CHAPTER 3

BLDC MOTOR POSITION CONTROL USING PID CONTROLLER

3.1 BASICS OF PID CONTROLLER

There are basically 3 kinds of operational modes that PID controllers use namely: proportional also referred to as P, integrative also referred to as I and derivative also known as D. Zoran & Ognjen (2002) in their study have discussed that proportional as well as integrative modes are primarily utilized as single control modes, whereas the derivative mode on its own as such is seldom used in control systems. Practical systems often deploy PI and PD control combinations.

3.1.1 Proportional Controller

Generally it may be inferred that controller P as such does not have the ability to stabilize those processes that functionally or of a higher order. Kemal et al (2012) in their study show that those processes that basically have one storage energy reserve also referred to as 1st order processes, which generally can tolerate a relatively high increase in gain. The Proportional controller has the capability to stabilize only processes that are unstable and are of the 1st order. Changing controller gain $K$ has the ability and hence can change or alter the dynamics functionally of the closed loop. Controller gain which is of a large nature will effectively result in presenting a control system that is characterized by the following characteristics:
• Steady State error that is relatively smaller or in other words has an improved referential following

• Dynamics that are faster dynamics, or in other words the closed loop system has a broader signal frequency band additionally it has larger sensitivity in terms of its ability to measure noise

• An amplitude that is smaller and also a phase margin

Once controller P controller is deployed, a large gain is required in order to improvise steady state error. Stable systems necessarily do not face problems if large gain has been utilized. These kinds of systems are those that possess single energy storage or are referred to as 1st order capacitive systems. Considering that the steady state error is more or less constant it becomes relatively acceptable and suitable for such processes, than those where controller P may be deployed. Steady state errors that are small become acceptable only if the respective sensor presents an error whose value can be measured or if measure value is not significant enough as presented by Zoran & Ognjen (2002).

3.1.2 Proportional Derivative Controller

D mode is generally deployed when error prediction has the ability to improve control or under the circumstance when the system needs to be stabilized. When observing one notice that characteristic element D frequency has a 90° phase lead.

Normally the derivative necessarily is not obtained from the error signal except from the system output variable. This is carried out to steer clear of effects resulting from the reference input’s sudden change that leads to a sudden change in the error signal value. The sudden change existing in these error signals can further lead to a sudden control output change as presented
in Kemal et al (2012). In order to steer clear of that it is appropriate to thus design D mode as one that is proportional to output variable change.

PD controller generally extensively utilized as a means to control moving objects like vehicles that fly or operate underwater, rockets and ships and other. The main reason in doing so is basically to stabilize PD controller’s effect on heading variable y(t) sudden changes.

3.1.3 Proportional Integral Controller

PI controller has the capability to eliminate forced oscillations as well as steady state error that results in on-off and P controllers’ operation respectively. Nonetheless, through the introduction of the integral mode there is a negative effect that affects response speed and generally the overall system stability. Therefore, PI controller as such cannot increase response as presented by Vinod & Kishor (2013). However this may be anticipated as the PI controller essentially doesn’t possess the ability to predict what will occur when an error takes place in the near future. There is however a way to resolve this through the introduction of the derivative mode which possess the ability to more or less assess and predict what follows next when the error takes place in the near future and thus resolves the problem by decreasing controller’s reaction time.

PI controllers quite popularly utilized in the industry, particularly when response speed is not the main issue. A control that does not have D mode is deployed when:

- It is not mandatory for system to respond fast
- While the process is operational large disturbances as well as noise prevail
• There exists a single energy storage that is either inductive or it is capacitive

• When in the system there are inherently large transport delays

The PID controller is well equipped in terms of possessing the dynamics that are necessary for its effectiveness: when controller input is changed in D mode it reacts quickly, lead error is directed towards zero that is the I mode when there is an increase in the control signal and its fitting action within the area of control error that enables eliminating oscillations in the P mode. The Derivative mode as such helps in improving system’s stability and additionally facilitates in increasing gain K and simultaneously decreasing $T_i$ which is the integral time constant, resulting in increasing controller response speed.

The PID controller basically is deployed to counteract with those processes that have more than a single energy storage or those with a higher order capacity, especially when process dynamics is not like the integrator dynamics as in various thermal processes as presented by Kemal et al (2012). The PID controller is frequently utilized in the industry; additionally it is used in controlling movement of mobile objects which includes course and trajectory ones when precision as well as stability referencing becomes a prerequisite as below.

**Table 3.1 Effects of coefficients**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Speed response</th>
<th>Stability</th>
<th>Accuracy</th>
</tr>
</thead>
<tbody>
<tr>
<td>Increasing $K_p$</td>
<td>Increases</td>
<td>Deteriorate</td>
<td>Improves</td>
</tr>
<tr>
<td>Increasing $K_i$</td>
<td>Decreases</td>
<td>Deteriorate</td>
<td>Improves</td>
</tr>
<tr>
<td>Increasing $K_d$</td>
<td>Increases</td>
<td>Improves</td>
<td>No impact</td>
</tr>
</tbody>
</table>
Table 3.1 visibly shows influences or responses of the PID controller parameters $K_p$, $K_i$ and $K_d$. Evidently as shown in the table every parameter possesses an individualistic importance in terms of its ability to control system’s accurate responses, stability as well as speed.

This study main focus is sensorless BLDC motor speed control by deploying PID controller throughout the voltage magnitude that varies at intervals across the BLDC motor.

3.2 PROPOSED ARCHITECTURE OF SENSORLESS BLDC MOTOR

Figure 3.1 shows sensorless BLDC motor’s proposed architectural framework. In the BLDC motor VSI or the voltage source inverter has an important role for generating flux and also fixing for the motor angle ‘$\theta$’ to facilitate its operation. On the basis of the BLDC motor principle, there inherently exists a supply among the two phases. Emf is generated alongside one of the phases. The back emf has a basic sensing capability which helps in generating pulse after zero crossing detection that is basically the position encoding. On the basis of the switching table, switches operative decision can thus be controlled.
3.2.1 BLDC Motor Mathematical Modelling

The three phase star connected BLDC motor can be described by the following four equations

\[ v_{ab} = R(i_a - i_b) + L \frac{d}{dt}(i_a - i_b) + e_a - e_b \]  
\[ v_{bc} = R(i_b - i_c) + L \frac{d}{dt}(i_b - i_c) + e_b - e_c \]  
\[ v_{ca} = R(i_c - i_a) + L \frac{d}{dt}(i_c - i_a) + e_c - e_a \]  
\[ T_e = k_f \omega_m + J \frac{d\omega_m}{dt} + T_L \]

The symbol \( v, I \) and \( e \) denote the phase-to-phase voltages, currents and phase back-emf’s respectively, in the three phase \( a, b \) and \( c \). The
resistance $R$ and the inductance $L$ are per phase values and $T_e$ and $T_L$ are the electrical torque and the load torque. $J$ is the rotor inertia, $k_f$ is a friction constant and $\omega_m$ is the rotor speed. The back emf and the electrical torque can be expressed as

$$e_a = \frac{k_e}{2} \omega_m F(\theta_e)$$

(3.5)

$$e_b = \frac{k_e}{2} \omega_m F(\theta_e - \frac{2\pi}{3})$$

(3.6)

$$e_c = \frac{k_e}{2} \omega_m F(\theta_e - \frac{4\pi}{3})$$

(3.7)

$$T_e = \frac{k_t}{2} [F(\theta_e) i_a + F(\theta_e - \frac{2\pi}{3}) i_b + F(\theta_e - \frac{4\pi}{3}) i_c]$$

(3.8)

Respectively, where $k_e$ and $k_t$ are the back-emf constant and the torque constant. The electric angle $\theta_e$ is equal to the rotor angle times the number of pole pairs ($\theta_e = \frac{p}{2} \theta_m$). The function $F(\theta_e)$ gives the trapezoidal waveform of the back-emf. One period of this function can be written as

$$F(\theta_e) = \begin{cases} 
1, & 0 \leq \theta_e < \frac{2\pi}{3} \\
1 - \frac{6}{\pi} \left( \theta_e - \frac{2\pi}{3} \right) & \frac{2\pi}{3} \leq \theta_e < \pi \\
-1 & \pi \leq \theta_e < \frac{5\pi}{3} \\
-1 + \frac{6}{\pi} \left( \theta_e - \frac{5\pi}{3} \right) & \frac{5\pi}{3} \leq \theta_e < 2\pi
\end{cases}$$

(3.9)

The equations here need a little modification so that state-space representation is permissible. As each and every voltage equation is basically a linear combination of the other, hence what is required is only a two voltage
equation. If one of the equations is discarded and one variable eliminated the following current relationship may be used:

\[ i_a + i_b + i_c = 0 \]  \hspace{1cm} (3.10)

The voltage equation becomes

\[ v_{ab} = R(i_a - i_b) + L \frac{d}{dt} (i_a - i_b) + e_a - e_b \]  \hspace{1cm} (3.11)

\[ v_{bc} = R(i_a + 2i_b) + L \frac{d}{dt} (i_b + 2i_b) + e_b - e_c \]  \hspace{1cm} (3.12)

and the complete model is then

\[
\begin{pmatrix}
i_a \\
i_b \\
o_{m}' \\
\theta_m'
\end{pmatrix}
= \begin{pmatrix}
-\frac{R}{L} & 0 & 0 & 0 \\
0 & -\frac{R}{L} & 0 & 0 \\
0 & 0 & -\frac{k_f}{J} & 0 \\
0 & 0 & 1 & 0
\end{pmatrix}
\begin{pmatrix}
i_a \\
i_b \\
o_{m}' \\
\theta_m'
\end{pmatrix}
+ \begin{pmatrix}
\frac{2}{3L} & 1/3L & 0 \\
-1/3L & 1/3L & 0 \\
0 & 0 & 1 \\
0 & 0 & 0
\end{pmatrix}
\begin{pmatrix}
v_{ab} - e_{ab} \\
v_{bc} - e_{bc} \\
T_e - T_L
\end{pmatrix}
\]  \hspace{1cm} (3.13)

\[
\begin{pmatrix}
i_a \\
i_b \\
i_c \\
o_{m}' \\
\theta_m'
\end{pmatrix}
= \begin{pmatrix}
1 & 0 & 0 & 0 \\
0 & 1 & 0 & 0 \\
0 & 0 & 1 & 0 \\
0 & 0 & 0 & 1
\end{pmatrix}
\begin{pmatrix}
i_a \\
i_b \\
i_c \\
o_{m}' \\
\theta_m'
\end{pmatrix}
\]  \hspace{1cm} (3.14)

It is but quite often that rotating reference frames are machine models that have been transformed for simplifying purposes and also to
improvise overall computational efficiency. However the above approach has not been deployed for the study as it has proven that supply voltage not as such sinusoidal, respective transformation will not be in a position to improvise overall computational efficiency.

3.2.2 BLDC Motor Control

The basic means to acquiring an effective torque and BLDC motor speed control is primarily based on the torque that is relatively simple torque along with back EMF equations that are almost same as the DC motors.

The back EMF magnitude can be written as:

\[ E = 2NlrB\omega \]  

(3.15)

and the torque term as:

\[ T = \left( \frac{1}{2} i^2 \frac{dL}{d\theta} \right) - \left( \frac{1}{2} B^2 \frac{dR}{d\theta} \right) + \left( \frac{4N}{\pi} Brl\pi i \right) \]  

(3.16)

where \( N \) is the number of winding turns per phase, \( l \) is the length of the rotor, \( r \) is the internal radius of the rotor, \( B \) is the rotor magnet flux density, \( \omega \) is the motor’s angular velocity, \( i \) is the phase current, \( L \) is the phase inductance, \( \theta \) is the rotor position, and \( R \) is the phase resistance.

In the torque expression the two terms that come initially are basically components that are parasitic reluctance torque components. The term that is appears third produces shared torque, which really is the BLDC motor’s torque production mechanism. Overall, motor speed and back EMF are directly proportional to one another and production of torque with respect to the phase current is more or less directly proportional.
BLDC motor functions basically by a two phase ON operation that enables inverter control. The torque production as per the control scheme, functions on the particular principle wherein current flows in three phases at any given point of time and wherein no torque production should ideally be present especially in the back EMF zero crossings. Figure 3.2 illustrates formation of BLDC motor electrical wave in two of the phases that are operational.

There are many advantages that this particular control structure has:

- At any given point of time only one current requires controlling
- A single current sensor is required as such (or for speed loop none are required, as explained in following sections)
- Current sensor positioning permits use of shunt which are basically low cost sensors

BLDC motor operating principle is to make sure that the phase pair is energized generally throughout the process at all times so that highest torque is produced. The shape of the back Emf is trapezoidal to ensure that the effect is optimized. This blend of trapezoidal back EMF in the DC current theoretically makes constant torque feasible. However practically, in any motor phase establishing current does not take place instantaneously; as a result of which at every 60° phase commutation there is a presence of a torque ripple.
Figure 3.2  Electrical waveforms in the two phases on operation and torque ripple

This control is applicable only when motor deployed has a sinusoidal back EMF shape, as a consequence produced torque is:

Irregular but formed from sine wave portions. This is because it is a combination of both the trapezoidal current control strategy as well as the sinusoidal back EMF. It is important to take note of the fact that a sinusoidal back EMF shape motor controlled that has a sine wave strategy (three phase ON) is one where constant torque is produced.

3.2.3  Back emf (Bemf) Zero Crossing Point Computation

Resistor divider circuit specifications are configured in a manner that facilitates maximized output from the voltage sensing circuit which deploys ADC complete conversion range. Filtering capacitor must have the
capability wherein chopping frequency is filtered, hence only those values that are very small are required (within nF range or perhaps even lesser). The basis for the Sensorless algorithm is the three motor terminal voltage measurements and wherein just four ADC input lines are necessary.

The phase commutation takes place only once at the rotor’s 60° mechanical rotation in this sensored control structure. This basically means that in terms of sufficiency six commutation signals are adequate in driving the BLDC motor. Additionally, an efficient control encompasses synchronization between both the phase supply as well as Bemf such that Bemf in the non-fed 60° sector crosses zero at least once. The following details show the possibility of getting three Bemfs and their respective zero crossings. As per the motor terminal model, \( R \) refers to the phase resistance, \( L \) refers to the phase inductance, \( V_n \) refers to the star connection voltage that is referenced to ground, \( E \) refers to the back electromotive force, and \( V_x \) refers to the phase voltage referenced to ground. \( V_x \) Voltages are measured using ADC unit and through resistance bridge.

Presuming phase \( C \) as non-fed phase, with respect to the three terminal voltages the following equations are possible:

\[
V_a = R\dot{\alpha} + L\frac{dla}{dt} + Ea + V_n
\]  
(3.17)

\[
V_b = R\dot{b} + L\frac{dlb}{dt} + Eb + V_n
\]  
(3.18)

\[
V_c = Ec + V_n
\]  
(3.19)
The currents at the two phases are both opposite to each other and equal as there are only two currents that flow through the stator windings at any given point of time. Hence,

\[ l_a = -l_b \]  

(3.20)

Therefore, by adding the three terminal voltage Equations, we have:

\[ V_a + V_b + V_c = E_a + E_b + E_c + 3V_n \]  

(3.21)

**Table 3.2 Switching sequence of BLDC motor**

<table>
<thead>
<tr>
<th>Switching Interval</th>
<th>Seq. Number</th>
<th>Switch Closed</th>
<th>Phase Current</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
<td>A</td>
</tr>
<tr>
<td>0°-60°</td>
<td>0</td>
<td>Q1</td>
<td>Q4</td>
</tr>
<tr>
<td>60°-120°</td>
<td>1</td>
<td>Q1</td>
<td>Q6</td>
</tr>
<tr>
<td>120°-180°</td>
<td>2</td>
<td>Q3</td>
<td>Q6</td>
</tr>
<tr>
<td>180°-240°</td>
<td>3</td>
<td>Q3</td>
<td>Q2</td>
</tr>
<tr>
<td>240°-300°</td>
<td>4</td>
<td>Q5</td>
<td>Q2</td>
</tr>
<tr>
<td>300°-360°</td>
<td>5</td>
<td>Q5</td>
<td>Q4</td>
</tr>
</tbody>
</table>

As shown in Table 3.2 on the basis of the switching sequence, BLDC generated speed requires controlling using an appropriate controlling technique. Research work that has been presented here deploys PID controlling technique so the BLDC motor speed is controlled.

### 3.3 PROPOSED SPEED CONTROL TECHNIQUE OF BLDC MOTOR THROUGH PID CONTROLLER

In terms of importance PID controller so the most commonly used control algorithm. PID control or its minor variations form the basis for many
practically designed feedback loops. There quite a few controllers that necessarily don’t utilize any form of derivative action. PID controllers apparently have several different forms, for example they appear as stand-alone controllers, or these controllers can be a part of the Direct Digital Control (DDC) package or even they could be inherent in hierarchical distributed process control system or incorporated into embedded systems. There are practically thousands of engineers who study instrument and control all over the world who utilize these kind of controllers as part of their day-to-day tasks or routine as presented by Desborough & Miller (2002). PID algorithm can be viewed from varying directions using different approaches. In terms of being a device there are several viewpoints under which it may be seen in light of its operative functionalities under a couple of empirical rules, however it may also be looked at using a more analytical approach. Figure 3.3 shows a block diagram of the PID Controller along with the error feedback.

![Figure 3.3 PID Controller with error feedback](image)

Figure 3.3 PID Controller with error feedback

The PID Controller can be described by the following equation.
\[ u = k_p e(t) + k_i \int_0^t e(\tau) d\tau + k_d \frac{d e}{d t} \] (3.22)

where \( u \) is the control signal from the BLDC motor and \( e \) is the control error \((e = y - r)\). The reference value is also called the set point as discussed in Desborough & Miller (2002). The control signal is thus a sum of three terms: the P-term (which is proportional to the error), the I-term (which is proportional to the integral of the error), and the D-term (which is proportional to the derivative of the error). The controller parameters are proportional gain \( k_p \), integral gain \( k_i \) and derivative gain \( k_d \). The controller can also be parameterized as

\[ u(t) = k_p \left( e(t) + \frac{1}{T_i} \int_0^t e(\tau) d\tau + T_d \frac{de(t)}{dt} \right) \] (3.23)

where \( T_i \) is called integral time and \( T_d \) derivative time. The section that is proportional works on the error’s present value, wherein the integral is symbolic of the past errors average and here the derivative may also be interpreted as being the future error prediction on the basis of the linear extrapolation.

3.3.1 Proportional Action

Figure 3.4 below shows output response to BLDC motor’s speed \( \omega_m \) that has pure proportional control as previously discussed and presented by Desborough & Miller (2002). Here output almost never is at a state where it can reach steady state error. Considering transfer functions for both the process as well as controller to be \( P(s) \) and \( C(s) \). Equation (3.24) here gives
the transfer function that exists from point of reference to the output as presented by Karl & Richard (2009).

\[
G_{yr} = \frac{P(s) C(s)}{1 + P(s) C(s)}
\]

(3.24)

The steady state error for a unit step is

\[
G_{yr}(0) = \frac{k_p P(0)}{1 + k_p P(0)}
\]

(3.25)

![Figure 3.4 A PID controller takes control action based on past, present and prediction of future control errors](image)

Here increasing gain results in an error decrease, however the system develops a tendency making it more oscillatory. As per figure 3.4 it is clear that the control signal’s initial value is equal to the controller gain. Hence to avoid steady state error, proportional controller may be altered as per the equation that is below as presented by Karl & Richard (2009).

\[
u(t) = k_p e(t) + u_b
\]

(3.26)
Here $u_b$ refers to the bias or reset term which may be adjusted in order to produce desired steady state value.

### 3.3.2 Integral Action

The integral action here more or less ensures that process output is in agreement with reference having a steady state which may be illustrated as below. Here the systems are assumed to possess a steady state having a constant control signal ($u_0$) and constant error $e_0 \neq 0$.

It follows from Equation (3.6) that

$$u_0 = k_pe_0 + k_ie_0t \quad (3.27)$$

Here left hand side as such has a constant state however the right one is basically one of the functions of $t$. Hence, it is considered as a contradiction wherein $e_0$ has to be zero. The observations derived from the argument indicate that the only possible assumption from this is that a steady state exists as presented by Desborough & Miller (2002).

![Diagram of integral action as automatic biasadjustment](image)

**Figure 3.5** Implementation of integral action as automatic biasadjustment

Among other arguments one of them is the controller’s transfer function that has an integral action of an infinite gain at zero frequency ($C(0) \propto 1$). Figure 3.5 then shows that $G_{yr}(0)=0$. However the aforementioned
argument has to be supported by the fact that the system in operation here is linear.

The integral action here may also be considered as a methodology to generate bias term \( u_b \) in the proportional controller (3.26) automatically. Figure 3.4 shows how bias \( u_b \) has been generated by permitting low pass filtering through the output. This particular implementation is referred to as automatic reset, which apparently was among the list of integral control early inventions. Figure 3.5 showing the system’s transfer function has been obtained through the process of loop tracing. Presuming exponential signals as well as tracing these that go round the loop then gives the following:

\[
u = k_p e + \frac{1}{1 + sT} u
\] (3.28)

Solving for \( u \) gives

\[
u = k_p \frac{1 + sT}{sT} e
\]

\[= (k_p + \frac{k_p}{sT})
\]

Which is the transfer function of a PID controller.

The proportional gain is constant, \( k_p = 1 \) and the integral gain is changed. The case \( k_i = 0 \) corresponds to pure proportional control; with a steady state error is 50%. The steady state error is removed when integral gain \( k_i \) is increased. The response creeps slowly towards the reference for small values of \( k_i \). The approach apparently is relatively quicker for the larger integral gains additionally the system is more oscillatory thus.
3.3.3 Derivative Action

Figure 3.6 here presents how derivative action has the ability to improvise closed-loop system stability. Hence input-output relation between controller with the proportional as well as the derivative action is as follows:

\[ u(t) = k_p e(t) + k_d \frac{de}{dt} \]

(3.29)

\[ = k_p (e(t) + T_d \frac{de}{dt}) \]

\[ = k_p e_p(t) \]

where \( T_d = \frac{k_d}{d} \) is the derivative time. Interpretation of the controller’s action with proportional and derivative action may read as if the control is made proportional to the predicted process output, whereas prediction is derived from error \( T_d \) time units extrapolating into the future by deploying tangent to the error curve. System is generally oscillatory when the derivative action is not utilized and it is kind of damped as there is an increase in the derivative gain.

3.3.4 Filtering the Derivative

Major disadvantage with derivative action is that ideal derivative needs to have a relatively high gain for the high frequency signals. This basically implies that high frequency measurement noise would result in generating control signal variations that large as shown by Hemchand (2013). Measurement noise effect may however be lessened by simple replacement of term \( k_{ds} \) by
\[ D_a = -\frac{k_{ds}}{1 + sT_f} \]  \hspace{1cm} (3.30)

**Figure 3.6** Implementation of the transfer function \( sT = (1 + sT) \) which approximates derivative action

This can be interpreted as an ideal derivative that is filtered using a first order system with the time constant \( T_f \). For small \( s \), the transfer function is approximately \( k_{ds} \) and for large \( s \) it is equal to \( k_d = T_f \). The approximation acts as a derivative for low-frequency signals and as a constant gain for the high frequency signals. The high-frequency gain is \( k_d = T_f \). The filtering time is chosen as \( k_d=\kappa=N \), with \( N \) in the range of 2 to 20. The transfer function of a PID controller with a filtered derivative is

\[ C(s) = K (1 + \frac{1}{sT_i} + \frac{sT_d}{1 + sT_d/N}) \]  \hspace{1cm} (3.31)

The high-frequency gain of the controller is \( K(1 + N) \).

As an alternative for derivative filtering, one could look at the possibility of using an ideal controller as well as filter measured signal. This controller transfer function with the filter can be presented as follows:

\[ C(s) = K \left( 1 + \frac{1}{sT_i} + sT_d \right) \frac{1}{(1 + sT_f)^2} \]  \hspace{1cm} (3.32)
Where, a second order filter is used.

Among the earlier implementations one of them is presented whose derivative is shown in Figure 3.6. In the particular system, derivative is really the difference between both signal and its filtered version. System’s transfer function may be shown as:

\[
C(s) = \frac{1}{1 + sT} \\
\]  

\[
= \frac{sT}{1 + sT} \cdot u(s) \quad (3.33)
\]

The system’s transfer function thus is \( G(s) = \frac{sT}{(1 + sT)} \) which basically estimates low frequency derivative. The observation on the basis of this indicated that implementation automatically generates filtering. Therefore BLDC motor speed is controlled using PID controller by voltage magnitude variations voltage that are across BLDC motor with parameters such as \( k_p, k_i \) and \( k_d \). Figure 3.7 shows the overall working of proposed system.
Figure 3.7  Flowchart of the proposed speed control technique using PID Controllers

Step 1: Supply initialization fed in to the inverter

Step 2: Back emf/voltage sensing, $e_\alpha, e_\beta, e_\gamma$
Step 3: Zero Crossing Detection to attain pulses

Step 4: BLDC motor Position Identification ‘θ’

Step 5: Sequential application in the inverter using the switching table

Step 6: Flux generation, Voltage $V_α, V_β, V_γ$ and Current $I_α, I_β, I_γ$

Step 7: BLDC motor rotation happens on the basis of flux, which further generates speed, $ω_m$.

Step 8: Speed, $ω_m$ is controlled using PID controller by voltage magnitude variations across the BLDC motor

Step 9: In order to control BLDC motor speed PID control parameters $k_p, k_i, k_d$ and saturation limit which basically refers to the peak to peak voltage is altered.

3.4 SUMMARY

In chapter 3 various types of controllers have been discussed like P, PI, PD and PID. Generally controller description presents the relevance as well as efficiency of the PID controller which have been discussed in chapter 3 here. Proposed methodology herein gives a brief description of PID controller utilization for controlling BLDC motor speed which has been elaborately presented in this particular chapter.