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Detailed Review of Induction Motor Drive

J. Ahamad[1], A. Srivastava[2], P. Singh[3],
[1]Assistant Professor H.B.T.U. Kanpur,
[2]Guest Faculty H.B.T.U. Kanpur,
[3]Assistant Professor Axis Colleges Kanpur

Abstract: The induction motor was invented in 1888 by Nikola Tesla [1]. Dolivo Dobrowolski extended its structure in 1890 which is known as squirrel-cage motor [2].In modern days industrialized countries, more than half the total electrical energy used is converted to mechanical energy through induction motor This paper deals with literature survey of various existing converter topologies, and control strategies which have been proposed for induction motor drives. A study of the merit and demerit of different converter topologies have been carried out and various control strategies have been analyzed in this paper.
Keyword: D.T.C., I.F.O.C., I.F.O.C.I.M., F.G.A., T.H.D. etc

I. INTRODUCTION

An electrical machine has gone through slow but sustained evolution during the past century. The advent of powerful digital computers, incipient and ameliorated materials, coupled with extensive R&D, has resulted in higher power density, higher efficiency, and many performance enhancements of machines. The dramatic improvements that are required in the performance, reliability, and cost-effectiveness of electric drives can only be achieved by developing an integrated system approach based on the advanced packaging of semiconductor devices, and innovative circuits with integrated functionality, suitability for control, and application versatility. Three- phase induction motor outshines in terms of machine efficiency, robustness, reliability, durability, power factor, ripples, stable output voltage and torque [3]. When operated directly from the line voltages, they operate at a nearly constant speed. However, with the help of power electronic converter it is possible to vary the speed of an induction motor. The fundamental elements needed in an Electric motor drive system (as shown in Figure 1) include:

A. Power electronic Converter
B. Electric Motor
C. Controller (Analog/Digital)

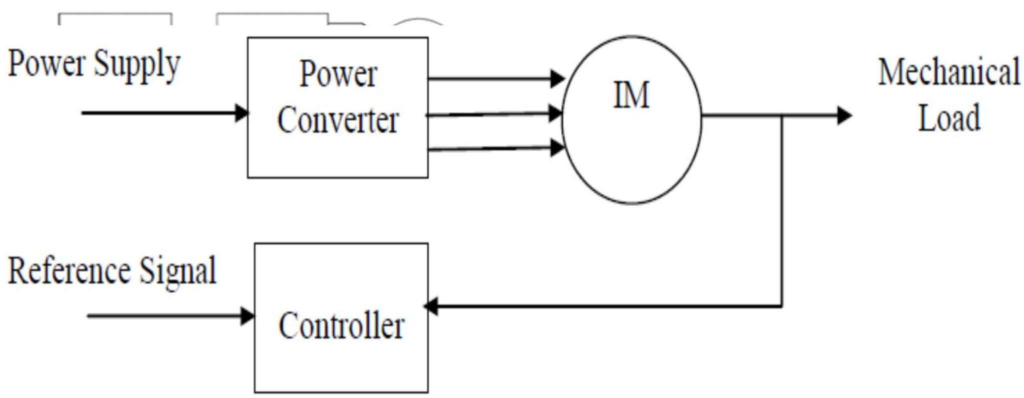

Figure 1: Major drive type categories

II. CONTROL ALGORITHM FOR INDUCTION MOTOR DIVES

High performance drive applications usually require a fast torque response, with DC drives preferred in the past. The advantages of AC drives include robustness, compactness, economy, and low maintenance. Previously torque response control was a problem. Advances in power switching devices, electronic processing, and control have led to great improvements. Such controllers build upon good steady state performance and can give excellent transient behavior. Variable-frequency AC machine control can be divided into scalar and field oriented or VC shown in figure 2. Scalar control uses magnitude and frequency control. VC uses orientation in addition. Variants include direct torque control (DTC) which also exploits spatial orientation but aims to control current and hence torque by more directly switching the voltage rather than using PWM [4], [5].

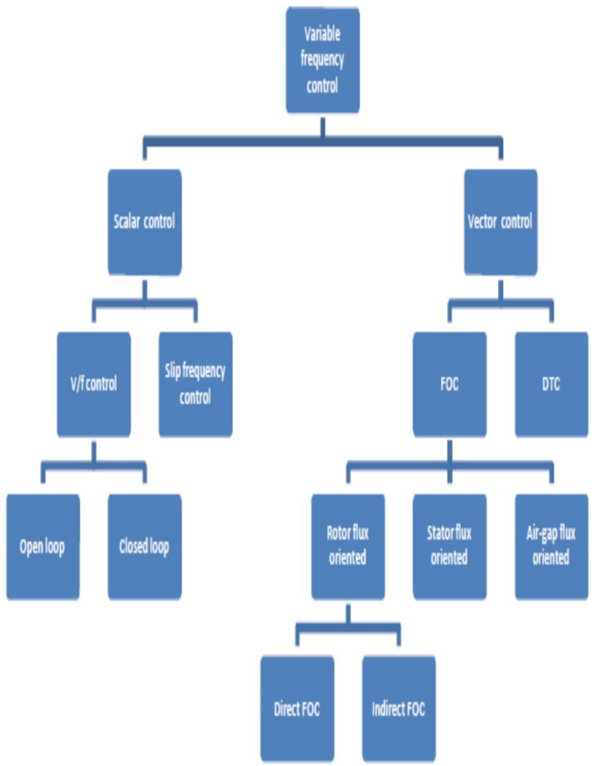

Figure 2: Classification of variable frequency control of induction motor drive

A. Scalar Control

It is based on steady state relationships, usually only magnitude and frequency are controlled, not space vector orientation. Making terminal voltage magnitude proportional to frequency results in approximately constant stator flux, desirable to maximize the capability of the motor. The classical variable frequency V/f scheme is a scalar control based on this principle, with voltage boost at low frequency usually introduced to counteract the larger effect of stator resistance at low speeds. Scalar control, often open-loop apart from stator current monitoring for fault detection, gives an economical drive with good behavior, but transients may not be well controlled. More sophisticated variants can improve behavior, perhaps with better handling of parameter variations, particularly of stator resistance.

B. Vector Control

1) *Direct Torque Control:* Direct torque control (DTC) is an IM drive that is frequently used in IM control because of its balance between simplicity of design and its decent performance. It is a motor drive that uses little parameter information, and therefore is generally not considered to be a high performance drive. It essentially has two user inputs: torque and stator flux. The typical DTC IM drive uses these two inputs in hysteretic control as inputs to a look-up switching state table. In addition to the previous two inputs, standard DTC uses the angle of the stator flux to determine its “sector.” The typical table uses six flux sectors to distinguish where the flux angle lies. The output of the table is a vector that contains the information telling which gates of the inverter should be on at any point in time. It should be emphasized that this configuration is not necessarily optimal, but it is indeed common. Sikorski et al. [16] compare linear DTC-SVM to nonlinear DTC methods, such as DTC-δ, DTC-2x2, and DTFC-3A, using steady-state performance metrics. Excellent numerical hardware results were given that compared the variations in DTC. The average switching frequency was kept the same throughout these trials in an attempt to keep one

variable constant. It was found that the ripple for the current and torque was found to be smallest in the standard DTC case, which was unexpected considering the higher level of complexity using the other methods.

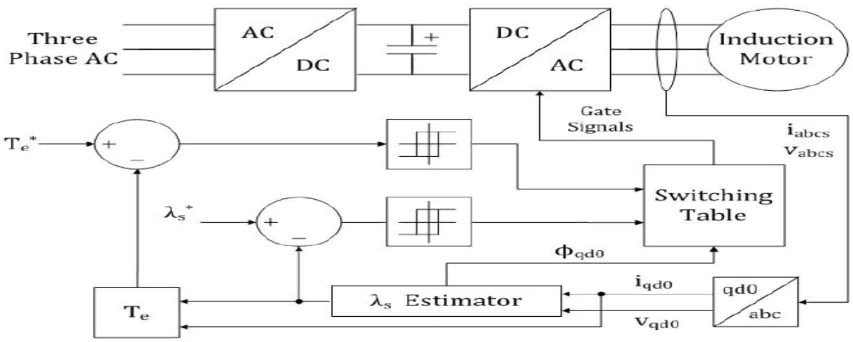

Figure 3. DTC Block Diagram

2) *Indirect Field-Oriented Control:* Indirect field-oriented control, or IFOC, is the most common IM drive because of its use of moderate amounts of parameter information to give it respectable performance 21 while also not requiring a high level of sophistication. Field-oriented control (FOC) was introduced by Blaschke in 1971 [17, 18]. FOC was created to imitate the control of a separately excited dc motor. In a similar fashion to the dc motor, the FOC drive keeps the rotor flux perpendicular to the stator flux to get the maximum output torque possible. The big advantage of FOC is that the flux and the torque can be decoupled by insuring that the other is in steady state. In this fashion, the dynamics can be independently controlled by the user. Because of this, the classical feedback control can be used to obtain desired motor performance. The basic attribute of IFOC is that it uses an estimate of the rotor flux in determining the next state of the inverter. In particular, it uses the angle of the rotor flux to determine where the flux is in vector-space. The angle is calculated by (1).

$$\rho = \tan^{-1}\left(\frac{\lambda_{qr}}{\lambda_{dr}}\right) \quad (1)$$

This flux vector angle is then used in a matrix transformation that converts the stator current and rotor flux values into a new state space { ω_r,ψ, I_q, I_d, ρ} [19] where

$$\begin{aligned}
\omega_r &= \omega_r \\
\psi &= \sqrt{\lambda_{qr}^2 + \lambda_{dr}^2} \\
i_q &= \frac{\lambda_{dr} i_{qs} - \lambda_{qr} i_{ds}}{\psi} \\
i_d &= \frac{\lambda_{dr} i_{ds} + \lambda_{qr} i_{qs}}{\psi} \\
\rho &= \tan^{-1}\left(\frac{\lambda_{qr}}{\lambda_{dr}}\right)
\end{aligned} \quad (2)$$

This is equivalent to using the following matrix conversion for the same currents and flux values:

$$\begin{pmatrix} i_d \\ i_q \end{pmatrix} = \Gamma_\rho \begin{pmatrix} i_{ds} \\ i_{qs} \end{pmatrix}$$

$$\begin{pmatrix} \psi \\ 0 \end{pmatrix} = \Gamma_\rho \begin{pmatrix} \lambda_{dr} \\ \lambda_{qr} \end{pmatrix}$$

$$where\ \Gamma_\rho = \begin{pmatrix} \cos(\rho) & \sin(\rho) \\ -\sin(\rho) & \cos(\rho) \end{pmatrix} \quad (3)$$

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If we let the vector be equal to that in (4),

$$\begin{pmatrix} v_{qs} \\ v_{ds} \end{pmatrix} = \psi \begin{pmatrix} \lambda_{dr} & \lambda_{qr} \\ -\lambda_{qr} & \lambda_{dr} \end{pmatrix}^{-1} \begin{pmatrix} v_d \\ v_q \end{pmatrix}$$

$$where \begin{pmatrix} v_d \\ v_q \end{pmatrix} = \sigma \begin{pmatrix} -n_p \omega_r i_q - \dfrac{M r_r i_q^2}{L_r \psi} + u_{flux} \\ n_p \omega_r i_d + \dfrac{n_p \omega_r M \psi}{\sigma L_r} + u_{speed} \end{pmatrix} \qquad (4)$$

$$and\ where\ \sigma = \frac{L_s L_r - M^2}{L_r}$$

the unwanted nonlinear terms cancel, and the closed loop dynamic system equations become similar to that of a dc motor, as seen in (5). The quadrature axis current represents the speed-producing element, while the direct axis current represents the torque-producing element. As revealed in (5), i_q and i_d are asymptotically decoupled in this reference frame. This will allow the user to independently control their steady-state values as well as their dynamic performance. The only downside of FOC is the nonlinear nature of ρ and the fact that it is already very difficult to estimate accurately. This is a common downside of many field-oriented controllers.

$$\begin{aligned}
\frac{d\omega_r}{dt} &= \frac{3 n_p M \psi i_q}{2 J L_r} - \frac{T_{load}}{J} \\
\frac{d\psi}{dt} &= -\frac{r_r \psi}{L_r} - \frac{r_r M i_d}{L_r} \\
\frac{di_q}{dt} &= -\left(\frac{M^2 r_r}{\sigma L_r^2} + \frac{r_s}{\sigma} \right) i_q + u_{speed} \\
\frac{di_d}{dt} &= -\left(\frac{M^2 r_r}{\sigma L_r^2} + \frac{r_s}{\sigma} \right) i_d + \frac{r_r M \psi}{L_r} + u_{flux} \\
\frac{d\rho}{dt} &= n_p \omega_r + \frac{r_r M i_q}{L_r \psi}
\end{aligned} \qquad (5)$$

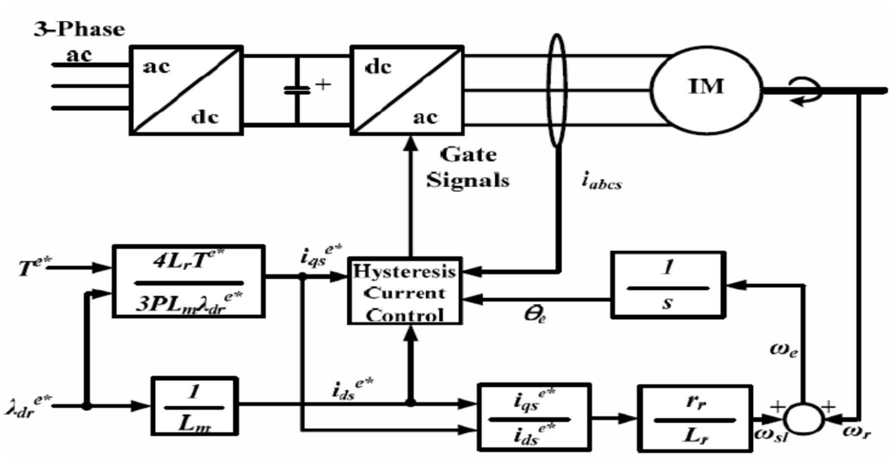

Figure 4. Block Diagram of IFOC with Current Hysteresis

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C. Feedback Linearization: Input-Output Decoupling Control

In [10], Krzeminski comes up with a type of nonlinear state feedback control which is completely input-output decoupled at all times even through transients. This differs from IFOC because IFOC is decoupled only when flux and speed are in steady state. FL-IODC achieves better performance than IFOC in theory due to accounting for the stator resistive drop and other terms that allow it to have complete decoupling. The drawback of this type of control is the additional parameter sensitivity that results from the addition of these extra terms. The inputs in this control scheme are and v_{qs},v_{ds} while the outputs are the rotor mechanical speed, ω_r , and the flux magnitude squared,

ψ^2.Krzeminski assumes in the paper that the load torque response is known. If this is the case, then the new state space is:

$$\begin{aligned}
\frac{d\omega_r}{dt} &= \frac{3n_p M\psi i_q}{2JL_r} - \frac{T_{load}}{J} \\
\frac{d\psi}{dt} &= -\frac{r_r\psi}{L_r} - \frac{r_r M i_d}{L_r} \\
\frac{di_q}{dt} &= -\left(\frac{M^2 r_r}{\sigma L_r^2} + \frac{r_s}{\sigma}\right) i_q + u_{speed} \\
\frac{di_d}{dt} &= -\left(\frac{M^2 r_r}{\sigma L_r^2} + \frac{r_s}{\sigma}\right) i_d + \frac{r_r M\psi}{L_r} + u_{flux} \\
\frac{d\rho}{dt} &= n_p\omega_r + \frac{r_r M i_q}{L_r\psi}
\end{aligned} \tag{6}$$

From (6) and the original dynamic motor model, we get the following dynamic system, as seen in [8]:

$$\begin{aligned}
\frac{dy_1}{dt} &= y_2 \\
\frac{dy_2}{dt} &= f_{21}(y_1 \cdots y_5) + f_{22}(y_1 \cdots y_5) v_{ds} + f_{23}(y_1 \cdots y_5) v_{qs} \\
\frac{dy_3}{dt} &= y_4 \\
\frac{dy_4}{dt} &= f_{41}(y_1 \cdots y_5) + f_{42}(y_1 \cdots y_5) v_{ds} + f_{43}(y_1 \cdots y_5) v_{qs} \\
\frac{dy_4}{dt} &= f_5
\end{aligned} \tag{7}$$

If one sets nd *vq, vds* to the vector

$$\begin{pmatrix} v_{ds} \\ v_{qs} \end{pmatrix} = \begin{pmatrix} f_{22}(y_1 \cdots y_5) & f_{23}(y_1 \cdots y_5) \\ f_{42}(y_1 \cdots y_5) & f_{43}(y_1 \cdots y_5) \end{pmatrix}^{-1} \begin{pmatrix} -f_{21}(y_1 \cdots y_5) + u_{speed} \\ -f_{41}(y_1 \cdots y_5) + u_{flux} \end{pmatrix} \tag{8}$$

$$\begin{aligned}
\frac{dy_1}{dt} &= y_2 \\
\frac{dy_2}{dt} &= u_{speed} \\
\frac{dy_3}{dt} &= y_4 \\
\frac{dy_4}{dt} &= u_{flux} \\
\frac{dy_5}{dt} &= n_p y_1 + \frac{2r_r}{3n_p y_3}(Jy_2 + T_{load})
\end{aligned}$$

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The new system look like (9)

The following inputs in (10) can be set to completely decouple the inputs from the outputs when using constant design parameters $k_1, k_2\ k_3\ k_4$. Thus, if there is a transient in the flux magnitude squared or the speed of the rotor, the transient will not affect the other variable [9].

$$u_{speed} = -k_1\left(\omega - \omega_{ref}\right) - k_2\left(\frac{T_e}{J} - \frac{T_{load}}{J} - \dot{\omega}_{ref}\left(t\right)\right) + \ddot{\omega}_{ref}\left(t\right) \tag{10}$$

$$u_{flux} = k_3\left(\psi^2 - \psi_{ref}^2\right) - k_4\left(2\frac{R_r}{L_r}\left(M\left(\psi_d i_d + \psi_q i_q\right) - \left(\psi_d^2 + \psi_q^2\right)\right) - \dot{\psi}_{ref}^2\right) + \ddot{\psi}_{ref}^2$$

D. Vectorized Volts-per-Hertz

Vectorized volts-per-hertz is by far the simplest motor drive since it requires no parameter knowledge and is essentially an open-loop drive. Similar to the standard volt/ hertz, it requires a desired operating frequency, f^*, from the user to create a desired voltage on the IM. In addition to the frequency, it also requires a current, i^*_d, to run the drive. This current essentially creates a desired voltage vector,v^*_d, via the stator resistor, R_s . From these two variables, i^*_d and f^*, two reference voltage vectors, v^*_d and v^*_q , are created that are used in driving the inverter output. With the knowledge of both v^*_d and v^*_q, the whole voltage vector is created. Another way of looking at this is that v^*_d and v^*_q represent the voltage vector in rectangular coordinates, but it can also be thought of as a voltage vector in polar coordinates with a magnitude, V , and an angle, θ .

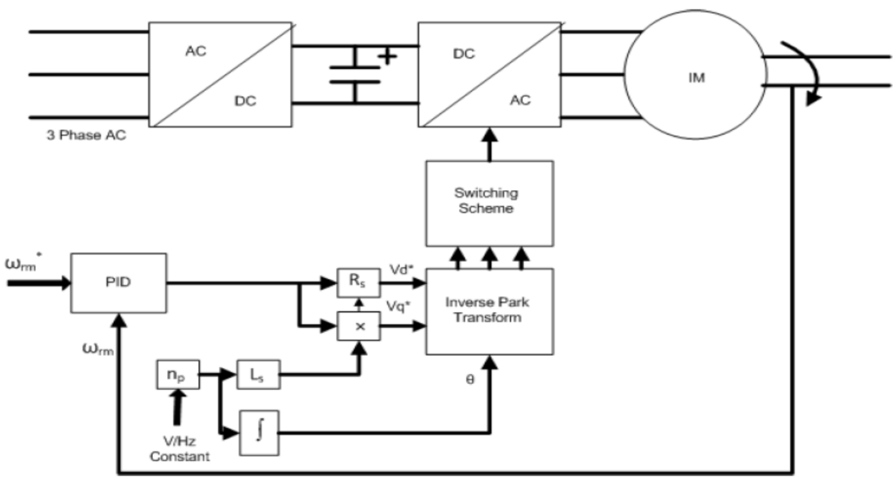

Figure 5. Block Diagram of Vectorized Volts-per-Hertz

III. STRATEGIES FOR PULSE WIDTH MODULATION TECHNIQUE

The Pulse Width Modulation (PWM) is one of the most widely used technique applied in the inverter (DC/AC converter) to output an AC waveform with variable voltage and variable frequency for use in mostly variable speed motor drives. The implementation of the complex PWM algorithms have been made easier with the advent of fast digital signal processors, microcontrollers and Field Programmable Gate Arrays (FPGA) [1]. The concept of harmonic distortion is used as the performance index to compare various PWM algorithms. The most important aim of the modulation strategies is to achieve the maximum voltage with the lowest harmonic distortion. Total Harmonic Distortion (THD) and Weighted Total Harmonic Distortion (WTHD) are the commonly used performance evaluation methods [16], [15]. In the literature, the enormous amount of material published made it challenging one to identify basic modulation principles and apply them to particular implementations and the research is still active in this area [12]-[14]. In spite of these massive research works, only three significantly different alternatives for determining the converter switching pattern that have been proposed for fixed frequency modulation systems. These alternatives are [11]:

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A. Naturally sampled PWM

Switching at the intersection of a target reference waveform and a high frequency carrier;

B. Regular sampled PWM

Switching at the intersection between a regularly sampled reference waveform and a high frequency carrier and

C. Direct PWM

Switching so that the integrated area of the target reference waveform over the carrier interval is the same as the integrated area of the converter switched output. Of the many variations of these three alternatives published, the classification according to switching frequency is [13]:

D. Sinusoidal PWM

The classical Sinusoidal PWM (SPWM) is a form of naturally sampled PWM which uses a triangular carrier to compare against the reference (control/modulating) waveform. The ratio of the amplitude of carrier signal and the control signal is called 'modulation ratio'. If the reference signal is higher than the carrier, the corresponding inverter cell outputs positive voltage; otherwise, the corresponding inverter cell outputs negative voltage. The major limitation with this naturally sampled PWM is the difficulty of its implementation in a digital modulation system due to the intersection between the reference sinusoid and triangular carrier as defined by a transcendental equation which is complex to calculate. This limitation can be overcome by adopting a regular sampled PWM strategy, in which the low frequency reference waveforms are sampled and then held constant during each carrier interval. These sampled values are compared against the triangular carrier waveform to control the switching process of each phase leg, instead of the sinusoidally varying reference [6]. Even though the method is very simple, it has the following drawbacks:

1) SPWM is unable to fully utilize the available DC bus supply voltage to the voltage source inverter.
2) The method cannot completely eliminate the low order harmonics. Therefore the low order harmonics cause loss and high filter requirements.
3) The high switching frequency causes high switching loss and low efficiency.
4) To reduce run-time processing load for slow controllers, three 120° phase shifted sine tables are created in the controller memory. This is an inefficient usage of the controller memory
5) Resulting waveform is tied to the selected switching frequency [7].

E. Space Vector PWM

In the mid 1980's a form of PWM called Space Vector Modulation (SVM) was proposed, which was asserted to offer significant advantages over natural and regular sampled PWM in terms of performance, ease of implementation and maximum transfer ratio [8]. The SVM technique is one of the most popular for bi-level PWM inverter control. In SVM, the three phase stationary reference frame voltages for each inverter switching state are mapped to the complex two phase orthogonal a-b plane. The reference voltage is represented as a vector in this plane and duty-cycles are computed for the selected switching state vectors in proximity to the reference. These vector diagrams are universal regardless of the topology of inverter. Therefore it can be used for diode-clamped or capacitor-clamped. The adjacent three vectors can synthesize a desired voltage vector by computing the duty cycle for each vector [1]. Space vector simultaneously represents three phase quantities as one rotating vector, hence each phase is not considered separately. The three phases are assumed as only one quantity. The space vector is valid for both transient and steady state conditions in contrast to phasor representation. The main benefit of SVM is the explicit identification of pulse placement as an additional degree of freedom that can be exploited to achieve harmonic performance gains [2]. SVM methods generally have the following advantages:

1) Line-to-line voltage amplitude can be as high as the available DC bus voltage and thus 100% DC voltage utilization is possible in the linear operating region
2) In the linear operating range, line-to line voltage amplitude is 15% more in SVM with the modulation index = 0.866, compared to the SPWM with modulation index = 1. Hence, it has the better usage of the modulation index depth.
3) With the increased output voltage, design of motor control system with reduced current rating, keeping the horsepower rating the same. The reduced current helps to reduce inherent conduction loss of the VSI.
4) Only one reference space vector is controlled to generate 3-phase sine waves.
5) The algorithm gives less THD and less switching loss compared to SPWM.
6) It offers several degrees of freedom that can be used effectively to design an improved harmonic spectrum and obtain the desired waveform quality [7].

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7) As the reference space vector is a two-dimensional quantity, it is feasible to implement more advanced vector control using SVM
8) Control is relatively easy and the hardware implementation can be easy by a digital signal processor (DSP).

These advantages make it popular for inverter control and hence the research is still active especially in the switching sequence of SVM. Various SVM sequences are investigated in [9] and its application in AC drives [10]. Space vector approaches with a higher number of degrees of freedom have been presented in [11]. Different vector sequences based on the value of the modulation index is employed in [12]. The application of SVM to a variable speed electric drive is studied in [13].

IV. CONCLUSION

This paper reviews different motor controllers using a second order motor load, typical of many loads like a fan or industrial pump. The progress of Electric drives in the last few years with increasing use of a standard, modular, integrated approach have led to increased penetration of power conversion equipment. The observation that power electronics technology may be becoming a mature field for Electrical machines after a lifetime of at least a century as many other technologies before it was an important motivation to critically examine the present state of the art and possible future development. In this process it is important to understand historically how and when the original driving philosophy for the spectacular development of power electronics technology has come about. On the evaluation of the state of the art, it can indeed be concluded that the historical development of Induction motor drives is approaching the limits of the most important internal metrics of the technology in its present form. This is a definite indication of maturity in the internal development process. It is also clear, however, that upcoming maturity only applies to the internal constituent technologies of power semiconductor switch technology and power electronic switching network technology. It is expected that converter, control, and machine will eventually be integrated as an intelligent machine of the future, particularly in the lower end of power rating.

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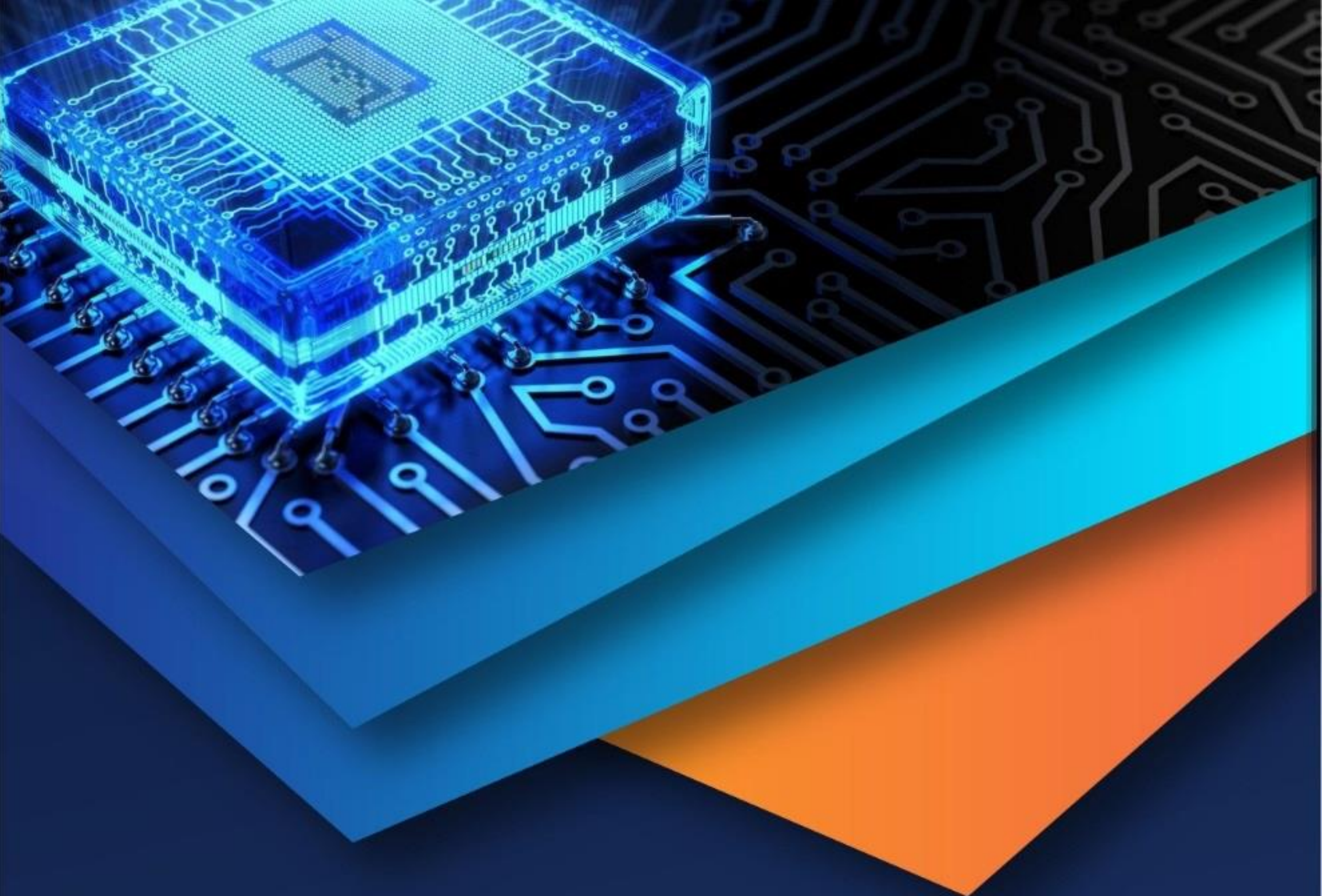

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