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Performance of PMSM Using SVPWM Technique for 2-Level Inverter

K. Lavanya[1], T. Sandeep Varma[2], Y. Manoj Maheshwar[3], P. Gowri[4]
[1, 2, 3, 4]*Department of Electrical and Electronics engineering, ANITS*

Abstract: This paper presents SVPWM technique for 2-level inverter used to control PMSM. Mathematical model of PMSM is done in abc frame and $\alpha - \beta$ frame. Various controllers can be used to control PMSM to analyze the performance of PMSM. In this paper conventional controllers such as PI, PID are used and performance characteristics of PMSM is obtained. SVPWM technique is applied to control the output of 2 level inverter which is to be applied to PMSM drive.
Keywords: SVPWM TWO LEVEL, THD, PMSM, PI, PID

I. INTRODUCTION

Permanent magnet synchronous motors are widely used in high performance drives such as industrial robots and machine tools. In recent years, the magnetic and thermal capabilities of the Permanent Magnet Synchronous Motors have been considerably increased by employing the high-coercive permanent magnet material. The speed control of synchronous motor depends upon two factors viz number of poles, P and supply frequency, f. as in case of shipping propulsion, the speed of the motor can be changed by changing the speed of the alternator – the speed of the motor changes exactly in the same proportion as that of the alternator supplying power to it. It is to be noted here that the voltage and frequency are directly proportional to the speed at which alternator is driven.The effective way of producing the variable speed Permanent Magnet Synchronous Motor drive is to supply the motor with variable voltage and variable frequency or constant V/f supply variable frequency is required because the rotor speed is directly proportional to the stator supply frequency. A variable voltage is required because the motor impedance is reduced at lower frequencies and consequently the current has to be limited by means of reducing the supply voltage. Unlike a DC motors, Permanent magnet synchronous motors (PMSM) are very popular in a wide range of applications. The PMSM does not have a Commutator, which makes it more reliable than a DC motor. The PMSM also has advantages when compared to an AC induction motor. The PMSM generates the rotor magnetic flux with rotor magnets, achieving higher efficiency. Therefore, the PMSM is used in applications that require high reliability and efficiency.

II. PMSM Drive Construction

In an electric motor the moving part is the rotor which turns the shaft to deliver the mechanical power. The rotor usually has conductors laid into it which carry currents that interact with the magnetic field of the stator to generate the forces that turn the shaft. However, some rotors carry permanent magnets, and the stator holds the conductors.

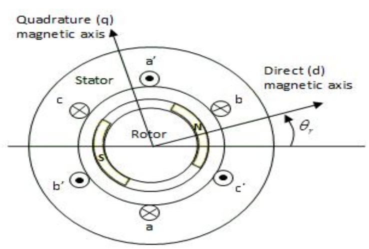

Fig1 Motor Construction with a Single Pole-Pair on the Rotor

The Permanent magnet motors have permanent magnets embedded in the steel rotor to create a constant magnetic field. At synchronous speed these poles lock to the rotating magnetic field. They are not self-starting. Because of the constant magnetic field in the rotor these cannot use induction windings for starting, and must have electronically controlled variable frequency stator drive. Fig1 shows the motor construction with a single pole-pair on the rotor. The rotor magnetic field due to the permanent magnet(s) creates a sinusoidal rate of change of flux with rotor angle. For the axes convention in the preceding Fig.1, the a-phase and permanent magnet fluxes are aligned when rotor angle $\theta_r$is zero. When there is a large number of coils , instead of the mmf wave being stepped, a smooth traction may be assumed, and the mmf diagram becomes triangular. If the conductors only occupy a portion of the armature surface, the mmf wave will be trapezoidal. Such mmf wave is found with the rotor winding of a smooth rotor alternator.

A. Control Aspectsof PMSM Drives

Historically several controllers have been developed for the control of PMSM drives such as

1) *Scalar Control:* Despite the fact that "Voltage-Frequency" (V/f) is simplest controller, it is the most widespread method, being in the majority of the industrial applications. It is known as a scalar control and acts by imposing a constant relation between voltage and frequency. The structure is simple and it is normally used without feedback. However, this controller does not achieve a good accuracy in both speed and torque responses, mainly due to the fact that the stator flux and torque are inherently coupled and cannot be controlled independently.
2) *Vector Control:* In these types of controller, there are control loop for controlling both the torque and flux. The most widespread controllers of this type are the ones that use vector transform such as PARK. The main disadvantages are the huge computational capability required and the compulsory good identification of the motor parameters.
3) *Direct Torque Control:* This method has emerged over the last one and half decade to becomes one possible alternative to the well-known vector control of PMSM. Its main characteristic is good performance, obtaining results as good as classical vector control with several advantages based on its simpler structure and control diagrams. However, DTC suffers with variable switching frequency and flux, torque ripples

III. MATHEMATICAL MODELLING OF PMSM

In a motor with more than one pair of magnetic poles the electric angle differ the mechanical. Their relationship is

$$\theta_r = \frac{p}{2}\theta_m$$

The voltage V, over each stator winding is the sum of the resistive voltage drop and the voltage induced from the time varying flux linkages, $d\psi / dt$.

$$V_a = r_a i_a + \frac{d}{dt}\psi_a$$

$$V_b = r_b i_b + \frac{d}{dt}\psi_b$$

$$V_c = r_c i_c + \frac{d}{dt}\psi_c$$

The stator windings are wound with the same number of turns so the resistance is equal in all three windings,

$$r_a = r_b = r_c = r_s$$

In matrix form these voltage equations (3.2) to (3.4) becomes

$$V_{abc} = r_s i_{abc} + \frac{d}{dt}\psi_{abc}$$

$$= \begin{bmatrix} r_s & 0 & 0 \\ 0 & r_s & 0 \\ 0 & 0 & r_s \end{bmatrix}\begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} + \frac{d}{dt}\begin{bmatrix} \psi_a \\ \psi_b \\ \psi_c \end{bmatrix}$$

Flux linkages in a linear magnetic circuit is the product of inductance and current, the motor model was assumed linear, which is a fairly accurate approximation if saturation does not occur, hence

$$\psi_{abc} = L_s i_{abc} + \psi_m$$

$$L_s = \begin{bmatrix} L_{aa} & L_{ab} & L_{ac} \\ L_{ba} & L_{bb} & L_{bc} \\ L_{ca} & L_{cb} & L_{cc} \end{bmatrix}$$

The diagonal elements in the inductance matrix L_s are self inductances and the off diagonal elements are mutual inductances. The matrix is symmetric because the flux coupling between two windings is equal in both directions. A current in stator windings gives

rise to a leakage flux and a magnetizing flux. The magnetizing flux is confined to the air-gap and give rise to the rotating MMF wave. Leakage flux is assumed to only affect its own winding. In a magnetically linear circuit flowing in the winding with all currents set to zero.Let the self inductance be $L_{aa} = L_{ls} + L_m$ where L_{ls} is the leakage inductances .
The minimum value of L_{aa} occurs at $\theta_r = 0, \pi, 2\pi \cdots\cdots$ and maximum values at $\theta_r = \pi/2, 3\pi/2, 5\pi/2 \cdots\cdots$
Assume $L_{aa}(\theta_r)$ varies sinusoidally, then $L_{aa} = L_{ls} + L - L_\Delta \cos(2\theta_r)$

$$L_{bb} = L_{ls} + L - L_\Delta \cos(2\theta_r + 2\pi/3)$$

$$L_{cc} = L_{ls} + L - L_\Delta \cos(2\theta_r - 2\pi/3)\ L_{ab} = -\frac{L}{2} - L_\Delta \cos\left(2\theta_r - \frac{2\pi}{3}\right)$$

$L_{ac} = -\frac{L}{2} - L_\Delta \cos\left(2\theta_r + \frac{2\pi}{3}\right)$ & $L_{bc} = -\frac{L}{2} - L_\Delta \cos(2\theta_r)$

$$L_{aa}(\theta_r) = \begin{bmatrix} L_{ls} + L - L_\Delta \cos(2\theta_r) & -\frac{L}{2} - L_\Delta \cos(2\theta_r - 2\pi/3) & -\frac{L}{2} - L_\Delta \cos(2\theta_r + 2\pi/3) \\ -\frac{L}{2} - L_\Delta \cos(2\theta_r - 2\pi/3) & L_{ls} + L - L_\Delta \cos(2\theta_r + 2\pi/3) & -\frac{L}{2} - L_\Delta \cos(2\theta_r) \\ -\frac{L}{2} - L_\Delta \cos(2\theta_r + 2\pi/3) & -\frac{L}{2} - L_\Delta \cos(2\theta_r) & L_{ls} + L - L_\Delta \cos(2\theta_r + 2\pi/3) \end{bmatrix} \psi_m$$

$$= |\psi_m| \begin{bmatrix} \sin(\theta_r) \\ \sin\left(\theta_r - \frac{2\pi}{3}\right) \\ \sin\left(\theta_r + \frac{2\pi}{3}\right) \end{bmatrix}$$

A. Transformation to qdo Frame

Now there is a set of equations which describe the motor. These equations though depend on rotor position and make the equation system quite involved to solve. If the variables are transformed into reference frame attached to the rotor, reluctance which in the ABC frame depends on rotor position will be constant. The transformation used is called the park transform

Let K_s be the park transform matrix then

$$\begin{bmatrix} S_q \\ S_d \\ S_o \end{bmatrix} = K_s \begin{bmatrix} S_a \\ S_b \\ S_c \end{bmatrix}$$

$$K_s = 2/3 \begin{bmatrix} \cos(\theta) & \sin(\theta) & 1 \\ \cos(\theta_r - 2\pi/3) & \sin(\theta_r - 2\pi/3) & 1 \\ \cos(\theta_r + 2\pi/3) & \sin(\theta_r + 2\pi/3) & 1 \end{bmatrix}$$

$$K_s^{-1} = (KA)^{-1} = K^{-1}A^{-1}$$

$|A| = \cos\left(\theta - \frac{2\pi}{3}\right)\sin\left(\theta + \frac{2\pi}{3}\right) - \sin\left(\theta - \frac{2\pi}{3}\right)\cos\left(\theta + \frac{2\pi}{3}\right) - \left\{\cos\theta \sin\left(\theta + \frac{2\pi}{3}\right) - \sin\theta\cos\left(\theta + \frac{2\pi}{3}\right)\right\} + \left\{\cos\theta\sin\left(\theta - \frac{2\pi}{3}\right) - \sin\theta\cos\left(\theta - \frac{2\pi}{3}\right)\right\}$

$$= \sin\left(\theta + \frac{2\pi}{3} - \theta + \frac{2\pi}{3}\right) + \sin\left(\theta - \theta - \frac{2\pi}{3}\right) + \sin\left(\theta - \frac{2\pi}{3} - \theta\right)$$

$$= \sin\left(\frac{4\pi}{3}\right) + \sin\left(-\frac{2\pi}{3}\right) + \sin\left(-\frac{2\pi}{3}\right)$$

$$= -\frac{\sqrt{3}}{2} - \frac{\sqrt{3}}{2} - \frac{\sqrt{3}}{2}$$

$$K_s^{-1} = 2/3 \begin{bmatrix} \cos(\theta) & \cos(\theta_r - 2\pi/3) & \cos(\theta_r + 2\pi/3) \\ \sin(\theta) & \sin(\theta_r - 2\pi/3) & \sin(\theta_r + 2\pi/3) \\ 1/2 & 1/2 & 1/2 \end{bmatrix} \quad (3.23)$$

In K_s there is a factor of $2/3$ in front of the matrix. This factor can be understood by discussion about mmf wave. For a balanced set, of say voltages the resultant voltage vector has amplitude $3/2$ times that of the individual amplitude. The factor $2/3$ makes the amplitude of quantities expressed in the qdo reference frame correspond to that of each individual phase in the stator abc frame. The last row in K_s is the zero sequence. Another feature that may be noted with the above definition of the park transform, it is not power invariant. This is because $|K_s| \neq 1$.

$$P_{\mathrm{qdo}} = (v_{\mathrm{qdo}}, \quad i_{\mathrm{qdo}}) = W_1 v_q i_q + W_2 v_d i_d + W_3 v_o i_o$$

$$P_{\mathrm{abc}} = v_{abc}{}^T i_{abc} = v_{abc} i_{abc}{}^T$$

$$P_{\mathrm{abc}} = v_{abc}{}^T i_{abc} = (K_s{}^{-1} v_{\mathrm{qdo}})^T K_s{}^{-1} i_{\mathrm{qdo}} = v_{\mathrm{qdo}}{}^T K_s{}^{-T} K_s{}^{-1} i_{\mathrm{qdo}} = v_{\mathrm{qdo}}{}^T \begin{bmatrix} 3/2 & 0 & 0 \\ 0 & 3/2 & 0 \\ 0 & 0 & 3/2 \end{bmatrix} i_{\mathrm{qdo}}$$

$$= 3/2 (v_q i_q + v_d i_d + 2 v_o i_o) = P_{\mathrm{qdo}}$$

$$P_{\mathrm{qdo}} = 3/2 (v_q i_q + v_d i_d + 2 v_o i_o)$$

We are now going to transform v_{abc}, first to an arbitrary qdo reference frame and then let this transformation be attached to the rotor. Express v_{abc} in qdo variables

$$v_{abc} = r_s i_{abc} + \frac{d}{dt} \Psi_{abc}$$

$$= r_s K_s{}^{-1} i_{\mathrm{qdo}} + \frac{d}{dt} K_s{}^{-1} \Psi_{qdo}$$

$v_{\mathrm{qdo}} = K_s r_s K_s{}^{-1} i_{\mathrm{qdo}} + \frac{d}{dt} K_s{}^{-1} \Psi_{qdo}$ The resistance does not change when transformed since

$$K_s r_s K_s{}^{-1} = r_s K_s K_s{}^{-1} = 1 * r_s = r_s$$

$$K_s \frac{d}{dt}\left(K_s{}^{-1} \Psi_{qdo}\right) = K_s \left(\frac{d}{dt} K_s(\theta_T)\right) \Psi_{qdo} + K_s{}^{-1} \frac{d}{dt} \Psi_{qdo}(\theta_T, \theta_r)$$

$$K_s(\theta_T) \frac{d}{dt} K_s(\theta_T) = \omega_T \begin{bmatrix} 0 & 1 & 0 \\ -1 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix}$$

$$K_s \frac{d}{dt}\left(K_s{}^{-1} \Psi_{qdo}\right) = \omega_T \begin{bmatrix} \Psi_d \\ -\Psi_q \\ 0 \end{bmatrix} + \frac{d}{dt} \Psi_{qdo}$$

Voltage equation in the arbitrary frame

$$v_{\mathrm{qdo}} = r_s i_{\mathrm{qdo}} + \omega_T \begin{bmatrix} \Psi_d \\ -\Psi_q \\ 0 \end{bmatrix} + \frac{d}{dt} \Psi_{qdo}$$

Now let us express flux in component form

First expand Ψ_{abc}

$$\Psi_{abc} = L_s i_{abc} + = L_s K_s{}^{-1} i_{\mathrm{qdo}} + \Psi_m$$

$$\Psi_{qdo} = K_s L_s K_s{}^{-1} i_{\mathrm{qdo}} + K_s \omega_m$$

$$K_s L_s(\theta_r) K_s{}^{-1} =$$

$$\begin{bmatrix} L_{ls} + 3/2 (L - L_\Delta \cos(2\theta_T - 2\theta_r)) & -3/2 L_\Delta \sin(2\theta_T - 2\theta_r) & 0 \\ -3/2 L_\Delta \sin(2\theta_T - 2\theta_r) & L_{ls} + 3/2 (L - L_\Delta \cos(2\theta_T - 2\theta_r)) & 0 \\ 0 & 0 & L_{ls} \end{bmatrix}$$

$$K_s(\theta_T) \omega_m(\theta_r) = \omega_m \begin{bmatrix} -\sin(\theta_T - \theta_r) \\ \cos(\theta_T - \theta_r) \\ 0 \end{bmatrix}$$

$$L_{mq} = 3/2 (L - L_\Delta)$$

$$L_{md} = 3/2 (L + L_\Delta)$$

$$\psi_{qdo} = \begin{bmatrix} \frac{L_d+L_q}{2} - \frac{L_d-L_q}{2}\cos(2\theta_T-2\theta_r) & -\frac{L_d-L_q}{2}\sin(2\theta_T-2\theta_r) & 0 \\ -\frac{L_d-L_q}{2}\sin(2\theta_T-2\theta_r) & \frac{L_d+L_q}{2} + \frac{L_d-L_q}{2}\cos(2\theta_T-2\theta_r) & 0 \\ 0 & 0 & L_{ls} \end{bmatrix} \begin{bmatrix} i_d \\ i_q \\ i_o \end{bmatrix} + \omega_m \begin{bmatrix} -\sin(\theta_T-\theta_r) \\ \cos(\theta_T-\theta_r) \\ 0 \end{bmatrix}$$

Where $L_q = L_{ls} + L_{mq}$ and $L_d = L_{ls} + L_{md}$. θ_T is the angle which rotates the transformed reference frame and θ_r the rotor position in electrical radians. This is the flux expressed in the arbitrary reference frame.If the reference frame rotates in synchronism with the rotor and both angles have the same initial conditions then, and total flux in the rotor reference becomes

$$\psi_{qdo}^{\ r} = \begin{bmatrix} L_q & 0 & 0 \\ 0 & L_d & 0 \\ 0 & 0 & L_{ls} \end{bmatrix} \begin{bmatrix} i_q^{\ r} \\ i_d^{\ r} \\ i_o^{\ r} \end{bmatrix} + \begin{bmatrix} 0 \\ \psi_m \\ 0 \end{bmatrix}$$

Stator voltage expressed in the rotor qdo frame then is

$$v_{qdo}^{\ r} = r_s i_{qdo}^{\ r} + \omega_m \begin{bmatrix} \psi_d \\ -\psi_q \\ 0 \end{bmatrix} + \frac{d}{dt}\psi_{qdo}^{\ r}$$

$$v_q^{\ r} = (r_s + \rho L_q) i_q^{\ r} + \omega_r L_d i_q^{\ r} + \omega_r L_m$$

$$v_d^{\ r} = (r_s + \rho L_d) i_d^{\ r} - \omega_r L_q i_q^{\ r}$$

$$v_o^{\ r} = (r_s + \rho L_{ls}) i_o^{\ r}$$

Where $\rho = d/dt$

$$\therefore T_e = \frac{dW_e}{d\theta_m} = \frac{P}{2}\left(\frac{dW_e}{d\theta_r}\right)$$

$$= \frac{P}{2}\left(\frac{1}{2} i_{abc}^{\ T} \frac{d}{d\theta_r} L_s i_{abc} + i_{abc}^{\ T} \frac{d}{d\theta_r}\psi_m\right)$$

By substituting various variables, we have

$$T_e = \frac{P}{2}\left(\frac{9}{4} L_\Delta i_{qdo}^{\ T}\right) \begin{bmatrix} -\sin(2\theta_T-2\theta_r) & -\cos(2\theta_T-2\theta_r) & 0 \\ \cos(2\theta_T-2\theta_r) & \sin(2\theta_T-2\theta_r) & 0 \\ 0 & 0 & 0 \end{bmatrix} i_{qdo} + \frac{3}{2}\psi_m i_{qdo}^{\ T}\omega_m \begin{bmatrix} \cos(\theta_T-\theta_r) \\ \sin(\theta_T-\theta_r) \\ 0 \end{bmatrix}$$

$$\therefore T_e = \frac{3}{2}\frac{P}{2}\left(\psi_d i_q - \psi_q i_d\right)$$

$$\psi_d = L_d i_d + T_e = \frac{3}{2}\frac{P}{2}\left((L_d - L_q) i_q^{\ r} i_d^{\ r} + \psi_m i_q^{\ r}\right)$$

In the above equation the first term is due to reluctance variations and disappears in a salient free machine. The second term is due to the permanent flux. These equations describe the electro mechanical behavior of the machine in the qdo reference frame. From above equations , can be rewritten as follows

$$\psi_d = \frac{1}{s}\left(v_d + w_r\psi_q - i_d r_s\right)$$

$$\psi_q = \frac{1}{s}\left(v_q + w_r\psi_d - (i_d r_s)\right)$$

IV. TWO LEVEL SVPWM INVERTER

In three phase two level inverter, each arm contains of two IGBT'S and two anti parallel diodes. Each IGBT'S simply considered as switches. Each pole in a two level inverter can assume two values namely 0 &Vdc. S1 to S6 are the six power switches that shape the output, which are controlled by the switching variable a, a', b, b', c and c'. When an upper transistor is switched on, i.e., when a, b or c is 1, the corresponding lower transistor is switched on, i.e., the corresponding a', b' or c' is zero .Therefore, the on and off states of the transistors can be used to determine the output voltage. In this PWM technique180^0 conduction is used for generating the gating signals. If two switches: one upper and one lower switch conduct at the same time such that the output voltage is ± Vs. the switch state is 1. If these two switches are off at the same time, the switch state is 0. (S1, S4), (S3, S6), (S5, S2) are switch pairs. These are shifted each other by 180^0 . For example S1conducts at 0^0, S4 conducts at 180^0. Upper switches S1, S3, S5 are displaced by 120^0.That is S1conducts at 0^0, S3 conducts at 120^0, and S5 conducts at 240^0. Similarly lower switches S4, S2, S6 are displaced by 120^0. That is S4conducts at 0^0, S2 conducts at 120^0, and S6conducts at 240^0. In any phase leg of inverter switches cannot turn on simultaneously, that would result short circuit across the dc link voltage. To avoid this switches are turned on complementally. Similarly, in order to avoid undefined states in the VSI, and thus undefined ac output line voltages, the switches of any leg of the inverter cannot be switched off simultaneously as this will result in voltages that will depend upon the respective line current polarity. Of the eight valid states, two of them produce zero ac line voltages. In this case, the ac line currents freewheel through either the upper or lower components. The remaining states produce non-zero ac output voltages.

1) No of switching states & selection of switching states.
2) No of space vectors.
3) Determination of location of space vectors.
4) Sector identification.
5) Calculation of active vectors switching time periods.
6) Generation of gating signals for the individual power devices.
7) Determination of switching sequence for the individual sectors.

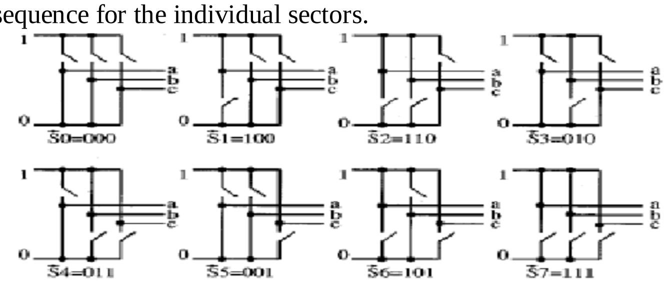

Figure: 2 Switching states of two level inverter

STATE	ON	OFF
0	S4,S6,S2	S1,S3,S5
1	S1,S2,S6	S4,S5,S3
2	S2,S3,S1	S5,S6,S4
3	S3,S4,S2	S6,S1,S5
4	S4,S5,S3	S1,S2,S6
5	S5,S6,S4	S2,S3,S1
6	S6,S1,S5	S3,S4,S2
7	S1,S3,S5	S4,S6,S2

Table1. Functioning of switches

A. Space Vector Diagram Of Two-Level Inverter

Space vector diagram is divided into six sectors. The duration of each sector is 600. V1, V2, V3, V4, V5, V6 are active voltage vectors and V0 & V7 are zero voltage vectors. Zero vectors are placed at origin. The lengths of vectors V1 to V6 are unity and lengths of V0 and V7 are zero.

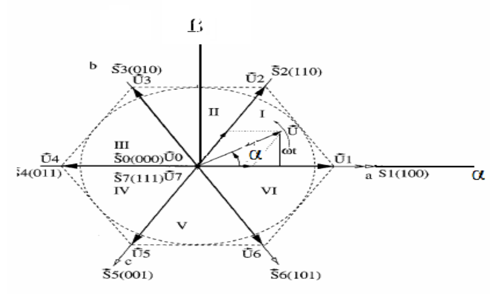

Figure.3. Space vector diagram of two level inverter

The space vector Vs constituted by the pole voltage Vao, Vbo, and Vco is defined as [4]
Vs = Vao + Vbo e j (2π/3) + Vco e j (4 Π/3)
Vao=Van + Vno , Vbo=Vbn + Vno and Vco=Vcn+ Vno
Van+Vbn+Vcn=0 & Vno= (Vao+Vbo+Vco) /3
Vab = Vao – Vbo , Vbc = Vbo – Vco , Vca = Vco – Vao
FOR example voltage vector V1 that is 100
Vao=Vdc, Vbo=0and Vco=0, then Vn= (Vdc+0+0)/3=Vdc/3)
Van=Vao-Vno = (2/3) Vdc, Vbn=Vbo-Vno= (-1/3) Vdc & Vcn=Vco-Vno= (-1/3) Vdc
Vab=Vao-Vbo=Vdc , Vbc=Vbo-Vco=0 & Vca=Vco-Vao=-Vdc
V α = 2/3 (Va – 1/2 Vb – 1/2 Vc) & V β =1/√3 V b – 1/√3 VC

Voltage Vectors	Switching Vectors			Line to neutral voltage			Line to line voltage		
	a	b	c	V_{an}	V_{bn}	V_{cn}	V_{ab}	V_{bc}	V_{ca}
V_0	0	0	0	0	0	0	0	0	0
V_1	1	0	0	2/3	-1/3	-1/3	1	0	-1
V_2	1	1	0	1/3	1/3	-2/3	0	1	-1
V_3	0	1	0	-1/3	2/3	-1/3	-1	1	0
V_4	0	1	1	-2/3	1/3	1/3	-1	0	1
V_5	0	0	1	-1/3	-1/3	2/3	0	-1	1
V_6	1	0	1	1/3	-2/3	1/3	1	-1	0
V_7	1	1	1	0	0	0	0	0	0

(Note that respective voltage should be multiplied by V_{dc})
Table.2.switching vectors, phase voltages and output line to line voltages

B. *Calculation Of Active Vector Switching time Periods*

In a two level inverter, on time calculation is based on the location of reference vector with in a sector.In one sampling interval, the output voltage vector V can be written as

V=T0/TS * V0 +T1 /TS* V1 +----T1/TS *V7

or example for Sector 1

V=T0/TS*V0+T1/TS*V1+T2/TS*V2+T7/TS*V7 (1)

Lengths of vectors V0 & V7 are zero.

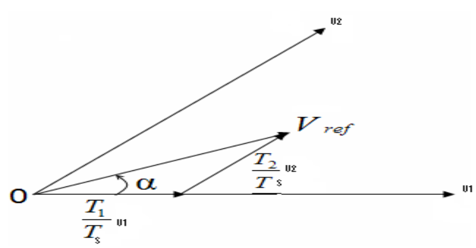

Figure 4.Reference vector as a combination of adjacent vectors at sector 1

Along the α axis

V = V ref cos α

V1 = V dc cos 0 & V2 = V dc cos 60

Along the β axis

V = V ref sin α

V1 = V dc cos 90 & V2 = V dc cos (90-60) = V dc sin 60

Vref cos α = V dc ×*T1/Ts + (V dc cos 60) *T2 / TS -

V ref sin α = (V dc sin60) × T2 / TS

T1 = (2/√3) (Ts/ Vdc) V ref sin (Π/3- α)

T2= (2/√3) (Ts/ Vdc) V ref sinα

V. CONTROLLING PART

The design of the speed-controller is important from the point of view of imparting desired transient and steady-state characteristics to the speed-controlled PMSM drive system. A proportional-plus-integral controller is sufficient for many industrial applications; hence, it is considered in this section. Selection of the gain and time constants of such a controller by using the symmetric-optimum principle is straightforward if the d axis stator current is assumed to be zero.

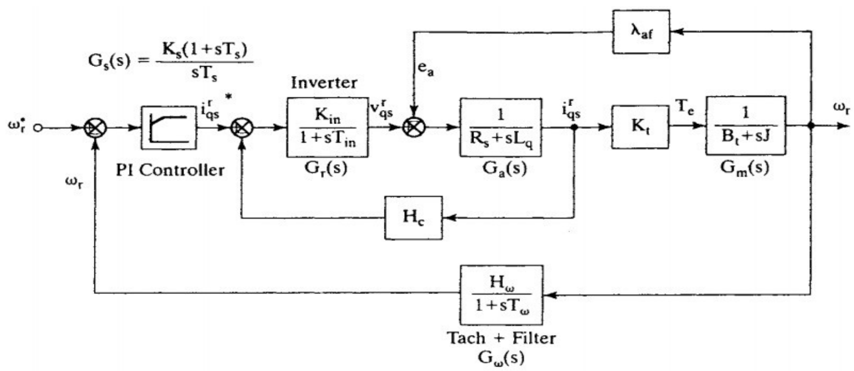

Figure:5 Block Diagram of the speed controlled PMSM Drive

In the presence of a d axis stator current, the *d* and *q* current channels are cross-coupled, and the model is nonlinear, as a result of the torque term. Under the assumption that $i_{qs}{}^{r} = 0$, the system becomes linear and resembles that of a separately-excited dc motor with constant excitation. From then on, the block-diagram derivation, current-loop approximation, speed-loop approximation, and derivation of the speed-controller by using symmetric optimum are identical to those for a de or vector-controlled induction-motor-drive speed-controller design.

A. Proportional Controller

The proportional term makes a change to the output that is proportional to the current error value. The proportional response can be adjusted by multiplying the error by a constantK_p, called the proportional gain. The transfer function of a proportional controller is simply a gain sayK_p. If the input of the controller is $e(t)$ then the output is $u(t) = K_p e(t)$ or in a Laplace transform domain $U(s) = K_P E(s)$. As K_P increases the unit-step response may becomes faster and eventually the feedback system may becomes unstable. For the same unit-step reference input the steady-state plant outputs are different for different K_P.

B. Proportional Action

Proportional action provides an instantaneous response to the control error. This is useful for improving the response of a stable system but cannot control an unstable system with a nonzero steady-state error. By using this controller rise time increases and also steady state error decreases. And peak overshoots increases. This will be done only by proper selection of K value.

C. Integral Controller

In this controller the output u(t) is altered at a rate proportional to the error signal e(t). The output u(t) depends upon the integral of the error signal e(t).

Mathematically

$\frac{du(t)}{dt} = K.e(t)$ Or $u(t) = K\int_0^1 e(t)dt$ Or U(s) =$\frac{KE(s)}{s}$

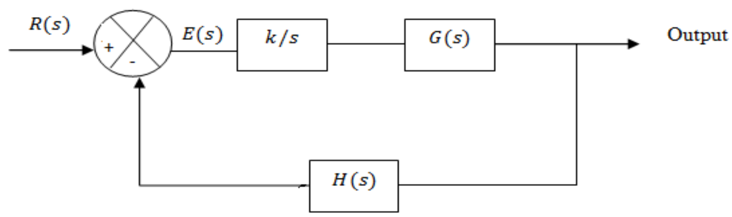

Figure:6 Block Diagram of Integral controller

D. Proportional Plus Integral Control

Integral control action itself is not sufficient, as it introduces hunting in the system. Therefore a combination of Proportional and integral control action is introduced to improve the system performance. In this type of system, the actuating signal consists of proportional error signal added with the integral of the error signal.

$u(t)$ = e(t)=K $\int_0^t e(t)dt$ = integral of error signal Or U(S) = E(s) [1+$\frac{K}{S}$]

Time response

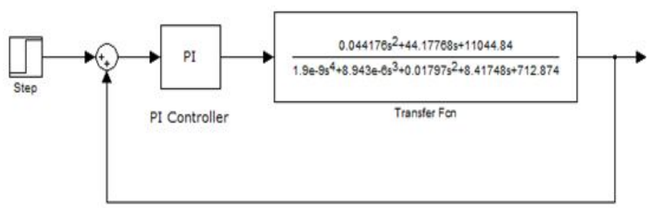

Figure:7 Block diagram of PMSM drive using PI controller

E. PID controller

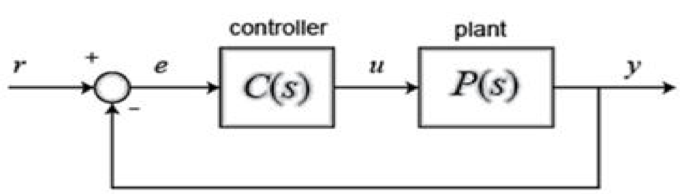

Figure:8 Block Diagram of Controller and Plant

The output of a PID controller, equal to the control input to the plant, in the time-domain is as follows: $u(t) = K_p e(t) + K_i \int e(t)dt + K_p \frac{de}{dt}$

The new output (y) is then fed back and compared to the reference to find the new error signal(e). The transfer function of a PID controller is found by taking the Laplace transform $K_p + \frac{K_i}{s} + K_d s = \frac{K_d s^2 + K_p s + K_i}{s}$

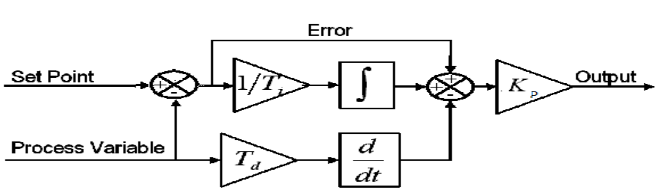

Figure:9 Block diagram of a basic PID controller

VI. SIMULINK MODEL OFSVPWM-PMSM

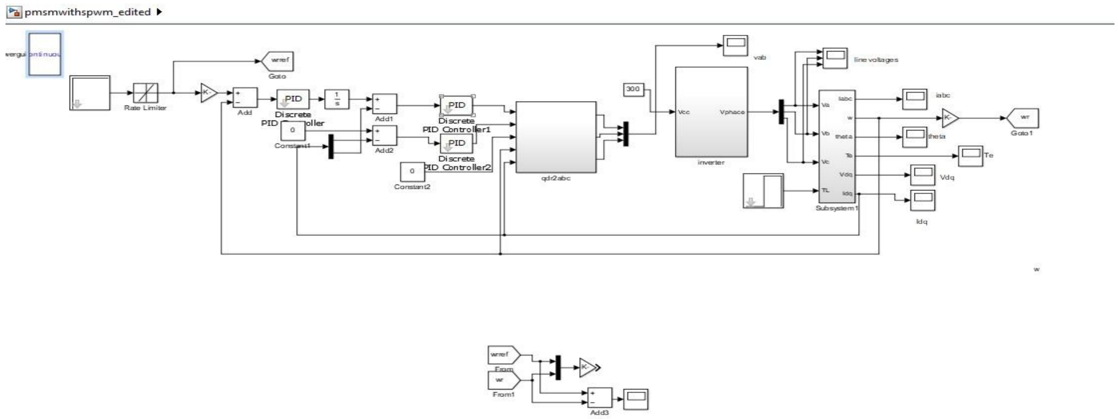

Figure:10 Simulink model

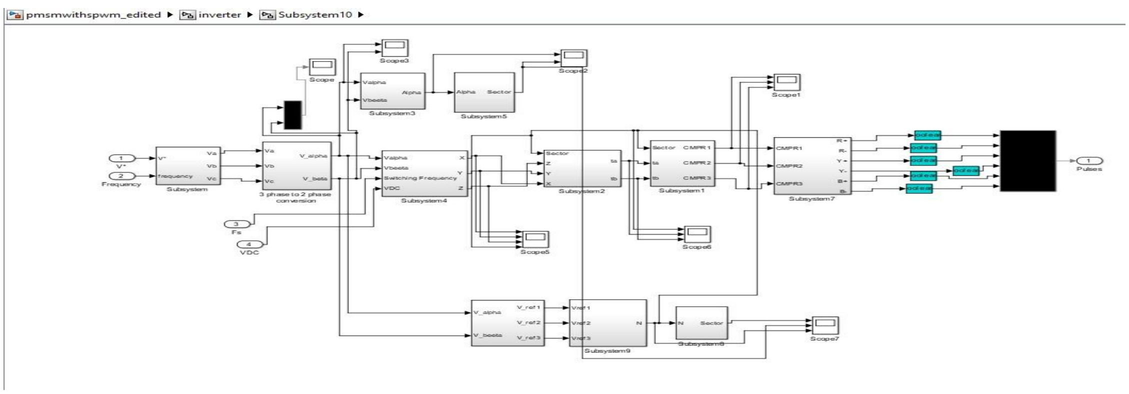

Figure: 11SVPWM MODEL:

VII.SIMULINK RESULTS

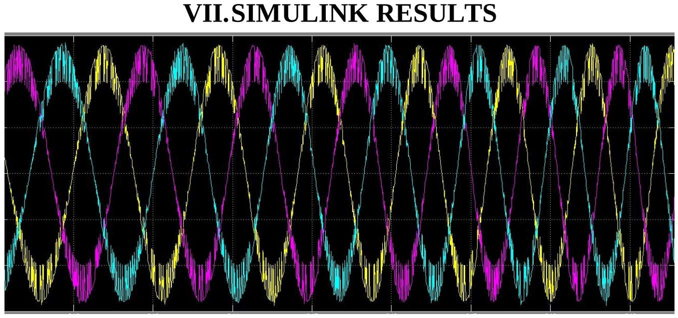

Figure:12 Waveforms of three phase current- Iabc

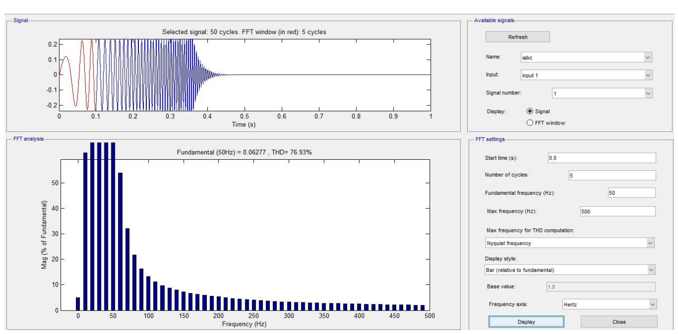

Figure:13 THD of Iabc

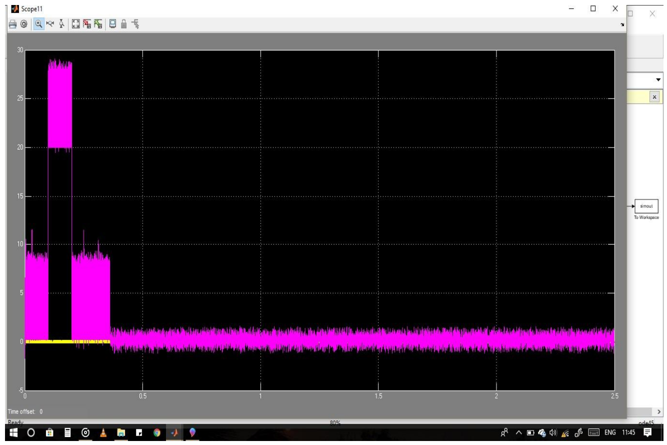

Figure:14 Direct axis, quadrature axis current- Idq

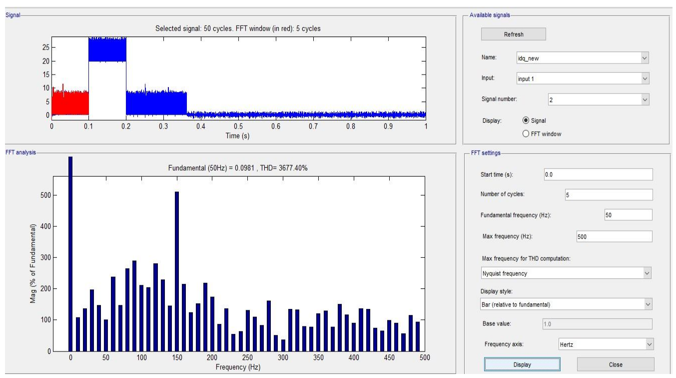

Figure:15 THD of Idq

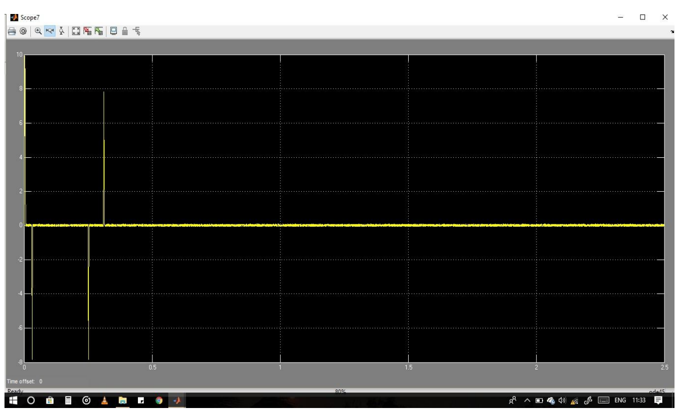

Figure:16 Speed variation

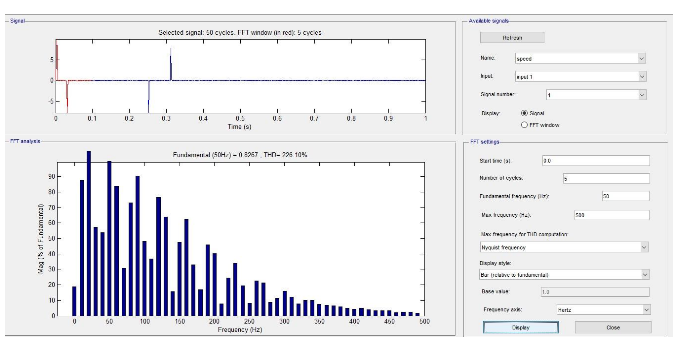

Figure:17 THD of speed

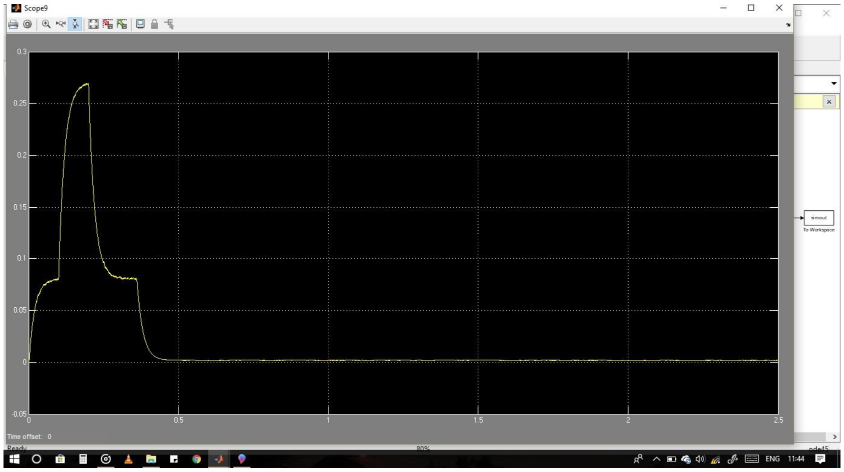

Figure:18PMSM torque

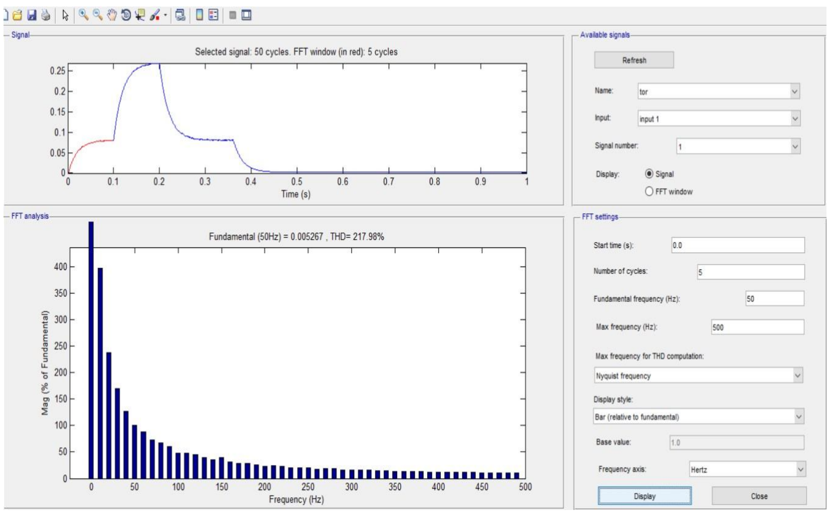

Figure:19 THD of torque

VIII. CONCLUSION

In this paper, the performance of PMSM which is controlled by a 2-level inverter using SVPWM technique is studied. Mathematical modelling of PMSM is observed. The control action of PI, PID controllers and their usage in controlling a PMSM is determined. The harmonics in a SVPWM-PMSM drive are observed using Matlab software. Total Harmonic Distortion is calculated and concluded that SVPWM technique when employed reduces the harmonic distortion. This paper work provides successful implementation of SVPWM technique used to control a PMSM drive.

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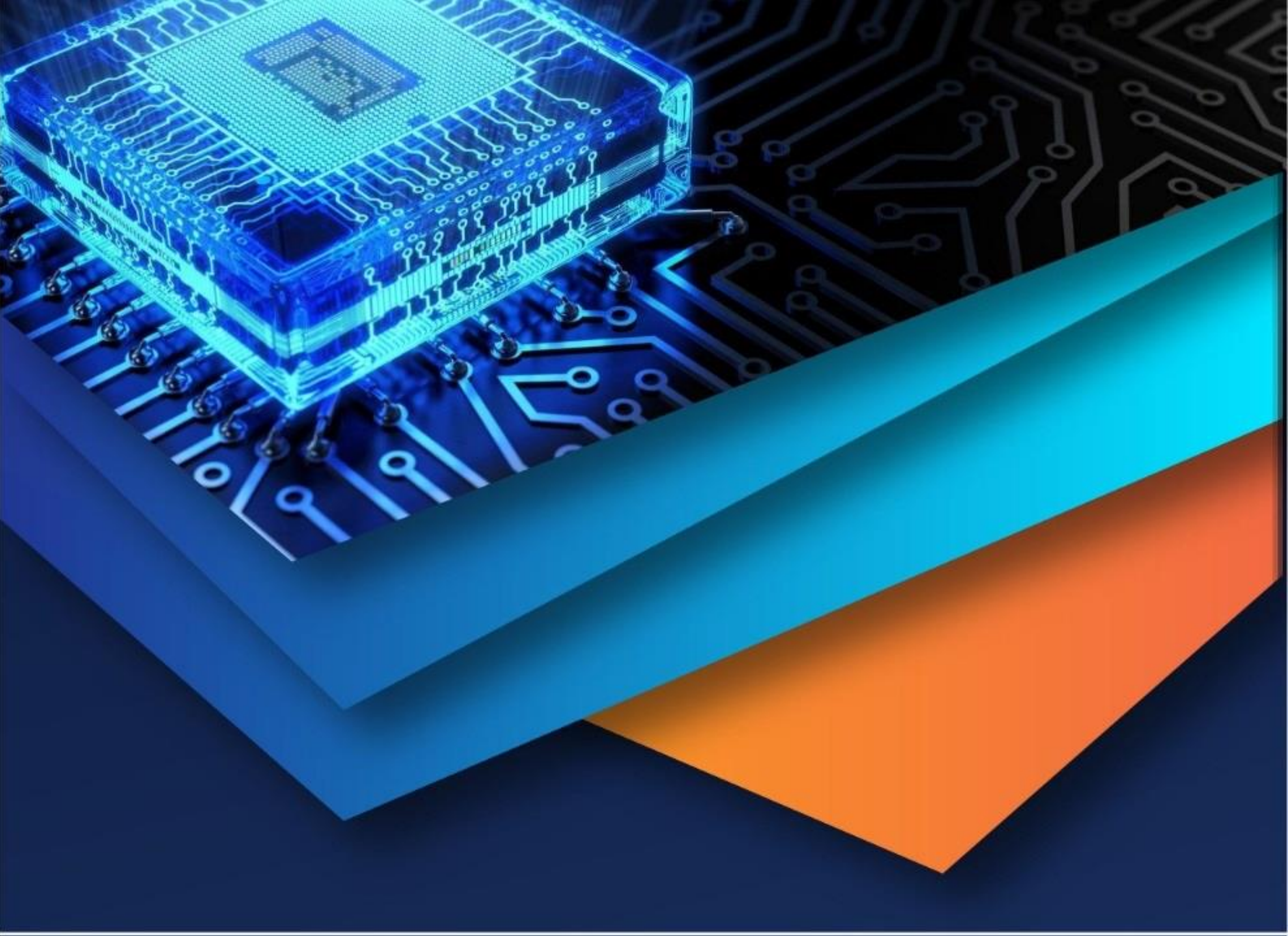
AUTHOR'S BIOGRAPHY

Lavanya Komma Received her Master degree in 2010 in Power Electronic from JNTU University (AURORA Eng College), Hyderabad, India, and Bachelor degree in 2005 in Electrical & Electronics Engineering from TPIST, Bobbili, India. She is pursuing her PhD in Power Electronics at JNTUniversity,Kakinada She had 10 years teaching experience. She is currently working as an Assistant Professor of Electrical and Electronics Engineering Department at ANITS, Visakhapatnam, India.
Mail id:lavanyagumpena1008@gmail.com
Sandeep Varma Tirumalarajupursing Bachelor degree (2014-2018) in E.E.E from Andhra University(ANITS Eng College), Visakapatnam, India. Mail id:svarma459@gmail.com

ManojMaheshwarYettharla pursing Bachelor degree (2014-2018) in E.E.E from Andhra University(ANITS Eng College), Visakapatnam, India.Mail id: manojyetcharla@gmail.com

Gowri Pentapallipursing Bachelor degree (2014-2018) in E.E.E from Andhra University(ANITS Eng College), Visakapatnam, India. Mail id: gowri17031997@gmail.com



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