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Chapter

FUZZY LOGIC BASED ENCODER-LESS SPEED CONTROL OF PMSM FOR HUB MOTOR DRIVE

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ABSTRACT

Today the interest in electric vehicles (EVs) is high due to their ability to reduce the fuel consumption. By employing an electric motor with high power density, which can be directly mounted inside the wheel, known as in-wheel motors, or hub motor the transmission losses of EVs are minimized and operating efficiency are improved. Amongst different type of electrical motor, permanent-magnet synchronous motor (PMSM) play a major role in the future, development of wheel hub drive, because of its high power density, smooth torque and wide speed range. This study has incorporated the Fuzzy logic into the speed control of PMSM to improve its dynamic speed performances. The Fuzzy-logic controller (FLC) is based on the encoder-less vector control. The Fuzzy-logic speed controller is employed in the outer loop. The effectiveness of the FLC-based PMSM drive is investigated and compared to those obtained from the conventional proportional-integral (PI) controller-based

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drive at different dynamic operating conditions such as sudden change in command speed, step change in load, etc. This study attempts to provide such an in-depth comparison of the operation an encoder-less vector controlled PMSM with FLC and PI controller. The comparative results show that the Fuzzy controller offers faster settling response under different load disturbances and overcomes the nonlinearity problem of PMSM, as does not need an exact system mathematical model. So, FLC found to be a suitable replacement of the conventional PI controller for the high performance industrial drive applications with robust performance.

**Keywords:** PMSM, fuzzy speed controller, sensor-less, proportional-integral controller, hub motor

### 1. Introduction

Concerns for the global warming and increasing oil, prices have resulted to increase the interest in the area of electrical vehicles (EVs). By employing high-torque at low-speed electric motors, which could be able to directly mounted inside the wheel, known as in-wheel motors, or hub motors the transmission losses are minimized and operating efficiency is improved [1]. In-wheel motors provide noteworthy advantages for EVs such as, eliminating mechanical transmission including gearboxes, differentials, drive shafts and axles provide a significant weight and manufacturing cost saving, while also decreasing the environmental impact of the product. Furthermore, all-wheel drive (AWD) is easily provided by applying in-wheel motors for all wheels without any mechanical transmission and further efforts for reconfiguration the mechanical system. Addition to potential advantage of AWD/4WD, all wheels equipped with in-wheel motors can provide 90-degree steering lock, which help to park and get out of the smallest spaces even equal to length of the vehicle. Although the in-wheel motor provide significant advantages, still the major disadvantage of wheel hub motors are that the weight of the electric motors would increase the unsprung weight, which adversely affects handling and ride. So the in wheel electrical motor should have special criteria such as large starting torque, overload capability, wide speed range, and high power density, in order to reduce motor weight (i.e., unsprung weight) to mount inside the wheel [2].

Three different type of electrical motor, namely induction motor (IM), permanent magnet synchronous motor (PMSM) and switched reluctance motor (SRM) are potential devices in the future development of in-wheel applications. The comparison between there potential candidate electrical motor for in-wheel applications is provided in Table 1. Amongst different type of electrical motor,
PMSM play a major role in the future development of in-wheel applications, because of its high power density and efficiency, smooth torque and wide speed range. The unsprung assembly of in-wheel permanent magnet synchronous motor (PMSM) manufactured by Protean Company for Mercedes-Benz Brabus E-class [3] (4WD full electric/plug-in hybrid) is shown in Figure 1.

There are two main types of Brushless motors. One is known as the Brushless DC Motor (BLDCM), characterized by constant flux density in the air gap around the pole faces [4]. The motor windings should be supplied with currents in the form of rectangular pulses [5-7]. The other motor ideally has sinusoidal flux and sinusoidal distribution of its windings. It is supplied with a sinusoidal current and is known as the Permanent Magnet Synchronous Motor (PMSM) or Brushless AC motor (BLACM). The commutation process has to ensure that the action of switching the current direction is synchronized with the movement of the flux in the air gap, and so the motor must have a sensor for measuring the position of the flux wave relative to that of the stator windings. Simple Hall-effect sensors are used with BLDCM in order to manage the commutation sequence and form the appropriate current waveform. On the other hand, a high-resolution encoder or resolver is necessary for the PMSM control mode to generate sinusoidal currents [1].

BLDC and PMSM differ in the number of states they use to control position. A trapezoidal BLDC motor, for example, has six states compared to the “continuous” states of a sinusoidal PMSM. The more states supported, the more precisely position can be controlled. However, more states require more complex control mechanisms and processing as well. BLDC provides efficient, reliable operation, medium-high torque and can be used in combustible environments for applications such as automation, traction, precision. A key advantage of these types of motors is that they have no brushes, eliminating a primary source of wear, maintenance, and EMI. In addition, they use magnets, giving them higher power density and greater efficiency. Technological advancements continue to bring BLDC technology to a wider range of low-cost applications traditionally served by standard DC brush motors, making it one of the fastest growing types of motors. While BLDC has been relatively more expensive than standard DC brush motors, BLDC is attractive in those applications where lower wear and maintenance play a major role in overall total cost of ownership [8].
Table 1. Comparison between different types of electrical motor for in-wheel applications

<table>
<thead>
<tr>
<th>Specification</th>
<th>Different type of in-wheel electrical motor</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Induction motor (IM)</td>
<td>Permanent magnet synchronous motor (PMSM)</td>
</tr>
<tr>
<td>Rotor and stator</td>
<td>Squirrel-cage rotor or wound rotor, radial and axial flux type</td>
<td>Damper winding, radial flux (RFPM) and axial flux (AFPM) type</td>
</tr>
<tr>
<td>Torque</td>
<td>Speed varies with load torque, Low starting torque</td>
<td>High starting torque</td>
</tr>
<tr>
<td>Speed</td>
<td>Must run less than Synchronous speed, speed varies with load, reach to higher speed by Flux-Weakening Control</td>
<td>Synchronous speed, need induction cage for line start, reach to higher speed by Flux-Weakening Control</td>
</tr>
<tr>
<td>Efficiency</td>
<td>Degraded efficiency at low load operation, need soft starting method (high inrush current)</td>
<td>High efficiency than induction motor</td>
</tr>
<tr>
<td>Reliability</td>
<td>More heat degrade insulation and life time.</td>
<td>Lower temperature reduce maintenance, trouble-free operation for years</td>
</tr>
<tr>
<td>Power density</td>
<td>Induction field &amp; rotor iron limits power density</td>
<td>Higher than induction and synchronous motors, axial flux type has higher power density</td>
</tr>
<tr>
<td>Regenerative breaking</td>
<td>Run higher than synchronous speed</td>
<td>No need for additional efforts</td>
</tr>
<tr>
<td>Cost</td>
<td>Simple design and high manufacture quantities lower price, but high operating cost</td>
<td>Higher cost, but higher efficiency gives low operating cost, AFPM due to compact design has less production cost, and SPM has lower cost as it has a simple construction and shorter end connections.</td>
</tr>
</tbody>
</table>
BLDC is also replacing induction motor (IM) in applications that require low or variable speed operation, such as blower motors in HVAC systems that can adjust their speed to run more efficiently. In addition to providing better performance at variable speeds than IM, BLDC motors are smaller and more efficient for an equivalent cost. Driving BLDC motors through commutation is simple, resulting in savings in motor design, electronic controls, and lighter overall weight. Because of the inherent flux in the magnets, less current is required to drive the motor, leading to better efficiency for BLDC motors. While BLDC motors have historically been used in many position-control applications, PMSM are used in those applications requiring precise position control, very high speed, and/or high torque. Common applications include traction control, precision automation (robotics), and hybrid/electrical vehicles. PMSM offers higher efficiency and torque, has minimal torque ripple at commutation, offers a higher maximum achievable speed, works with low-cost distributed windings, and has low noise. Because PMSM is based on continuous sinusoidal control, less noise is observed during operation, leading to the ability to react to load changes more quickly and precisely hit required torque, speed, and position [8], these advantages of PMSM make this kind of motor one of the most attractive and efficient motor for different application especially for hybrid electrical vehicles (HEV). The comparison between BLDC and PMSM are shown in Table 2 and Figure 2 [9].

Surface permanent magnet (SPM) and interior permanent magnet (IPM) are two different topologies for PMSM according to the technique for mounting the PM into the rotor. The SPM motor has a simple construction and shorter end connections, but it is penalized by eddy-current loss at high speed, has a very limited transient overload power, and has a high uncontrolled generator voltage. The SPM motor has barely no overload capability, independently of the available inverter current. Moreover the loss behavior of the two motors is rather
different in the various operating ranges with the SPM one better at low speed due to short end connections but penalized at high speed by the need of a significant de-excitation current. The IPM motor shows the better performance in terms of power overload curve and efficiency at any load and any speed, compromise, but it might be more complicated to be manufactured [10]. Figure 3 show SPM and IPM configuration topologies for PMSM [11].

Table 2. Comparison between BLDC motor and PMSM for in-wheel applications

<table>
<thead>
<tr>
<th>BLDC motor</th>
<th>PMSM</th>
</tr>
</thead>
<tbody>
<tr>
<td>Easier to control (six trapezoidal states)</td>
<td>More complex control (continuous 3Ph sine wave)</td>
</tr>
<tr>
<td>Torque ripple at commutations</td>
<td>No torque ripple at commutation</td>
</tr>
<tr>
<td>Better for lower speed</td>
<td>Higher maximum achievable speed</td>
</tr>
<tr>
<td>Noisy</td>
<td>Low noise</td>
</tr>
<tr>
<td>Doesn’t work with distributed winding</td>
<td>Work with low-cost distributed winding</td>
</tr>
<tr>
<td>Not as efficient, lower torque</td>
<td>Higher efficiency, higher torque</td>
</tr>
<tr>
<td>Lower cost</td>
<td>Higher cost</td>
</tr>
</tbody>
</table>

Figure 2. Measured terminal line-to-line voltage waveform (1000rpm). (a) BLDC machine and BLAC machine, (b) FFT result to back-EMF in BLDC machine and BLAC machine.
In addition, there are two different permanent magnet machine topologies. Axial flux interior permanent magnet (AFMs) and Radial flux permanent magnet (RFMs). While the PM motors have small magnetic thickness, which results in small magnetic dimensions. As for the axial flux PM machines, they have a number of distinct advantages over radial flux machines (RFM). They can be designed to have a higher power-to-weight ratio resulting in less core material. Moreover, they have planar and easily adjustable air gaps. The noise and vibration levels are less than the conventional machines. In addition, the direction of the main air gap flux can be varied and many discrete topologies can be derived. These benefits present the AFMs with certain advantages over conventional RFMs; therefore, there is more interest to apply axial flux permanent magnet in various applications such as gearless wind turbine, in-wheel motor for electric vehicle [12]. The axial and radial flux permanent magnet topology is shown in Figure 4 [13].
1.1. Feature of Axial Flux Permanent Magnet Synchronous Motor (AFPMSM)

The axial flux permanent magnet (AFPM) is a disc-type that has a pancake shape and flux distribution as shown in figure above. It is different from conventional radial flux machine in terms of the flux direction that runs parallel with the mechanical shaft of the machine. Axial flux permanent magnet synchronous motor (AFPMSM) is an alternative to the Radial flux permanent magnet synchronous machine (RFPMSM). Because of its pancake shape, compactness, and high density. Due to this reason, AFPM are high power density, compact machine construction and short frame, high efficiency (no rotor copper losses due to the permanent magnet excitation), and having a short rotor in axial direction with the ability of the construction without rotor steel. AFPM machine have usually been used in integrated high–torque applications because it can provide high torque. Nevertheless, the winding losses of this type of machines are comparably high especially at high-speed application. This type of motor is particularly suitable and widely used for electrical vehicles, fans, valves, and pumps. Another example of high dynamic application for AFPMSM motor drive can be found in aerospace and robotic fields, as well as EVs and Hybrid EVs (HEVs), nowadays for such applications the interest for this kind of electrical motor drive is growing very fast. The list below shows some prominent example of AFPM application with the corresponding machine power range, types, and operating speed [14];


II. Wheel direct drive for electrical vehicles: Double –side configuration with internal stator, slotless toroidal winding arrangement, with surface –mounted permanent magnet, rated speed less than 1000 rpm or internal permanent magnet rotors,

III. High –speed generation driven by a gas turbine in a hybrid traction system: Three types of multi-disc (2, 4, 6,-rotor verses 1, 3, 5–stator respectively) AFPM machine were investigated for 10, 30, 50 KW at 60000 rpm.

IV. Engine driven and starter motor: Slotless toroidal stator, two rotor disks, 2.5 KW at 3000 rpm.
V. Adjustable speed pump application: A slotless stator core with concentrated coils, surface-mounted permanent magnet (Ferrite for cost minimization) on rotor disks, 880 W at 2800 rpm.

There are several types of AFPM motor, which can be classified on the stator–rotor arrangement, the technique to integrate the permanent magnet to the rotor and the existence of armature slots.

I. Stator–rotor arrangement:
   Multi–disk structure

II. Single-side structure:
    Internal stator
    Internal rotor

III. Existence of armature slots and cores:
    Slotted stators
    Toroidally-wound slotless stators

Unlike the radial flux, AFPM machines can be designed as double side, single-side or even as multi-disk configuration. The simplest and cheapest construction is the single–side type. Nevertheless, this type of construction is not popular due to the low torque production and bearing problem caused by the high attractive force normal to the plane of the air-gap, which tends to bring the two parts together. This problem can be solved by the use of double–side structure. The high attractive force between the rotor and stator can be counterbalanced by using second stator or rotor mounted as the mirror image of the first. This makes double-side motor the most promising and widely used types [4].

Double–side motor can be constructed either with an internal permanent–magnet disk rotor sandwiched by two stator cores or with internal stator sandwiched by two permanent magnet disk. A double side with an internal permanent magnet disk rotor has poor utilization of the machine copper. The flux paths are in the stators and relatively large iron losses are more pronounced compared to the other type. The double–rotor configuration has lower copper and iron losses because the flux return paths are in the disks. Besides, it also has higher power density compared to the double-stator configuration. The cooling for this type of machine is much easier, since the rotors are rotating next to the sides of the windings, act as a fan. Nevertheless, this structure is more difficult to be manufactured and the windage losses are more pronounced. For the power requirement, multi-disk types can be the most attractive solution [15]. The Double–side structure axial flux motors are shown in Figure 5.
The objective of this study is to develop a FL speed controller to control the speed of the AFPMSM. The field oriented control is applied for torque control while Fuzzy logic is applied to control the speed by setting the reference value for the quadrant current $I_q$. In addition, a sensorless method base on Fuzzy logic observer is developed to estimate the rotor position and motor speed base on the motor current and voltage to eliminate the usage of encoder. Because applying the encoder implies additional cost for electronics parts, extra wiring, extra space, and careful mounting which gets far from the inherent robustness of AFPMSM the sensor less method for controlling the speed of permanent magnet synchronous motors (AFPMSM) is applied in this project. By applying the sensor less method base on FL observer to estimating the motor position at the same time applying the FL controller to control the speed of AFPMSM therefore the advantages of FL controller for controlling the speed, estimation the motor position and estimation the motor speed are taken for both controlling and estimation part. Although the cost of the controlling part of the system reduced significantly, the system became more complex. However, the objective is to keep the performance of the system within acceptable range by applying this method. The speed responses of FL speed controller is compare with conventional PI controller and PI controller tuned by metaheuristic algorithm to prove the superiority of FL speed controller. The estimated motor speed and estimated motor position are compared with actual rotor speed and position to check the performance of the estimation part.

Figure 5. Double –side structure axial flux motor, (a) double-side structure (internal rotor), (b) double-side structure (internal stator).
2. **Speed Control of Permanent Magnet Synchronous Motor (PMSM)**

The main feature of synchronous motors (SMs) is the fact that the rotating speed of the rotor is equal to the frequency of the supply voltage divided by the number of pole pairs. Thus different speed control strategies for SMs have been proposed to track the reference frequency of supplied voltage (reference speed) more accurately. In comparison with the scalar control technique (V/f), vector control (VC) methods provide high accuracy for speed control of SMs. Two different speed control strategies for PMSMs are field oriented control (FOC) and direct torque control (DTC). These two strategies can be considered among the family of vector control (VC) methods and provide a solution for high-performance drives. The comparisons between two VC strategies for speed control of SMs are presented in Table 3 [16]. The comparison shows that, although FOC is more complicated, it does not suffer from high torque ripple thus it widely used for wheel direct drive.

VC is based on projection, which transform a three-phase time and speed dependent system into a two coordinate (d and q coordinates) time invariant system. In a field control algorithm, there are two important mathematical transformations, namely Clark transformation and Park transformation. The Clarke transformation is to convert a three phase current vector into two-phase stationary current vector. The Parks transformation is to convert that two-phase stationary current vector into rotating current vectors. The use of the Clark and Park transform bring the stator currents, which can be controlled, into the rotor domain. Doing this allows a motor control system to determine the voltages that should be supplied to the stator to maximize the torque under dynamically changing loads. The fundamentals of field oriented control implementation can be explained with the help of the Figure 6 [17], where the machine model is represented in a synchronously rotation reference frame. The inverter is omitted from the figure. Assuming that, it has unity current gain. It generates $i_a, i_b, i_c$ as dictated by the corresponding command current $i_a^*, i_b^*, i_c^*$ from the controller. A machine model with internal conversions is shown on the right. The machine terminal phase current $i_a, i_b, i_c$ are converted to $i_d, i_q$ components by abc-to-dq transformation. These are then converted to synchronously rotating frame by the unit vector components ($\cos \theta_c$) and ($\sin \theta_c$) before applying them to the $d-q_e$ machine model. The controller makes two stages of inverse
transformation, as shown, so that control currents $i^*_{ds}$ and $i^*_{qs}$ correspond to the machine currents $i_{ds}$ and $i_{qs}$ respectively [18].

<table>
<thead>
<tr>
<th>Comparison property</th>
<th>FOC</th>
<th>DTC</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dynamic response to torque</td>
<td>Fast</td>
<td>Very fast</td>
</tr>
<tr>
<td>Coordinates reference frame</td>
<td>d, q (rotor)</td>
<td>alpha, beta (stator)</td>
</tr>
<tr>
<td>Low speed (&lt; 5% of nominal)</td>
<td>High performance with position or speed sensor</td>
<td>Requires speed sensor for continuous braking</td>
</tr>
<tr>
<td>Controlled variables</td>
<td>rotor flux, torque current $i_q$ &amp; rotor flux current $i_d$ vector components</td>
<td>torque &amp; stator flux</td>
</tr>
<tr>
<td>Torque/current/flux ripple</td>
<td>Low</td>
<td>High, moderate (requires high quality current sensors)</td>
</tr>
<tr>
<td>Parameter sensitivity, sensorless</td>
<td>d, q inductances, rotor resistance</td>
<td>Stator resistance</td>
</tr>
<tr>
<td>Rotor position measurement</td>
<td>Required (either sensor or estimation)</td>
<td>Not required</td>
</tr>
<tr>
<td>Current control</td>
<td>Required</td>
<td>Not required</td>
</tr>
<tr>
<td>PWM modulator</td>
<td>PWM - Space vector modulation (SVM)</td>
<td>SVM</td>
</tr>
<tr>
<td>Coordinate transformations</td>
<td>Required</td>
<td>Not required</td>
</tr>
<tr>
<td>Switching frequency</td>
<td>Constant</td>
<td>Varies widely around average frequency</td>
</tr>
<tr>
<td>Switching losses</td>
<td>Low</td>
<td>Lower (requires high quality current sensors)</td>
</tr>
<tr>
<td>Audible noise</td>
<td>constant frequency whistling noise</td>
<td>spread spectrum sizzling noise</td>
</tr>
<tr>
<td>Control tuning loops</td>
<td>Speed controller, rotor flux controller (flux-weakening method), $i_d$ and $i_q$ current controls (PI)</td>
<td>speed (PID control)</td>
</tr>
<tr>
<td>Complexity</td>
<td>Higher</td>
<td>Lower</td>
</tr>
<tr>
<td>Typical control cycle time</td>
<td>100-500 microseconds</td>
<td>10-30 microseconds</td>
</tr>
</tbody>
</table>

In addition, the unit vector assures correct the alignment of $i_{ds}$ current with the flux vector $\phi_r$ and $i_{qs}$ perpendicular to it, as shown. The transformation and
inverter transformation including the inverter ideally do not incorporate any dynamics, and therefore, the response to ids and \( i_{qs} \) is instantaneous (neglecting computational and sampling delays).

The transformation of voltage from three-phase into two-phase are prerequisite from field oriented control. Assume that the d-q axes are oriented at \( \theta \)-angle, the transformations of voltages from the three-phase stationary reference frame (a-b-c) into two-phase stationary reference frame (d-q) are:

\[
\begin{bmatrix}
  v_q \\
  v_d \\
  v_0
\end{bmatrix}
= \frac{2}{3}
\begin{bmatrix}
  \cos(\theta) & \cos(\theta - 120) & \cos(\theta + 120) \\
  \sin(\theta) & \sin(\theta - 120) & \sin(\theta + 120) \\
  0.5 & 0.5 & 0.5
\end{bmatrix}
\begin{bmatrix}
  v_d \\
  v_b \\
  v_c
\end{bmatrix}
\tag{1}
\]
While the zero sequence components are ignored, the transformation relation can be written as:

\[ v_q = \frac{2}{3} \left[ v_a \cos(\theta) + v_b \cos(\theta - 120) + v_c \cos(\theta + 120) \right] \]
\[ v_d = \frac{2}{3} \left[ v_a \sin(\theta) + v_b \sin(\theta - 120) + v_c \sin(\theta + 120) \right] \]  

(2)

Trigonometric identities are used to simplify the equation. Beside, by defining \( V_{ab} = V_a - V_b \) and \( V_{bc} = V_b - V_c \), the voltage equation can be further simplified as:

\[ v_q = \frac{1}{3} \left[ (2v_{ab} + v_{bc}) \cos(\theta) + (\sqrt{3} v_{bc}) \sin(\theta) \right] \]
\[ v_d = \frac{1}{3} \left[ (2v_{ab} + v_{bc}) \sin(\theta) - (\sqrt{3} v_{bc}) \cos(\theta) \right] \]  

(3)

The inverse transformation from d-q into a-b-c is:

\[
\begin{bmatrix}
V_a \\
V_b \\
V_c \\
\end{bmatrix} = \begin{bmatrix}
\cos(\theta) & \sin(\theta) & 1 \\
\cos(\theta - 120) & \sin(\theta - 120) & 1 \\
\cos(\theta + 120) & \sin(\theta + 120) & 1 \\
\end{bmatrix} \begin{bmatrix}
v_q \\
v_d \\
v_0 \\
\end{bmatrix}
\]  

(4)

The equation can be simplified as:

\[ V_a = V_q \cos(\theta) + V_d \sin(\theta) \]
\[ V_b = \frac{1}{2} \left[ \cos(\theta) \left( -v_q - \sqrt{3} v_d \right) + \sin(\theta) \left( \sqrt{3} v_q - v_d \right) \right] \]
\[ V_c = \frac{1}{2} \left[ \cos(\theta) \left( -v_q + \sqrt{3} v_d \right) + \sin(\theta) \left( \sqrt{3} v_q - v_d \right) \right] \]  

(5)

2.1. Field-Oriented Control (FOC)
Field-oriented control (FOC), is a variable-frequency drive (VFD) speed control method in which the stator currents of a PMSM are identified as two orthogonal components that can be visualized with a vector. FOC provides a very accurate steady state and transient speed control of the motors, which leads to robust and high performance in terms of response times and power conversion. The invention of FOC in the early of 1970 is had brought a renaissance in the high-performance control of ac drives. Basically, a FOC motor drive operates like a separately exited DC motor drive. In a DC machine, the torque developed in the motor is a result of the interaction between current in the armature winding and the magnetic field produced in the field system of the motor. The field should be maintained at a certain level so that it is sufficiently high to yield a high torque per unit ampere, at the same time not too high in order to prevent the excessive saturation of the magnetic circuit of the motor. With fixed field, the torque is proportional to the armature control. DC machine-like performance can also be extended to a PMSM if the machine control is considered in a synchronously rotating reference frame, where the sinusoidal variable appear as DC quantities in steady-state. FOC of in-wheel PMSM for EVs using SVPWM technique is shown in Figure 7.

2.2. Direct Torque Control (DTC)

In mid 80’s a new VC method appeared to become an alternative to FOC known as direct torque control (DTC). DTC is one method used in variable frequency drives to control the torque (and thus finally the speed) of three-phase AC electric motors. This involves calculating an estimate of the motor’s magnetic flux and torque based on the measured voltage and current of the motor [19]. Stator flux linkage is estimated by integrating the stator voltages. Torque is estimated as a cross product of estimated stator flux linkage vector and measured motor current vector. The estimated flux magnitude and torque are then compared with their reference values. If either the estimated flux or torque deviates too far from the reference tolerance, the transistors of the variable frequency drive are turned off and on in such a way that the flux and torque errors will return in their tolerant bands as fast as possible. Thus direct torque control is one form of the hysteresis or bang-bang control. In principle the DTC method selects one of the six nonzero and two zero voltage vectors of the inverter on the basis of the instantaneous errors in torque and stator flux magnitude. In DTC, a single stator voltage vector of the voltage source inverter (VSI) standard topology is selected during every control sampling period, and it is maintained constant for the whole period. By this switching technique,
based on hysteresis, large and small torque are not differentiated, which causes an extra torque ripple in motor steady state operation [20]. Figure 8 illustrates DTC of in-wheel PMSM for EVs using SVPWM technique.

This study covered a field oriented control algorithm, electrical model of PMSM, proportional-integral-derivative (PID) controller, Fuzzy logic (FL) controller, and sensor less method base on FL observer. The overall model is shown in Figure 9. An AFPMSM is modeled base on its electrical equation circuit. Two controlling loop is applied, the inner loop consist of two PI controllers designed base on the motor parameter which applied to control the I_d and I_q. The reference I_d is set to zero for nominated speed because, the maximum torque is obtained with I_d = zero which corresponds to the case when the rotor and stator fluxes are perpendicular. The flux weakening method control the I_d current while the motor runs above the rated speed. The estimator base on Fuzzy logic observer is applied to estimate the rotor angle; the speed will compare with reference speed to make a feedback and send the error signal to Fuzzy speed controller. The FL speed controller is built in order to obtain the desired quadrature-axis current (I_q) which is the reference for the inner loop.
Figure 7. Field-oriented control (FOC) of in-wheel PMSM using SVPWM technique.
Figure 8. Direct torque control (DTC) of in-wheel PMSM using SVPWM technique.
Figure 9. Speed control of AFPMSM by Fuzzy logic controller and sensor less method based on Fuzzy logic theory for in-wheel drive applications.
The PID controller calculates an error value as the difference between a measured process variable and a desired set point. The controller attempts to minimize the error by adjusting the process control input. The PID parameters used in the calculation must be tuned according to the nature and mathematical model of the system. The conventional PID speed controller can be successfully applied in controlling linear systems. However, it is not able to cope with nonlinear systems with the same success. Machine drives might behave as nonlinear systems, where non-linearity may appear due to speed coupling, armature current limitation, change of inertia, nonlinear load, poor measuring of the parameters of motor, friction, change of the environment, etc [21, 22]. Consequently, it might be useful to use controllers with nonlinear transfer function. The system step response for a given reference speed is one of the performance indicators of the speed controller. It is desired that the step response of the system have minimal rise time, without overshoot and negligible steady state error. However, conventional proportional–derivative (PD) or proportional–integral (PI) controllers cannot be tuned in such a way that optimum step response is achieved for different inertia, load, friction, and change of the speed reference. PI controller is often used to regulate the speed, torque, and current in synchronous rotating coordinates, because of its easy implementation. The PI is sensitive to external disturbance and parameters variations, and the constant gains cannot meet different working points. Sometimes several groups of gains are needed to obtain excellent performance in both low speed and high-speed ranges. This is why a nonlinear controller is needed, like the Fuzzy controller, even in case of DC drives.

The PI controller generally cancels the steady-state error and cancels disturbance effects arising from load torque changes. Whereas the Fuzzy logic control theory performs in the case of sufficiently large reference input change (large error) at the starting time and slightly slower than conventional PI controller. The advantages of Fuzzy logic controllers (FLC) compared to conventional and classical controller are the ability to control the nonlinear system, ability to control the system without any precise or accurate mathematical model of the system have some and simplify controlling system. The concept of Fuzzy logic (FL) theory is to implement the experience and sense of the human, researcher, and operator for controlling the system [23, 24]. In order to get control schemes that would be less sensitive to parameters of the
motor variations that conventional PI controllers, a variable input membership Fuzzy logic controller was suggested.

Previous work was done to test and prove superiority of the dynamic input membership scheme (DIMS) for induction motor (IM) and a DC motor controller. The Fuzzy logic (FL) controlling system center with narrow (CWN) was applied at starting time to cancel or reduce the speed overshoot and while the response was getting close to the set point to reach the best possible value for speed and current. For the next step the center with constant (CWC) was applied to reach the better performance of speed control. Finally, to cancel the speed oscillation the center with wide (CWW) was applied while the system response is reached to steady-state. As previous work, show the dynamic input membership scheme Fuzzy logic controller (DIMS FLC) has been verified the superiority of this controller for controlling the speed of DC motor. Nevertheless, for controlling the speed of AFPMSM via DIMS FLC has some problem such as slow responses for speed step and weak operation during the starting time. Therefore, in this study DIMS FLC is applied to utilized the advantage Fuzzy logic theory for accurate and fast response speed control of in-wheel AFPMSM. For high performance motor drives, some kind of speed estimation is required in order to perform better speed control. Speed estimation from terminal quantities can be obtained either by exploiting magnetic saliencies in the machine or by calculations according to feedback parameters and mathematical model of machine. Speed estimation using magnetic saliencies, such as rotor asymmetries, rotor slotting, or variations on the leakage reactance, depends on machine parameters and can be considered a true speed measurement. In some of these methods, machine modification or injection of disturbance signals is needed. Generally, these techniques cannot be used directly as speed feedback signal for high performance speed control, because they present relative large measurement delays, or because they can only be used within a reduced range of frequencies. Speed information can alternatively be obtained by using a machine model fed by stator quantities. These methods use simple open loop speed calculators, Model Reference Adaptive Systems (MRAS) and Extended Kalman filters. In this study, the speed information is calculated based on the motor voltage and current by applying the Fuzzy logic observer. This method has some similarity with sliding mode observer but for the sliding mode estimation method the back-EMFs are obtained from the machine model. Nevertheless, for position estimator based on Fuzzy logic observer the back-EMFs are obtained from FL observers.
Figure 10. Typical Mamdani implication method for speed control with two activated rules.
3.1. Fuzzy Logic Control (FLC)

In this study, Fuzzy logic theory is used to control the speed of an AFPMSM motor drive and also to estimate the rotor angle and speed. This theory uses incorporating engineering knowledge into the automatic control system by using the intuition and experience of the designer. In 1965, this strategy was first proposed by Zadeh complicated systems control that are too hard to be analyzed by traditional mathematics. However, Fuzzy logic theory did not find wide popularity in various applications such as economics, management, medicine, or process control until the 1970’s. Mamdani introduces the first application of Fuzzy set theory for controlling a small laboratory steam engine. After success of this theory, many scientists were inspired to attempt to control industrial processes such as automatic trains, chemical reactors, or nuclear reactors using Fuzzy algorithms. Experiments showed that Fuzzy controllers perform better at least the same as adaptive controllers. One of the advantages of this technique is requiring only a simple mathematical model to formulate the algorithm. It eases the digital implementation. For nonlinear processes which do not have any reliable model or their model is too complex due to the large number of equations involved Fuzzy model is appreciated. Although a Fuzzy logic control FLC method is used frequently for DC, induction motors and permanent magnet motors, not existence of any systematic procedure for the design of a Fuzzy controller is still one of the main problems.

FLC is being used in many engineering applications in spite of its relatively new concept due to its simplest available solutions for the specific problems, which is an advantage over more traditional solutions. It also allows computers to have more human-like reasoning. FLC is considered as low-cost method to be implemented on chip. The needed chip does not need high-resolution analog-to-digital converters. These systems can be improved very easily and only by inserting some new rules for better performance or new features. Furthermore, for a complicated mathematical model, which needs a too expensive powerful chip processor that is not always practical; in such cases, Fuzzy logic controller is so useful. Implementation of PID controllers for cases with complex model is a challenge. Especially if automatic PID tuning is requested, the Fuzzy logic provides less complexity with higher control quality. Approximation of the second-order switching curve used in time-optimal control systems by a polynomial of the first or higher order makes Fuzzy control a better candidate for time-optimal control applications. As a relatively new control method, it provides more space for further improvements. Zadeh introduced FLC as
knowledge and experiment of human that are prepared for uncertain control conditions. A lot of research and experiments are done on this new system control branch and now FLC has become so popular. Application of Fuzzy control so far has been so useful and it helps engineers in a vast area of industry, which has attracted a lot of attention. A typical Mamdani implication method for speed control with two activated rules is shown in Figure 10 [25].

3.2. Proportional-Integral-Derivative (PID) Control

A proportional–integral-derivative (PID) controller is a generic control loop feedback mechanism widely used in industrial control systems. The function of PID controller is to correct the error between a measured process variable and a desired set point by calculating and then set the output a corrective action that can adjust the process accordingly. The PID controller algorithm involves three separate, namely the Proportional, the Integral and Derivative values. The form of the PID controller expresses as:

$$G_P(S) = K_D S + K_P + K_I / S$$

A proportional gain ($K_P$) will have the effect of reducing the rise time and will reduce, but never eliminate, the steady-state error. An integral gain ($K_I$) will have the effect of elimination the steady state error, but it may make the transient response worse. A derivative gain ($K_D$) will have the effect of increasing the stability of the system, reducing the overshoot, and improving the transient response. Effects of each of controller $K_D$, $K_I$, and $K_P$ on a closed loop system are summarized in Table 4 [26].

<table>
<thead>
<tr>
<th>Gain</th>
<th>Rise time</th>
<th>Overshoot</th>
<th>Settling time</th>
<th>Steady-state error</th>
</tr>
</thead>
<tbody>
<tr>
<td>$K_D$</td>
<td>Decrease</td>
<td>Increase</td>
<td>Small change</td>
<td>Decrease</td>
</tr>
<tr>
<td>$K_I$</td>
<td>Decrease</td>
<td>Increase</td>
<td>Increase</td>
<td>Eliminate</td>
</tr>
<tr>
<td>$K_P$</td>
<td>Small change</td>
<td>Decrease</td>
<td>Decrease</td>
<td>Small change</td>
</tr>
</tbody>
</table>

Table 4. Control responses with the increase of $K_D$, $K_P$ and $K_I$ of PID controller

Tuning the control loop is a process of selecting its control parameters (gain/proportional band, integral gain/reset, derivative gain) to meet given
performance specifications. If the PID controller parameters are chosen incorrectly, the controlled process input can be unstable; its output diverges, with or without oscillation, and is limited only by saturation or mechanical breakage. It is important to fine-tune the PID controller parameter in order to get the acceptable responses from control systems and in order to meet the steady state and transient response specifications of a system, to tune the PID controller it is requires either a dynamic model or a detailed frequency response over a wide range of frequencies. There are several methods for tuning a PID loop. There are comparisons of different tuning methods. Conventionally, Ziegler and Nichols suggested rules for tuning PID controller based on experimental step response or based on the value of $K_p$ that results in marginal stability when only proportional control action is used. Using metaheuristic optimization algorithms for tuning PID controller is an alternative technique. The optimization methods change the gains in other to minimize the output error of the system.

3.3. Sensor Less Methods

The use of vector controlled PMSM drives provides several advantages over DC machines in terms of robustness, size, lack of brushes, and reduced cost and maintenance. However, the speed control of PMSM requires the use of an accurate shaft encoder for correct operation. Applying this encoder implies additional electronics, extra wiring, extra space, and careful mounting which gets far from the inherent robustness of PMSM. Moreover, at low powers the cost of the sensor is about the same as the motor. Even in high power, it can still be between 20 to 30% of the machine cost. Therefore, there has been great interest in the research community in developing a high performance PMSM drive that does not require a speed or position transducer for its operation.

Some kind of speed estimation is required for high performance motor drives, in order to perform speed control. Speed estimation from terminal quantities can be obtained either by exploiting magnetic saliencies in the machine or by using a machine model. Speed estimation using magnetic saliencies, such as rotor slotting, rotor asymmetries or variations on the leakage reactance, is independent of machine parameters and can be considered a true speed measurement. Some of these methods require specially modified machines and the injection of disturbance signals. Generally, these techniques cannot be used directly as speed feedback signal for high performance speed
control, because they present relative large measurement delays, or because they can only be used within a reduced range of frequencies.

Usually, the key problems related to the low-speed operation of model-based sensorless drives are imputed to stator voltage and current acquisition, machine parameter sensitivity, flux pure-integration problems and inverter non-linearity. Most of these techniques fail at or around zero speed because all model-based estimation techniques utilize rotor-induced voltages, which are too small at zero stator frequency.

Moreover, the methods proposed so far ultimately fail at low and zero speed in wheel motor tests due to the absence of measurable signals. Indeed, the position error and the torque losses are relatively large in these conditions. From the mathematical model of the PM Brushless Motor, it can be observed that the back-EMFs (BEMFs) or flux linkage varies as a function of the rotor position. Therefore, if these quantities are measured or estimated, the rotor position information can be determined. However, it is difficult to measure the back-EMFs, specifically at low operating speeds, or the flux linkages directly because of the integration drift and/or shift [1].

At zero speed, the BEMF induced voltage is equal to zero. The BEMF becomes significant from a certain rotor angular speed. Therefore, sensorless estimation techniques perform with a perfect precision in the medium to high-speed range. Usually, it can be used from 5% of nominal speed or lower. The sensorless estimation techniques that use any kind of feedback observer with a system model are the most popular. These techniques are based on a deterministic system model. Usually motor identification is provided before setting the controller parameters and the controller uses a system model that estimates the motor BEMF or flux with good enough precision. The motor model is usually based on an invariant stator coil inductance (insignificant variance). Many PMSMs have an inductance variance negligible over rotor angle especially if compared to the BEMF influence. Then the observer calculated in α and β coordinates can be used. Some PMSMs have significant inductance saliency that prevents the use of the α and β model. In this case, most of the PMSMs have an inductance almost constant in the rotor related d, q coordinates (lateral induction $L_q$ is constant but different from $L_d$, which is also constant). Then the rotor related d, q coordinate model is used [27]. The position of the rotor with its permanent magnet flux is not certain. It can be estimated from the BEMF as follows [28]:

While the model of the PMSM in the stationary frame ($\alpha$ and $\beta$) is expressed
\[ V_\alpha = R i_\alpha + e_\alpha \]
\[ V_\beta = R i_\beta + e_\beta \]  
(7)

where
\( e_\alpha \) and \( e_\beta \) are the back-EMF, \( V_\alpha \) and \( V_\beta \) are terminal voltage, \( i_\alpha \) and \( i_\beta \) are the terminal current \( R \) is the winding resistance.

The flux linkage in \( q\beta \)-frame are formulated according to Eq. 8.

\[ e_\alpha = L_\alpha i_\alpha + \Psi_m \cos(\theta) \]
\[ e_\beta = L_\beta i_\beta + \Psi_m \sin(\theta) \]  
(8)

where
\( \Theta \) is the rotor angle, \( \Psi_m \) is the maximum flux linkage of the permanent magnet. The \( L_\alpha \) and the \( L_\beta \) are the stator inductance in \( q\beta \)-frame for non-salient PMSM (AFPMSM) with sinusoidal back-EMF

\( (L_\alpha = L_\beta = L) \)  
(9)

The BEMF cannot be directly measured. Estimation from an observer is needed. The feedback observer estimates the BEMF using a system model with voltage and current vector inputs.

The estimated BEMF \( (e^*) \) are system variables. These system variables are used for position sine and cosine evaluation. Therefore the estimation rotor angle can expressed as:

\[ \theta^* = \arctan\left( -\frac{e^* s_\alpha}{e^* s_\beta} \right) \]  
(10)

4. **Electrical Model of Permanent Magnet Synchronous Motor (PMSM)**

The mathematical model of a PMSM is similar to that of the wound rotor synchronous machine. Figure 11 shows the per phase equivalent circuit of an AFPMSM machine with a coreless stator. This equivalent circuit can be used to calculate the steady-state performance of the machine. \( R_i \) is the stator resistance, \( X_i \) is the stator leakage reactance, \( E_i \) is the EMF induced in the stator winding by the rotor PM excitation system, \( E_i \) is the rms value of the internal phase
Voltage, $V$ is the terminal voltage, $I_{al}$, $I_{aq}$ and $I_d$ are the rms stator current. The shunt resistance $R_e$ is the stator eddy current loss resistance, which is defined in the same way as the core loss resistance for slotted PM brushless motors.

If we ignore the losses in PMs and losses in rotor backing steel discs, then the power can be calculated as follows:

Generator model (mechanical shaft power)

$$P_{in} = P_{elm} + \Delta P_{rot}$$

(11)

Motor mode (electrical power)

$$P_{in} = P_{elm} + \Delta P_{lw} + \Delta P_{e}$$

(12)

Where $P_{elm}$ is the electromagnetic power, $\Delta P_{lw}$ is the stator winding loss, $\Delta P_{e}$ are the eddy current losses in the stator conductors, and $\Delta P_{rot}$ are the rotational losses.

Correspondingly, the output power is:

Generator mode (mechanical shaft power)

$$P_{in} = P_{elm} + \Delta P_{lw} + \Delta P_{e}$$

(13)

Motor mode (electrical power)

$$P_{in} = P_{elm} + \Delta P_{rot}$$

(14)

Equivalent circuit of the motors are used for study and simulation of motors. The understanding of the PMSM fundamental equation may help in extracting the parameters from the machine specification. It is important to know fundamental equation before we can obtain the most important parameter, namely inductance for PMSM.

![Figure 11. Per-phase equivalent of PMSM machine. (a) Generator (b) Motor.](image)
4.1. Armature Reaction

The magnetic fluxes produced by the stator (armature), it can be expressed as:

\[ \Phi_{ad} = \frac{2}{\pi} B_{mad} \frac{\pi}{p} \left( \frac{R_{out}^2 - R_{in}^2}{2} \right) \]  
\[ \Phi_{aq} = \frac{2}{\pi} B_{maq} \frac{\pi}{p} \left( \frac{R_{out}^2 - R_{in}^2}{2} \right) \]  

Where \( B_{mad} \) and \( B_{maq} \) are the peak value of the first harmonic of the stator (armature reaction) magnetic flux density. The machine pole pair is expressed as \( P \), \( R_{in} \) is the inner radius of PMs equal to the inner radius of stator bars and \( R_{out} \) is the outer radius PMs equal to the outer radius of stator bars. The stator linkage fluxes are

\[ \Psi_d = \frac{1}{\sqrt{2}} N_1 k_{\omega 1} \Phi_{ad} = \frac{1}{\sqrt{2}} N_1 K_{mad} \frac{\pi}{p} \left( \frac{R_{out}^2 - R_{in}^2}{2} \right) \]  
\[ \Psi_q = \frac{1}{\sqrt{2}} N_1 k_{\omega 1} \Phi_{aq} = \frac{1}{\sqrt{2}} N_1 K_{maq} \frac{\pi}{p} \left( \frac{R_{out}^2 - R_{in}^2}{2} \right) \]  

Where \( N_1 \) is the number of stator turns per phase and \( K_{\omega 1} \) is the winding factor for the fundamental space harmonic. By neglecting saturation, the first harmonic of the stator magnetic flux density normal components are:

\[ B_{mad} = k_{fd} B_{mad} = K_{fd} \lambda_d F_{ad} = K_{fd} \frac{\mu_0}{g} \frac{m_1 \sqrt{2}}{\pi} N_1 k_{\omega 1} I_{ad} \]  
\[ B_{maq} = k_{fq} B_{maq} = K_{fq} \lambda_q F_{aq} = K_{fq} \frac{\mu_0}{g} \frac{m_1 \sqrt{2}}{\pi} N_1 k_{\omega 1} I_{aq} \]  

Where \( m_1 \) is the number of stator phase, \( \lambda_q \) and \( \lambda_d \) are the permanent per unit surface, \( K_{fd} \) and \( K_{fq} \) are the form factors of armature reaction. \( F_{ad} \) and \( F_{aq} \) are the armature reaction MMF. \( I_{ad} \) and \( I_{aq} \) are the d- and q- axis stator current. The equivalent air gaps for surface configuration of PMs with coreless stator are:

\[ g' = 2 \left\{ (g + 0.5t_w) + \frac{h_M}{\mu_{rec}} \right\} \]
\[ g_q^* = 2 \left( [g + 0.5t_w] + h_M \right) \] (22)

Where \( t_w \) is the stator winding axial thickness, \( h_M \) is the axial height of the PM and \( \mu_{\text{rec}} \) is the relative recoil permeability of the PM.

Armature reaction EMFs in the d-q axis are:

\[ E_{ad} = \pi \sqrt{2} f N_1 K_{ad} \Phi_{ad} \] (23)

\[ E_{aq} = \pi \sqrt{2} f N_1 K_{aq} \Phi_{aq} \] (24)

Armature reaction reactances are:

\[ X_{ad} = 2 \pi f L_{ad} = \frac{E_{ad}}{I_{ad}} \] (25)

\[ X_{aq} = 2 \pi f L_{aq} = \frac{E_{aq}}{I_{aq}} \] (26)

4.2. Calculation of Winding Inductance

Synchronous inductance, \( L_s \) consist of the armature reaction (mutual) inductance, \( L_{ad} \) and \( L_{aq} \) and the leakage inductance \( L_l \). Machine with a different in reluctances in the d-q axes (magnetic asymmetry), the synchronous inductances in the d-q axis, \( L_{sd} \) and \( L_{sq} \) are written as below:

\[ L_{sd} = L_{ad} + L_l \] (27)

\[ L_{sq} = L_{aq} + L_l \] (28)

Where \( L_{sq} \) and \( L_{aq} \) are the armature reaction (mutual) inductances:

\[ L_{ad} = \frac{\psi_d}{I_{ad}} = m_1 \mu_0 \frac{1}{\pi} \left( \frac{N_1 K_{w1}}{p} \right)^2 \left( \frac{R_{\text{out}}^2 - R_{\text{in}}^2}{g_{d}} \right) K_{fd} \] (29)

\[ L_{aq} = \frac{\psi_q}{I_{ad}} = m_1 \mu_0 \frac{1}{\pi} \left( \frac{N_1 K_{w1}}{p} \right)^2 \left( \frac{R_{\text{out}}^2 - R_{\text{in}}^2}{g_{q}} \right) K_{fq} \] (30)
For $\mu_{rec} \approx 1$ and surface configuration of PMs ($K_{fd} = K_{fq} = 1$), the d-and q-axis armature reaction inductances are equal:

$$L_d = L_aq = L_{aq} = m_1 \mu_0 \frac{1}{\pi} \left( \frac{N_1 K_{w1}}{p} \right)^2 \left( \frac{R_{out}^2 - R_{in}^2}{g'} \right)$$

(31)

The leakage inductance can be expressed as:

$$L_l = 2 \mu_0 \frac{N_1^2 L_{i1}}{p q_1} \left( \lambda_{1S} + \lambda_{1e} + \lambda_{1d} \right)$$

(32)

Where $L_l = 0.5 (D_{out} - D_{in})$ is the active length of a coil equal to the radial length $I_M$ of the PMs, $D_{out}$ and $D_{in}$ are outer and inner diameter of PM. The number of coil sides per pole per phase is expressed as:

$$q_1 = \frac{s_1}{2 p m_1}$$

(33)

Where $s_1$ is the number of stator slots and $m_1$ is the number of phases. The average length of the single-side end connection, $L_{ie}$ is expressed as:

$$L_{ie} = 0.5 (L_{in} + L_{out})$$

(34)

4.3. Dynamic d-q Modeling

The d-q model has developed on rotor reference frame, which is shown in Figure 12. At any time ($t$), the rotating rotor d-axis makes and angle $\theta_r$ with the fixed stator phase axis and rotating stator mmf makes an angle $\alpha$ with the rotor d-axis. Stator mmf and rotor are rotating at the same speed.
PMSM dynamic performance can be achieved by d-q current control system. Figure 13 shows the d and q-axis equivalent circuits of the sinusoidal permanent magnet brushless machine.

The voltage equation for the stator circuit are:

\[
V_{ld} = R_l i_{ld} + \frac{d}{dt} \psi_{ad} - w \psi_q \tag{35}
\]

\[
V_{lq} = R_l i_{lq} + \frac{d}{dt} \psi_{aq} + w \psi_d \tag{36}
\]
In which the linkage fluxes are defined as:

\[ \psi_d = (L_{ad} + L_d) i_{ad} + \psi_f = L_{ad} i_{ad} + \psi_f \]  \hspace{1cm} (37)

\[ \psi_q = (L_{aq} + L_q) i_{aq} + \psi_f = L_{sq} i_{aq} \]  \hspace{1cm} (38)

Where \( \psi_f = L_{fd} I_f \) is the maximum flux linkage per phase produced by the PM, \( L_{fd} \) is the maximum value of the mutual inductance between the armature and field winding. In the case of a PM excitation, the fictitious field current \( I_f \) is the maximum field current. \( H_c \) is a magnetic field intensity and \( h_M \) is the height of the PM. \( L_{ad} \) and \( L_{aq} \) are the d and q-axis components of the armature self-inductance, \( w \) is the angular frequency of the armature current. By substitute, the equation will be expressed:

\[ V_{ld} = R_l i_{ad} + \frac{d}{dt} L_{sd} - w L_{sq} i_{aq} \]  \hspace{1cm} (39)

\[ V_{lq} = R_l i_{ad} + \frac{d}{dt} L_{sq} + w L_{sd} i_{ad} + w \psi_f \]  \hspace{1cm} (40)

The instantaneous power to the motor input terminal is

\[ P_{in} = \frac{3}{2} (V_{ld} i_{ad} + V_{lq} i_{aq}) \]  \hspace{1cm} (41)

The electromagnetic power of a three-phase machine is

\[ P_{elm} = \frac{3}{2} w \left[ \psi_f + (L_{ad} - L_{aq}) i_{ad} \right] i_{aq} \]  \hspace{1cm} (42)

The electromagnetic torque of a three-phase motor with p-pole pairs is

\[ T_e = P \frac{P_{elm}}{w} \frac{3}{2} \left[ \psi_f + (L_{ad} - L_{aq}) i_{ad} \right] i_{aq} \]  \hspace{1cm} (43)

The electrical and mechanical torque balance equations are as follow:

\[ J \frac{dw}{dt} = p \left( T_e - T_m \right) - B w \]  \hspace{1cm} (45)

Where \( J \) is the inertia of the PMSM, \( T_m \) is the load torque; \( B \) is the approximated machine damping to friction.
The motor angular speed \( W \) is \( W = \frac{d\theta}{dt} \) where \( \theta \) is the rotor angle and 
\[ W_m = W_r \left( \frac{2}{p} \right) \]
While \( W_m \) is the mechanical rotor speed and \( W_r \) is the electrical rotor speed. The relationship between \( i_{ad}, i_{aq} \) and phase current \( i_{aA}, i_{aB} \) and \( i_{ac} \) are
\[
i_{ad} = \frac{2}{3} \left[ i_{aA} \cos(Wt) + i_{aB} \cos(Wt - \frac{2\pi}{3}) + i_{ac} \cos(Wt + \frac{2\pi}{3}) \right]
\]
\[
i_{aq} = \frac{2}{3} \left[ i_{aA} \sin(Wt) + i_{aB} \sin(Wt - \frac{2\pi}{3}) + i_{ac} \sin(Wt + \frac{2\pi}{3}) \right]
\]
The reverse relations in conjunction with \( i_{aA} + i_{aB} + i_{ac} = 0 \)
\[
I_{aA} = i_{ad} \cos(Wt) - i_{aq} \sin(Wt)
\]
\[
I_{aB} = i_{ad} \cos(Wt - \frac{2\pi}{3}) - i_{aq} \sin(Wt - \frac{2\pi}{3})
\]
\[
I_{aC} = i_{ad} \cos(Wt + \frac{2\pi}{3}) - i_{aq} \sin(Wt + \frac{2\pi}{3})
\]

### 4.4. Performance Characteristics

For most of the machine, the impact of the armature reaction must be taken into account in order to prevent sparking and flux distortion. Nevertheless, the armature reaction for PMSM with coreless stator is generally negligible. Thus, the influence of the armature reaction on the eddy current losses is usually insignificant. For Brushless motor with axial flux permanent magnet, with sinusoidal back-EMF the, Inductances in the d-q frame and \( \alpha-\beta \) frame are equal \( L_d = L_q \) and \( L_\alpha = L_\beta \).

Hence, the mathematical model of a PMSM is similar to that of the wound rotor synchronous machine. To perform the operation of PMSM or synchronous motor as DC (direct current) motor the field-oriented control is applied. The variable frequency and variable voltages are generated by an inverter to feed provide the variable voltage and frequency to connect to stator winding of the synchronous motor. The frequency and phase of the output wave are controlled using a position sensor instead of controlling the inverter frequency independently. The speed control block diagram of PMSMs is shown in Figure 14 [29].
4.5. Field-Oriented Control (FOC) of Permanent Magnet Synchronous Motor (PMSM)

Field oriented control was developed in the 1970s and it proved that an IM or synchronous motor (SM) could be controlled like a separately excited or self excited DC (direct current) motor by the orientation of the stator magnetomotive force (MMF) or current vector in relation to the rotor flux to achieve a desired objective. The control needs accurate knowledge of the position of the instant rotor position or the rotor flux of PMSM for behaving the motor like DC (direct current) motor. To measure the accurate position of the rotor a resolver or an absolute optical encoder is required. By applying the sensor less technique to estimate the position of the rotor instead of the encoder the controlling method are independent of any speed sensor. Some control options are constant torque and flux weakening.

The PMSM control is equivalent to that of the dc motor by a decoupling control known as FOC. The vector control separates the torque component of current and flux channels in the motor through its stator excitation. The vector control of the PM synchronous motor is derived from its dynamic model. Considering the currents as inputs, the three currents are:

\[
I_a = I_m \sin (W_r t + \alpha) \tag{51}
\]

\[
I_b = I_m \sin (W_r t + \alpha - \frac{2\pi}{3}) \tag{52}
\]

\[
I_c = I_m \sin (W_r t + \alpha + \frac{2\pi}{3}) \tag{53}
\]

Figure 14. Speed control block diagram of PMSM.
When $W_r$ is the electrical rotor speed and $\alpha$ is the angle between the rotor field and stator current phasors.

The previous currents obtained are the stator currents that must be transformed to the rotor reference frame with the rotor speed $W_r$, using Park’s transformation. The $q$ and $d$ axis currents are constants in the rotor reference frames since $\alpha$ is a constant for a given load torque. As these constants, they are similar to the armature and field currents in the separately excited dc machine. The $q$ axis current is distinctly equivalent to the armature current of the dc machine; the $d$ axis current is field current, but not in its entirety. It is only a partial field current; the other part is contributed by the equivalent current source representing the permanent magnet field. For this reason, the $q$ axis current is called the torque-producing component of the stator current and the $d$ axis current is called the flux-producing component of the stator current [29]. Which $I_d$ and $I_q$ in terms of $I_m$ is expressed as follows:

$$\begin{pmatrix} I_d \\ I_q \end{pmatrix} = I_m \begin{pmatrix} \sin (\alpha) \\ \cos (\alpha) \end{pmatrix}$$

(54)

In addition, the electromagnetic torque is obtained from the following equation.

$$T_e = P \frac{P_{elm}}{W} \frac{3}{2} [\psi_f + (L_{sd} - L_{sq}) i_{ad}] i_{aq}$$

(55)

4.6. Operation of Constant Torque

The constant torque control method is obtained from field-oriented control, while the maximum possible torque is obtained at all times like the DC (direct current) motor. In the other words to reach the maximum torque at all the time the $I_q$ should be equal to the supply current $I_m$. Which it results in selecting the $\alpha$ angle to be $(0, \frac{\pi}{2})$. By making, the $I_d$ current equal to zero the torque equation can be rewritten as:

$$T_e = \frac{3}{2} p \psi_f I_{aq}$$

(57)

By assuming $k_t$ as a constant value:
\[ K_e = \frac{3}{2} p \psi f \]  

(58)

\[ T_e = K_t I_{aq} \]  

(59)

Therefore, by applying the field orient control the electrical torque equation for PMSM resembles that of the regular DC motor. Therefore, may have very efficiently for controlling the machine torque. The motor currents are expressed into \( I_d \) and \( I_q \) components in the rotor based d-q coordinates system. The maximum torque is obtained while \( I_d = 0 \) which corresponds the case when the rotor and stator fluxes are at 90° degree angle. The operation of the drive is then similar to DC motor.

5. SIMULATION OF PERMANENT MAGNET SYNCHRONOUS MOTOR (PMSM) AND CONTROLLING PART

The simulation of the PMSM’s dynamic model is divided into two parts, mainly the electrical part and mechanical part. The electrical part executes the calculation of three phase current \( I_{abc} \) and calculation of tree phase voltage by using park transform. The mechanical part executes the calculation of rotor angle and motor angular speed. The details of the model is shown in Figure 15 and the parameters of simulated PMSM is listed in Table 5.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated Power</td>
<td>0.5 KW</td>
</tr>
<tr>
<td>Poles</td>
<td>4</td>
</tr>
<tr>
<td>Rated Voltage</td>
<td>220 V</td>
</tr>
<tr>
<td>Rated Speed</td>
<td>2000 rpm (210 rad·s⁻¹)</td>
</tr>
<tr>
<td>Rated Torque</td>
<td>1.7 Nm</td>
</tr>
<tr>
<td>Stator phase resistance</td>
<td>4.7Ω</td>
</tr>
<tr>
<td>Inertia</td>
<td>(3.57*10^{-3}) Kg.m²</td>
</tr>
<tr>
<td>Inductance (L_q = L_d)</td>
<td>0.0140 H</td>
</tr>
<tr>
<td>Friction</td>
<td>(0.47*10^{-3})</td>
</tr>
</tbody>
</table>
Two PI controllers are applied in the inner loop to control the \( u_d \) and \( u_q \) base on the difference between the reference current and the motor current in the d-q frame. The park’s transform is applied to convert the motor current to d-q frame. While the maximum torque is obtained with \( I_d = 0 \) which corresponds to the case when the rotor and stator fluxes are perpendicular therefore the reference value for \( u_d \) controller is set to zero and the reference value for the \( u_q \) controller is set by the speed controller (outer control loop). Figure 7 shows the simulation of the FOC control for PMSMs. Both PI controllers, which are applied in the inner loop, have same \( K_{ip} \) and \( K_{ii} \) gains. The \( K_{ip} \) and \( K_{ii} \) gains are calculated from the equation below

\[
K_{ii} = R/toi \quad \text{and} \quad K_{ip} = L/toi \ (toi=2.5e^{-3})
\]  

(60)

Where, \( L \) is the Inductance of the motor and \( R \) is the Stator phase resistance.

5.1. Proportional-Derivative (PI) Speed Controller

The proportional gain \( K_p \) and integrate gain \( K_i \) of PI controller (the outer control loop) is obtained from the mathematical model of the PMSM. Base on the mathematical model of the PMSM the \( K_p \) and \( K_i \) are obtained from the equation below:
Fuzzy Logic Based Encoder-Less Speed Control of PMSM …

\begin{align}
K_p &= \frac{J}{(gr*k_t*t_{ow})} \quad (61) \\
K_i &= \frac{f_r}{(gr*k_t*t_{ow})} \quad (62)
\end{align}

While in this study it is considered that:

\[ J = 3.57 \times 10^{-3} \text{ Kg.m}^2 \]
\[ f_r = 0.47 \times 10^{-3} \text{ Nm/rad.s}^{-1} \]
\[ t_{ow} = 0.1 \text{ s} \]
\[ gr = \frac{60}{2\pi} \]
\[ K_t = 0.45 \]

Where \( J \) is the Inertia of rotor and \( f_r \) is the friction.

The friction will change by operating the motor in different speed, different environment therefore the system will behave like a nonlinear system in addition accurate measuring the inertia of the rotor needs too many efforts and some time it is impossible to measure the accurate value. As reason that tune the PI controller to control the speed needs exact mathematical model of the motor, exact measurement of the parameters (sensitive to parameter variations) and some parameters make the system behaves like nonlinear system to overcome this problem the Fuzzy logic controller is proposed to control the speed PMSM. Fuzzy logic controller does not need exact system mathematical model in addition the Fuzzy logic controller will be able to handle intricate nonlinearity problem. Additionally, applying metaheuristic optimization algorithms for tuning PI controller is an alternative technique to eliminate the use of mathematical model for speed control of PMSMs. The optimization methods change the gains in order to minimize the output error of the system \([30, 31]\). So in this study the performance of FL speed controller is compared with PI controller tuned by genetic algorithm (GA) which is a most well-known metaheuristic optimization algorithm inspired by natural evolution \([32]\).

5.2. Fuzzy logic Speed Controller

The structure of the Fuzzy logic speed controller (outer control loop) is shown in Figure 16. The controller has two input variables. The first is speed error \( e(k) \) and the second is change of speed error \( ce(k) \). At the same time, change in reference phase current \( i_q^* \) is the output \( \Delta i_q^*(k) \) of the Fuzzy logic controller.
The definition of the FL rules is based on experimental and performance of the system to justify the FLC rules therefore, there are no special criteria to define the FL rules. However, the step responses of a PI controller provide an opportunity to define the rule justification. The step response of the system is shown in Figure 17. Moreover, the FLC rules can be extracted from rule based data mining algorithm.

Therefore, after thorough careful analysis of the system performance and the step response of the system the total 49 rules can be split into seven regions corresponding to each condition. The number of Fuzzy set, N applied here are seven. Therefore, for N number of Fuzzy sets, there will be $N^2$ number of rules in the Fuzzy control. The linguistic definitions of the membership function are negative large (NL), negative small (NS), positive large (PL), positive small (PS), negative medium (NM), approximately zero (ZE), positive medium (PM).

Table 6 shows the corresponding rule matrix for the Fuzzy control. There are altogether 49 rules in the matrix.

By analyzing the step response of the system, we will be able to write the rules. For example, when the speed error is zero $e(k) = 0$ and the change in the speed error is negative medium, $ce(k) = NM$ (C in Figure 17) the output is set to negative medium to minimize the speed overshoot and keep the speed at desired point, same logic is applied to obtain all of the 49 rules. Once speed error $e(k)$ and the second is change of speed error $ce(k)$ are determined, the output is computed using the Mamdani method where the min-max method is used for fuzzification purpose, whilst the center of area (CoA) method is used for defuzzification purpose.
Figure 17. Step response of the system.

Table 6. Fuzzy logic controller rules table

<table>
<thead>
<tr>
<th>$ce(k)$</th>
<th>NL</th>
<th>NM</th>
<th>NS</th>
<th>ZE</th>
<th>PS</th>
<th>PM</th>
<th>PL</th>
</tr>
</thead>
<tbody>
<tr>
<td>$e(k)$</td>
<td>NL</td>
<td>NL</td>
<td>NL</td>
<td>NL</td>
<td>NL</td>
<td>NM</td>
<td>NS</td>
</tr>
<tr>
<td></td>
<td>NM</td>
<td>NL</td>
<td>(D)</td>
<td>NL</td>
<td>NL</td>
<td>(E)</td>
<td>NM</td>
</tr>
<tr>
<td></td>
<td>NS</td>
<td>NVL</td>
<td>NL</td>
<td>NM</td>
<td>NS</td>
<td>ZE</td>
<td>PS</td>
</tr>
<tr>
<td></td>
<td>ZE</td>
<td>NL</td>
<td>(C)</td>
<td>NM</td>
<td>NS</td>
<td>(I)</td>
<td>ZE</td>
</tr>
<tr>
<td></td>
<td>PS</td>
<td>NM</td>
<td>NS</td>
<td>ZE</td>
<td>PS</td>
<td>PM</td>
<td>PL</td>
</tr>
<tr>
<td></td>
<td>PM</td>
<td>NS</td>
<td>(B)</td>
<td>PL</td>
<td>PS</td>
<td>(A)</td>
<td>PL</td>
</tr>
<tr>
<td></td>
<td>PL</td>
<td>ZE</td>
<td>PS</td>
<td>PM</td>
<td>PL</td>
<td>PL</td>
<td>PL</td>
</tr>
</tbody>
</table>
Figure 18. Surface view of Fuzzy logic rules.

Figure 19. Membership function of FLC first input, speed error $e(k)$.

Figure 20. Membership function of FLC second input, change of speed error $ce(k)$.
The membership function of the speed error $e(k)$, change of speed error $ce(k)$ and the reference phase current $i^*_q(k)$ is the output $\Delta i_q^*(k)$ of the Fuzzy logic controller are shown in Figures 19-21 respectively. Put into practice, one or two types of membership functions are enough to solve most of the problems. To simplify, triangular and trapezoidal shapes are used.

### 5.3. Fuzzy Logic Sensorless Method

The Fuzzy logic sensorless (encoder less) method is based on determination of rotor position and then speed by estimating back-EMF components, which result from Fuzzy logic observers. The rotor rotating speed and position angle of the rotor are estimated by evaluating the instantaneous values of PMSM’s current and voltage. Two Fuzzy logic observers are applied to estimate the rotor position. The Fuzzy logic observers have two inputs: the estimated stator current and the difference between the measured and estimated stator current [33]. The position estimator shows similarities with the sliding mode estimations. Instead of the machine model, back-EMF are obtained from the Fuzzy logic observers within proposed position estimator.

From mathematical model of the PMSM in $\alpha$ and $\beta$ frame (the Clarke Transform is applied to convert the PMSM current in to $\alpha$ and $\beta$ coordinates), it can be observed that the flux linkage or back-EMF varies as a function of the rotor position.

\[
U_\alpha = RI_\alpha + PLI_\alpha + KEc\sin(\theta) \quad (63)
\]
\[
U_\beta = RI_\beta + PLI_\beta + KEc\cos(\theta) \quad (64)
\]
\[
\begin{bmatrix}
U_\alpha \\
U_\beta
\end{bmatrix} = R \begin{bmatrix}
I_\alpha \\
I_\beta
\end{bmatrix} + PL \begin{bmatrix}
I_\alpha \\
I_\beta
\end{bmatrix} + KE W_e \begin{bmatrix}
\sin(\theta) \\
\cos(\theta)
\end{bmatrix}
\] (65)

\[
KE W_e \begin{bmatrix}
-\sin(\theta) \\
\cos(\theta)
\end{bmatrix} = e_a
\] (66)

\[
\frac{d\alpha}{dt} = \frac{v_\alpha}{L} - \frac{R I_\alpha - e_a}{L}
\] (67)

\[
\frac{d\beta}{dt} = \frac{v_\beta}{L} - \frac{R I_\beta - e_\beta}{L}
\] (68)

where, \( e_a \) represent the back-EMF, \( R \) is the stator resistance, \( p \) is the differential operator, \( KE \) is the back-EMF constant, \( W_e \) is the electrical speed and \( L \) is inductances (\( L_q = L_d = L \)).

Two Fuzzy logic observers are applied to estimate the back-EMF. Each obtains one of the components of the back-EMF from the estimated current and measured PMSM current. The input of the Fuzzy logic observers are estimated current (the output of the current estimator) and difference between the measured and estimated stator currents. The poising angle is obtained from the output of the Fuzzy logic observers while the back-EMF is calculated by Eq. (66) the position angle can be obtained by

\[
\arctan\left(\frac{-e_\alpha}{e_\beta}\right) = \theta
\] (69)

Eqs. (67) and (68) are applied inside the current estimator to estimate \( I_\alpha \) and \( I_\beta \) from the PMSM terminal voltage and estimated back-EMF while the estimated back-EMF are obtained from the Fuzzy logic observers. The output of each Fuzzy logic is estimated back-EMF (\( e^a_\alpha \) and \( e^a_\beta \)), the estimated position angel can be obtained from Eq. (69) [33]. Because of the function of \( \tan^{-1} \) the estimated position information varies between \( \left(\frac{\pi}{2}, -\frac{\pi}{2}\right) \). A mathematical function was written to convert the estimated angel varies between \( (0, 2\pi) \). The general structure of the Fuzzy position estimator is shown in Figure 22.
Rule base for the Fuzzy logic observer was obtained by evaluating the measured currents and truth degrees of the estimated current. Rules are extracted applying the Mamdani method. Two rules make no sense for operation the Fuzzy logic observer therefore; they are ignored from the rules. The center of area (CoA) defuzzification method is applied for both Fuzzy logic observers. The corresponding rule matrix for the Fuzzy logic observer is shown in Table 7 [33].

Table 7. Fuzzy logic observers rules table

<table>
<thead>
<tr>
<th>$i_a$ or $i_b$</th>
<th>NL</th>
<th>NM</th>
<th>NS</th>
<th>ZE</th>
<th>PS</th>
<th>PM</th>
<th>PL</th>
</tr>
</thead>
<tbody>
<tr>
<td>$i_a$ or $i_b$</td>
<td>NL</td>
<td>NL</td>
<td>NM</td>
<td>Z</td>
<td>PM</td>
<td>PL</td>
<td>PL</td>
</tr>
<tr>
<td></td>
<td>NM</td>
<td>NL</td>
<td>NM</td>
<td>NS</td>
<td>PM</td>
<td>PL</td>
<td>PL</td>
</tr>
<tr>
<td></td>
<td>NS</td>
<td>NM</td>
<td>NS</td>
<td>Z</td>
<td>PS</td>
<td>PM</td>
<td>PM</td>
</tr>
<tr>
<td></td>
<td>ZE</td>
<td>NM</td>
<td>NS</td>
<td>Z</td>
<td>PL</td>
<td>PS</td>
<td>PM</td>
</tr>
<tr>
<td></td>
<td>PS</td>
<td>NM</td>
<td>NS</td>
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<td>PS</td>
<td>PM</td>
<td>PM</td>
</tr>
<tr>
<td></td>
<td>PM</td>
<td>NL</td>
<td>NM</td>
<td>NS</td>
<td>Z</td>
<td>PS</td>
<td>PM</td>
</tr>
<tr>
<td></td>
<td>PL</td>
<td>NL</td>
<td>NM</td>
<td>Z</td>
<td>PM</td>
<td>PL</td>
<td>x</td>
</tr>
</tbody>
</table>
Figure 23. Surface view of Fuzzy logic observers rules.

Figure 24. Membership function of Fuzzy logic observers first input.

Figure 25. Membership function of Fuzzy logic observers second input. (difference between the measured and estimated stator currents $i_a - i_a^*$ or $i_\beta - i_\beta^*$).
Two Fuzzy logic observers are applied in this method to estimate the $e_\alpha$ and $e_\beta$ from the estimated and measured $i_\alpha$ and $i_\beta$. Two Fuzzy logic observers have the same membership function and same rules. Seven membership functions were used for the two inputs and output of two Fuzzy logic observers. Limits of the membership functions were determined by considering the parameters of PMSM. Put into practice, one or two types of membership functions are enough to solve most of the problems. To simplicity, triangular and trapezoidal shapes are used. Figures 24-26 show the memberships function of Fuzzy logic observers.

6. RESULTS AND DISCUSSION

The PI control, Fuzzy logic, and PI tuned by GA are applied for controlling the speed of the PMSM in different speed and load condition to compare the performance of each controller. There are three response characteristics to be observed, namely overshoot, or undershoot during the starting time, settling time, and steady state error. The step function is applied to simulate speed step and torque step for different condition. The performance of the controllers are observed in rated speed motor and rated motor torque to simulate the nominal condition of the motor and observe the performance of each controller for worst condition. The worst condition will happen while the full load was applied during the starting the motor. If the motor was applied to pull up the load during the starting time, the direction of rotation will change and the load will fall down. The simulation proves the superiority of Fuzzy logic for different conditions. In addition, while the Fuzzy logic controller is applied to control the motor speed the performance of the position estimator based on Fuzzy logic...
observer is observed in different speed and load conditions. The rotor position and the actual motor speed are compared to the estimated rotor position and motor speed respectively. The simulation proves well performance of the estimator for different conditions within acceptable error range.

A FLC for PMSM was simulated to improve the motor’s start-up characteristic and reduce the speed overshoot by applying the advantage of the Fuzzy theory. From the simulation results, it is shown that PI controller provided faster setting time. Whereas the PI controller tuned by GA provided smaller overshoot. The FL speed controller provided faster setting time compare to PI controller and smaller overshot compare to PI controller tuned by GA. The simulation has been done under the changes of the load torque from zero to rated torque and the speed changes from zero to (rated speed)/2 = (1000 rpm) and nominal speed (2000 rpm).

Figure 27 shows the step speed responses of the system for three different speed controlling methods (Fuzzy logic, PI and PI tuned by GA) at the rated speed (2000 rpm) while the no load condition was applied. In addition, the results are compared to the first order system response to make better sense for performance of each controller. The hybrid controller provided no overshoot and faster setting time (0.24 second) compare to PI and PI tuned by GA. The speed step was applied in 0.5 second.

Figure 28 shows the comparison between three different controlling methods under the full load torque and rated speed (2000 rpm). The PI controller reached to desired speed slower than the other method under the full load condition. The PI tuned by GA provided faster setting time, but during the starting time it could not performed well and the speed reached to (-20 rpm). If the motor is used to elevate the load during the starting time, the direction of rotating had change and the load will fell down. Figure 5.3 shows the superiority of Fuzzy logic controller at starting time when the full load was applied. Comparison between the Fuzzy logic, PI controller tuned by GA under the different load and speed conditions are shown in Figures 28-35, which confirm the superiority of Fuzzy logic speed controller in comparison with PI speed controller tuned by different techniques.

The actual rotor position and motor speed were compared by the estimated rotor position and motor speed under different load condition to observe the performance of the estimator for different situation. The simulation results are shown that the estimation position is more challenging during the low speed and high torque.
Figure 27. Step responses of PMSM to rated speed (2000 rpm) under no load condition.
Figure 28. Step responses of the PMSM to rated speed (2000 rpm) under full load condition.
Figure 29. Zoom in-view at starting time to rated speed (2000 rpm) under full load condition.
Figure 30. Step speed responses of PMSM from 2000 (rpm) to 1000 (rpm) at 0.5 second.
Figure 31. Step speed responses of PMSM from 1000 (rpm) to 2000 (rpm) at 0.5 second.
Figure 32. Step load responses of PMSM from full load to no load at rated speed (2000 rpm) in 0.5 second.
Figure 33. Step load responses of PMSM from no load to full load at rated speed (2000rpm) in 0.5 second.
Figure 34. Step speed and load response at 1(s), step speed (2000 to 1000 rpm) and step load (full load to no load).
Figure 35. Step speed and load response at same time (1 second), step speed (1000 to 2000 rpm) and step load (no load to full load).
Figure 36. Estimated and real rotor position at rated speed (210 rad/s) under no load condition.
Figure 37. Estimated speed and real measured speed at rated speed (210 rad/s) under no load condition.
Figure 38. Estimated and real rotor position at rated speed (210 rad/s) under full load condition.
Figure 39. Estimated speed and real measured speed at rated speed (210 rad/s) under full load condition.
The comparison between the estimated and actual rotor position at rated speed (210 rad/s) under no load condition is shown in Figure 36. Figure 37 shows the comparison between the actual motor speed and the estimated speed at rated speed (210 rad/s) under no load condition.

The estimated and real rotor position at rated speed (210 rad/s) under full load condition is shown in Figure 38. The actual speed and estimated speed at nominal motor speed (210 rad/s) under full load condition is shown in Figure 37.

7. Conclusion

The dynamic control algorithm for PMSM using Fuzzy logic speed controller and Fuzzy logic position observer was developed for in-wheel applications, and the comparison of the specification performance between conventional PI controller, Fuzzy logic controller (FLC) and PI controller tuned by GA was done. In addition, the sensor less method was applied base on the Fuzzy logic observer to estimate the rotor position and eliminate the usage of encoder to measure the speed, furthermore the FLC was applied to control the speed of the motor while the rotor position is obtained from Fuzzy estimator instead of speed sensor. By applying this method, the advantage of Fuzzy theory was utilized for controlling and estimating the speed of the motor at the same time.

FLC has ability to employ different rules base on the desired output on our designs. In addition, Fuzzy logic (FL) enables us to decide unsymmetrical for positive and negative errors while PI does not allow to differentiate between positive and negative errors, furthermore it is sensitive to motor parameter variation. As what simulation results depict, the FLC has lots of damped undershoots. FL improves the control, removes most of overshoots, and undershoots. It is also very easy to change the FLC rules and membership functions. One problems of Fuzzy logic method is that its steady state response. Since this method does not have any integrator hence any memory, steady state errors will not be removed exactly. The simulation proves the superiority of FLC in both speed and torque variation and steady state condition.

Moreover, by applying the position estimator based on Fuzzy logic, disadvantages of using different estimation methods for lower and higher speeds are eliminated. The proposed estimator base on FL has some similarities with the sliding mode estimations method, while in the estimated method base on Fuzzy logic observer; back-EMFs are obtained from the Fuzzy logic observers.
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instead of the machine model. In the proposed estimation method, the differences between the estimated currents, which are determined by means of back-EMFs, and the measured currents are used as input for the Fuzzy logic position estimator. Therefore, the error that occurs in the system is decreased by this feedback. For the proposed Fuzzy logic observer, the position estimation has been done with acceptable error in the wide speed range.

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