rd} = \frac{1}{\tau}\left[\int (L_M I_{sd} - \lambda_{Rd} - \omega_R \tau_r \lambda_{rq}\right]dt \tag{5.6}$$

$$\lambda_{rd} = \frac{1}{\tau}\left[\int (L_M I_{sd} - \lambda_{Rd} - \omega_R \tau \lambda_{Rq}\right]dt \tag{5.7}$$

Eq.(5.6),(5.7) contains rotor speed which shows it is an adjustable model. Now, from the reference model and adjustable model when compared gives error signal.

i.e. $$\in = \lambda_{Rq}\lambda_{Rd}' - \lambda_{Rd}\lambda_{Rq}' \tag{5.8}$$

Which is then fed to PI controller which outputs Estimated rotor speed.

5.3.3 Speed tuning Signal $\in_e = \mathrm{Im}(e\hat{e}_{ref})$

This scheme uses Back EMF to generate the speed error signal. It is advantageous to the previously mentioned scheme in a way that it doesn't use Integrators in reference model.

Direct and Quadrature Axis Back EMF Eq. for reference model;

$$\mathrm{E}_d = V_{sd} - R_s I_{sd} - L_s' \frac{dI_{sd}}{dt} \tag{5.9}$$

$$\mathrm{E}_q = V_{sq} - R_s I_{sq} - L_s' \frac{dI_{sq}}{dt} \tag{5.10}$$

As we can see here no integrators were used. d-q components of stator voltage could be obtained from measured terminal voltages or from the switching states of the inverters.

For Adaptive model;

$$\mathrm{E}_d' = \frac{L_M(L_M I_{sd} - \lambda_{Rd} - \omega_R \tau \lambda_{Rq})}{L_R \tau} \tag{5.11}$$

$$\mathrm{E}_q' = \frac{L_M(L_M I_{sq} - \lambda_{Rq} + \omega_R \tau \lambda_{Rd})}{L_R \tau} \tag{5.12}$$

Speed tuning signal would be;

$$\in = E_d E_q' - E_q E_d' \tag{5.13}$$

i.e. difference between back EMF vectors.

It can give very accurate control even at low speed if stator resistance is calculated accurately. Well, whichever process doesn't use stator resistance then that would be better because this stator resistance varies with temperature and it can affect the observer at low speed. Which is not desirable.

5.3.4 Speed tuning Signal $\in_{\Delta e} = \mathrm{Im}(\Delta e i_{sref})$

As discussed above in Sec. 5.3.2 the integrator present in reference model creates problem which was then compensated in Sec 5.3.3 using Back EMF components and new problem came into light where due to temperature, machine parameters get affected and it creates problem in low-speed region.

In this scheme tuning signal is $\in_{\Delta e} = \mathrm{Im}(\Delta E I_{sref}) = I_s x E - I_s x E'$

Where,

$$I_s x E = I_s x (V_S - L_s \frac{dI_s}{dt}) \tag{5.14}$$

is reference model output. It is shown here that it doesn't contain any resistance term. Here, $v_s = R_s I_s + L_s \frac{dI_s}{dt} + E$ therefore vector product $I_s x u_s$ doesn't contain any machine parameter. The other voltage components can be calculated as per earlier.

Adaptive model output;

From Eq. (5.11),(5.12) $\widehat{\mathrm{E}} = \mathrm{E}_D + j\mathrm{E}_Q$

Therefore,

$$I_s x \widehat{\mathrm{E}} = \frac{L_M}{L_R}\left[\frac{1}{\tau}\lambda_R' x I_s + \omega_r (I_s x j \lambda_R')\right] \tag{5.15}$$

Speed tuning signal can be calculated by Eq. (5.14),(5.15)

When this is implied good performance can be obtained at very low speeds. No stator resistance can affect this signal. IN rotor flux oriented control, the time constant (τ) used in observer which makes it insensitive to variation of the time constant too [31]. This speed observer has its limitations too, the stator inductance. Deviation of rotor time constant also affects the estimated rotor speed in steady state. It cannot be used at 0 frequency.

5.3.5 Speed Tuning Signal $\in_{\Delta e'} = \mathrm{Im}(\Delta \bar{\mathrm{E}} \delta i_{sref})$

In the previously proposed scheme, the Stator Inductance do affect the speed estimation. This is eliminated in this section. A perfect speed tuning signal can only be generated if the measured stator resistance and rotor time constant can be known precisely. This can be done with the derivative ($\delta = d/dt$) of the stator current and $\Delta \mathrm{E}$ signifies back EMF error.

From Reference Model;

$$\mathrm{E} = \left[v_{sq}\delta I_{sq} - u_{sd}\delta I_{sq} - R_s(I_{sq}\delta I_{sd} - I_{sd}\delta I_{sq}) \right] \tag{5.16}$$

From Adaptive Model;

$$\widehat{\mathrm{E}} = \frac{L_M}{L_R}\left[\widehat{\omega}_R(\lambda_{rd}\delta I_{sd} + \lambda_{rq}\delta I_{sq}) + (\lambda_{rd}\delta I_{sq} - \lambda_{rq}\delta I_{sd} + I_{sq}\delta I_{sd} - I_{sd}\delta I_{sq}) / \tau \right] \tag{5.17}$$

Speed Tuning Signal;

$$\Delta \mathrm{E} = \mathrm{E} - \widehat{\mathrm{E}}$$

Eq. (5.16),(5.17) both do not contain any voltage drops across Inductances. But in the reference model, it contains Stator resistance. Thus, better speed estimation depends upon the accuracy of estimation of stator resistance.

Also,

$$[(\delta I_s)x\mathrm{E}] = [(\delta I_s)x(\delta\lambda_s)] = [(\delta I_s)x(V_s - R_s I_s)] \qquad (5.18)$$

Which is clear that $(\delta I_s)xI_s \neq 0$ due to the presence of derivative of stator flux linkage, not stator flux linkage itself and therefore no integration present. In adjustable model rotor, flux linkage components use voltage equations (5.3) (5.4) and monitored stator current components and estimated rotor speed. It depends on rotor time constants. Similar to every scheme the estimated speed is obtained from PI controller.

6 Result

The following result has been taken on MATLAB Simulation.

Motor Parameters used here are: -

Rated Power =74.60kW (100hp)

Supply Voltage 460Vrms, 50Hz

Stator resistance = 14 mohm

Rotor resistance = 9 mohm

Stator Leakage Inductance = 0.3 mH

Rotor Leakage Inductance = 0.3 mH

Mutual Inductance = 10 mH

Pole Pairs (P) = 2

Rotor Inertia = $3.1 kgm^2$

Vector Control

Speed Reference in simulation: Time [0 1] sec, Speed [500 0] rpm

Torque Reference in simulation: Time [0 0.5 1.5]sec, Speed [0 800 -800] Nm

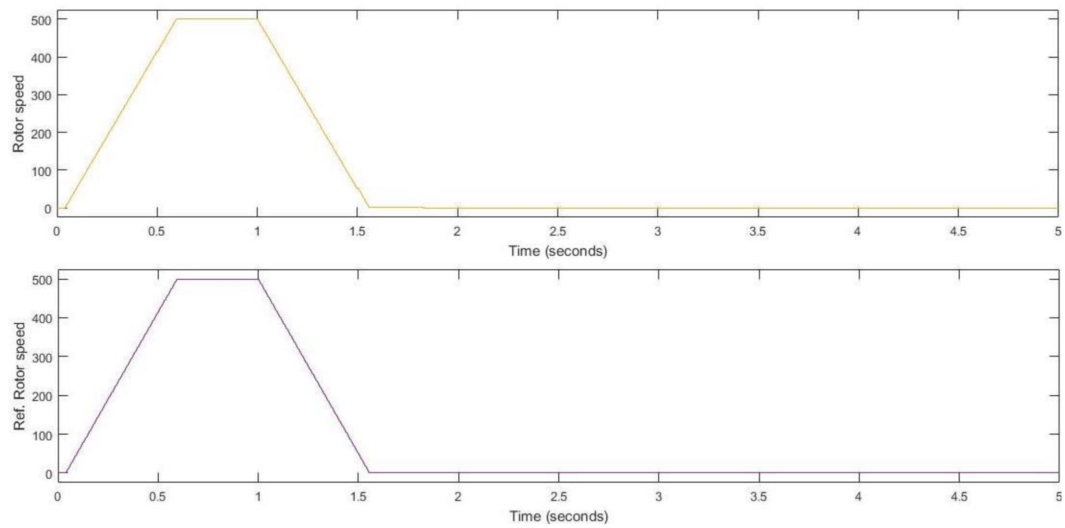

Figure 13: Rotor and reference Rotor Speed of Vector Controlled Induction Motor Drive

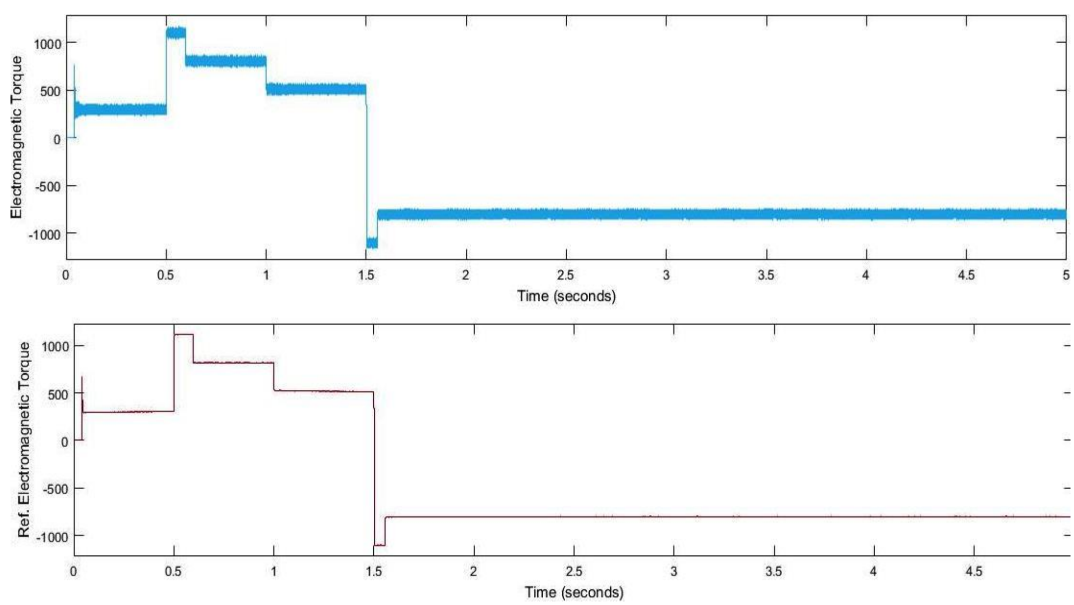

Figure12:Electromagnetic torque and Reference torque of vector controlled Induction Motor Drive

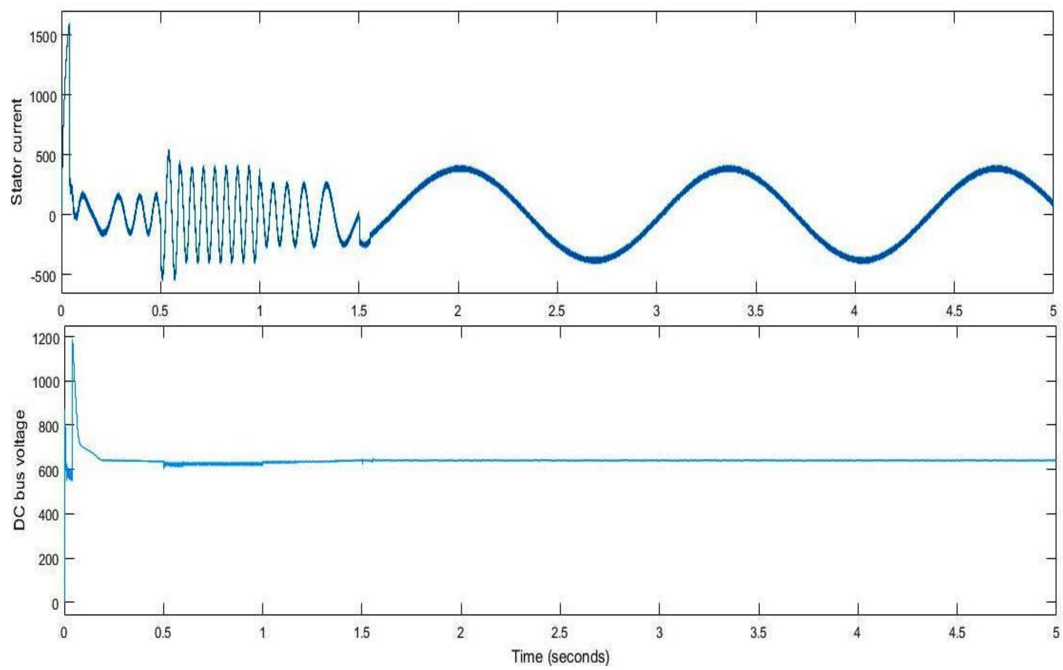

Figure 14:Stator Current and DC Bus Voltage of Vector controlled IM Drive

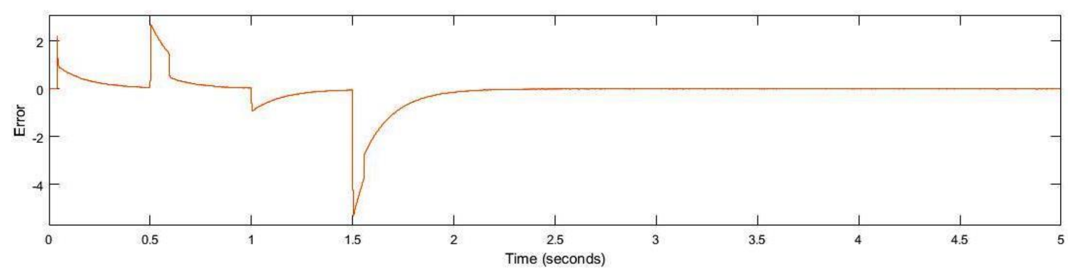

Figure 15: Error Signal

Fig.12, Fig.13, Fig.14, Fig.15 shows Speed, Torque, Stator Current, DC Bus voltage & error signal respectively, operational result of vector controlled induction motor drive. As the simulation starts at 0 sec the speed is set to 500 rpm which can be clearly seen here that it is ramping up to 500rpm at this instant the torque reference is 0, at 0.5 second torque is set to 800Nm as the speed is ramping till now so that the torque is increased to the maximum value given by user after that it stabilizes at 800Nm as the speed is set. At 1s the speed is changed to 0 rpm abruptly and mechanical load torque is inverted from 800 to -800 Nm so that speed is ramping down. Shortly after 1.5s the speed is set to 0.

During this whole operation, the DC Bus voltage is well regulated.

Sensorless Vector Control

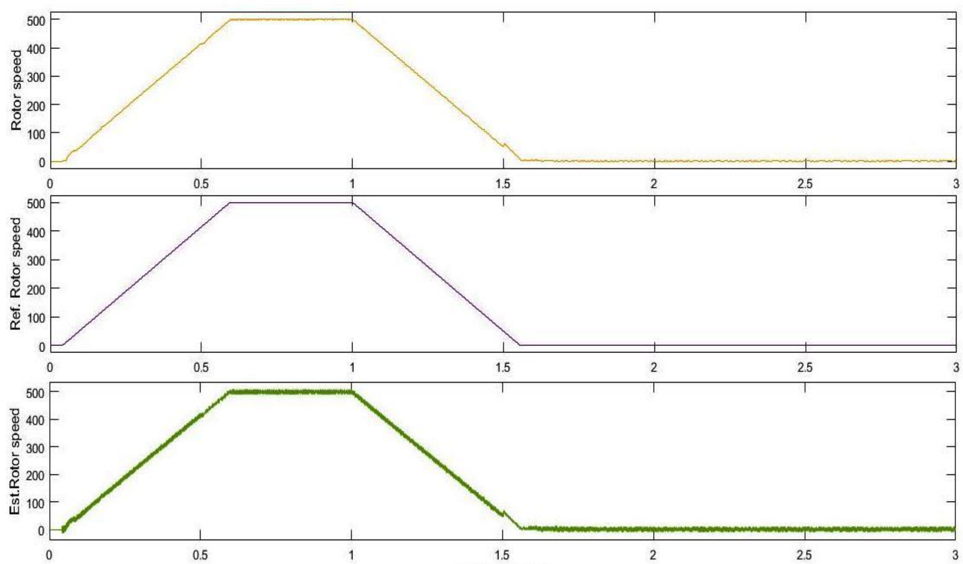

Figure 16:Rotor Speed, Reference Speed & Estimated Speed of sensorless Vector Controlled IM Drive

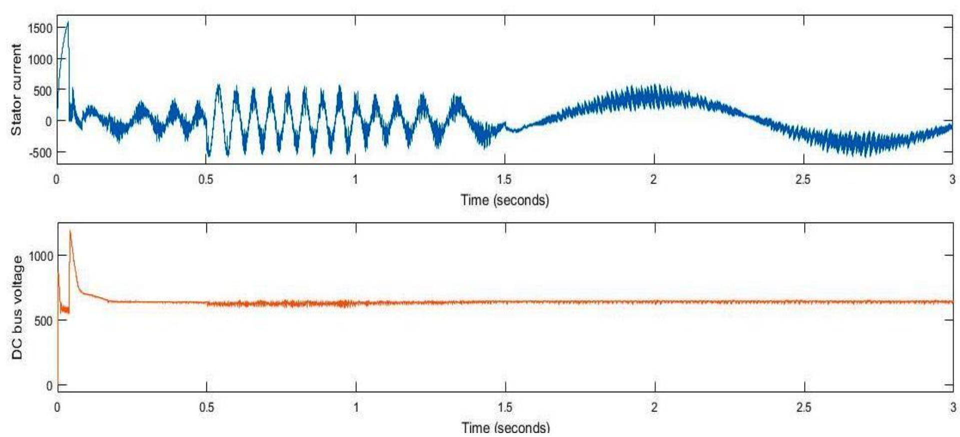

Figure 17:Stator Current and DC Bus Voltage of Sensorless Vector Controlled IM Drive

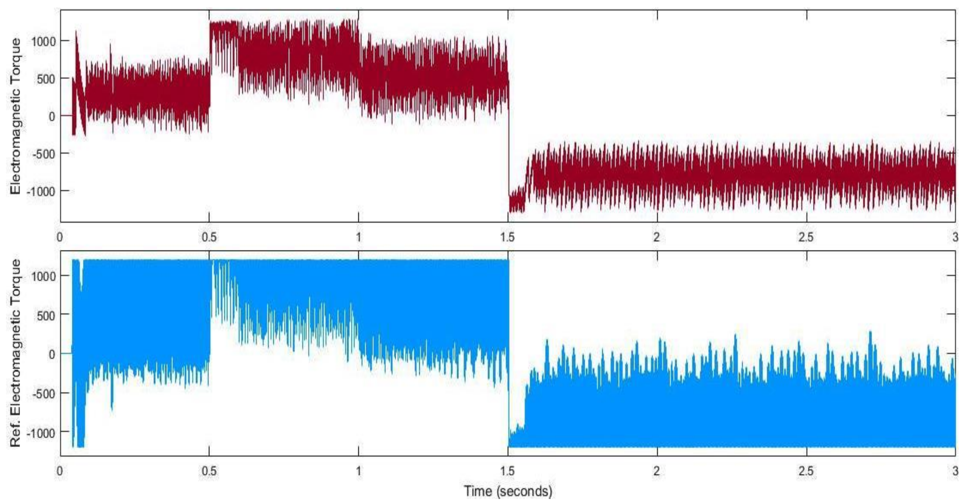

Figure 18:Electromagnetic Torque and Reference Torque of Sensorless controlled IM Drive

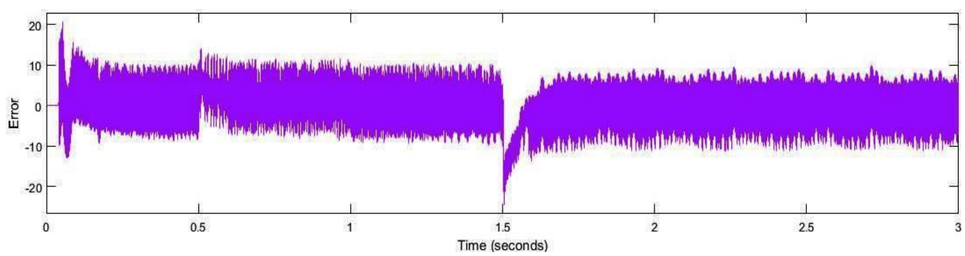

Figure 19:Speed Error Signal

Fig. 16, Fig. 17, Fig. 18, Fig.19 shows Speed, Stator Current & DC bus voltage, Torque and Error respectively of sensorless vector controlled induction motor drive. We can see that rotor follows the reference speed with the help of estimated rotor speed. Similar to vector control the speed here ramps up to 500 rpm and torque is then set to 800 Nm at 0.5s and then the torque sets to rated torque 1200Nm. And when the speed settles at 500 rpm the torque then follows the reference value. At 1s user defined speed is changed to 0 rpm thus speed decelerates to 0 rpm precisely even if the load torque is changed from 800 to -800Nm, at 1.5 sec. when the motor stabilizes at 0 rpm.

The DC Bus voltage could be seen regulated whole period.

Scalar Control

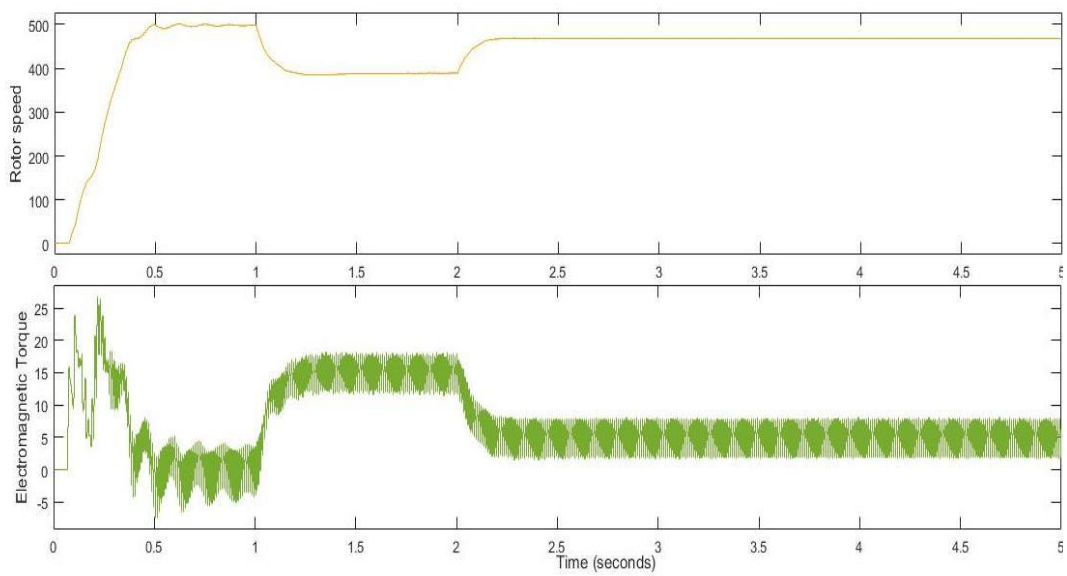

Figure 20: Constant V/f control Ref. Speed 500rpm Load Torque[0 15 5] at [0 1 2]sec

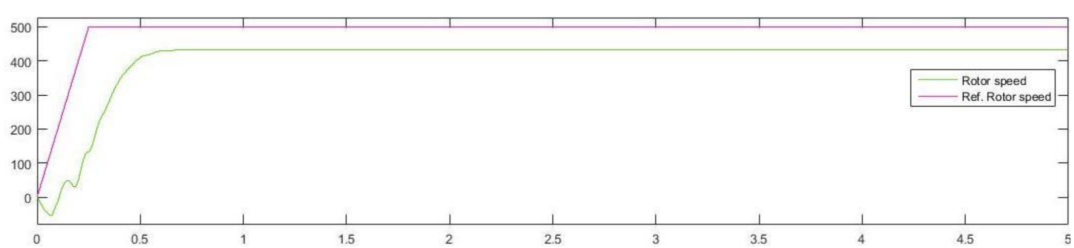

Figure 21: V/f Control reference speed 500rpm Load Torque 10 Nm

From Fig. 20 it is clear that in the scalar control load torque is inversely proportional to speed. Which means rotor speed fluctuates with change in load torque. Which clears that it is impossible to set constant speed without load torque variations.

Fig. 21 shows speed regulation of scalar control for reference speed 500 rpm which clearly shows poor performance.

It is slow method which is based on magnitude of voltage and frequency not current.

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