State Feedback Linearization of a Non-linear Permanent Magnet Synchronous Motor Drive

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Abstract
Control system design for inverter fed drives previously used the classical transfer function approach for single-input single-output (SISO) systems. Proportional plus Integral (PI) controllers were designed for individual control loops. It is found that the transient response of a PI controller is slow and is improved by pole placement through state feedback. However, the effective gains of the PI controller are substantially decreased as a function of the increase of motor speed. A control system is generally characterized by the hierarchy of the control loops, where the outer loop controls the inner loops. The inner loops are designed to execute progressively faster. The speed controller (PI controller) processes the speed error and generates the reference torque. In the inner loop, firstly a non-linear controller is designed for the system by which the system nonlinearity is canceled using state or exact feedback linearization. In addition, a linear state feedback control law based on pole placement technique including the integral of output error (IOE) is used in order to achieve zero steady state error with respect to reference current specification, while at the same time improving the dynamic response. The proposed scheme has been validated through extensive simulation using MATLAB.

Keywords: Permanent Magnet Synchronous Motor, Non-linear controller, PI controller, State feedback controller, Integral of Output Error

1. Introduction
Although ac drives require advanced control techniques for control of voltage, frequency and current, they have many advantages over dc drives like reduced power line disturbances, lower power demand on start, controlled acceleration, controlled starting current, adjustable operating speed and adjustable torque. The permanent magnet motors are similar to the salient pole motors, except that there is no field winding and the field is provided instead by mounting permanent magnets in the rotor. The equations of the salient pole motors may be applied to the PM motors, if the excitation voltage is maintained constant. Due to this, there is no excitation voltage source, field winding, collecting rings and brushes; resulting in improving efficiency when compared to other machines. By considering various features such as good dynamic performance, easy controllability, high torque to inertia ratio, high efficiency and improved power factor, Permanent Magnet Synchronous Motor (PMSM) drives [1-2] are used in robotics, machine tools, pumps, ventilators, compressors etc.

The mathematical model of a PMSM [1-2] is non-linear and cannot be represented in linear state space form. Thus, the conventional control system design techniques are not applicable to this system directly. Isidori [3] and M. Ilic-Spong et al. [4] developed the concept of dynamic feedback linearization to switched reluctance motor. KS Low et al. [5] applied the feedback linearization [6] technique to transform the nonlinear equations into a linear time invariant state model for a PMSM. The state transformation is essentially the familiar d-q transformation, whilst the non-linear feedback law performs decoupling and compensation for the influence of back emf in the motor. Zribi and Chiasson [7] proposed exact linearization for position control of PM stepper motor. Jun Zhang et al. [8] discuss decoupling control applied to PMSM using exact linearization. AK Parvathy et al. [9] applied quadratic linearization to PMSM, since PMSM can be adequately described by a quadratic model during normal operation. Safieh Izad and Mahmood Ghanbari [10] discussed speed control of permanent magnet synchronous motor using feedback linearization method.
In this paper, exact linearization of the model of a PMSM with damper windings has been attempted. The proposed controller represented in the conventional two-loop structure [11] for the motor drive is shown in Figure 1. The outer loop is the speed controller, the output of which is the reference value of the torque, $T_e^*$. From this value, the reference values of the currents such as $i_{qs}^*$ and $i_{ds}$ are computed for a desired internal angle $(\psi)$ and a desired torque angle $\delta$. This gives rise to the flexibility in choosing the power factor of the motor from lagging to leading values including unity. The field oriented control [12] can also be obtained as a special case, by setting the power factor angle to be equal to the torque angle, resulting in complete decoupling between the armature flux and the field flux, thus, producing a dc motor like behavior. In this sense, the proposed control scheme is more general than conventional field oriented control. The inner (current) loop is then considered. Here, firstly a non-linear controller is designed for the system by which the system nonlinearity is canceled. In addition to this, a linear state feedback control law [13-14] based on pole placement technique including the integral of output error (IOE) is used in order to achieve zero steady state error with respect to reference current specification, while at the same time improving the dynamic response [2].

2. Mathematical Modelling of PMSM

In order to design a control system for high performance drive, the mathematical model [17-18] of the machine is very much essential. To develop mathematical model of PMSM, the actual machine in a-b-c reference frame [19] is converted into d-q axis representation. By using mathematical modelling, the complexity of calculations is reduced while analyzing the system performance of any machine. Also the time variant inductance is treated as time invariant inductance and the sinusoidal quantities are represented as dc quantities. The schematic diagram of PMSM with damper windings is as shown in Figure 2. The model of PMSM with damper winding has been developed on rotor reference frame using d-q axis [19] representation.
The modeling equations of PMSM in rotor reference frame are given as below:

\[ v_{qs} = r_a i_{qs} + l_{qs} p i_{qs} + l_{aq} p i_{qr} + \omega_r l_{ds} i_{ds} + \omega_r l_{ad} i_{dr} \]  
\[ v_{ds} = r_a l_{ds} + l_{ds} p i_{ds} + l_{ad} p i_{dr} - \omega_r l_{qs} i_{qs} - \omega_r l_{aq} l_{qr} \]  
\[ v_{dr} = r_a l_{dr} + l_{dr} p i_{dr} + l_{ad} p i_{ds} \]  
\[ v_{qr} = r_a l_{qr} + l_{qr} p i_{qr} + l_{aq} p i_{qs} \]  

The electrical torque developed is,

\[ T_e = \frac{3}{2} \times \frac{P}{2} \left( l_{ad} - l_{aq} \right) i_{ds} l_{qs} + l_{ad} l_{qs} l_{dr} - l_{aq} l_{qr} l_{ds} \]  

The torque balance equation of the given system is

\[ p \omega_r = \frac{P}{2J} \left[ T_e - T_l - \frac{2}{P} J \omega_r \right] \]  

The above equations can be written in matrix form as,

\[
\begin{bmatrix}
  v_{qr} \\
  v_{ds} \\
  v_{dr}
\end{bmatrix} =
\begin{bmatrix}
  r_a + l_{qs} p & \omega_r l_{ds} & l_{aq} p & \omega_r l_{ad} & i_{qs} \\
  -\omega_r l_{qs} & r_a + l_{ds} p & -\omega_r l_{qs} & l_{ad} p & i_{ds} \\
  l_{aq} p & 0 & r_{qr} + l_{qr} p & 0 & i_{qr} \\
  0 & l_{ad} p & 0 & r_{dr} + l_{dr} p & i_{dr}
\end{bmatrix}
\begin{bmatrix}
  i_{qr} \\
  i_{ds} \\
  i_{dr}
\end{bmatrix}
\]  

Now to bring these equations in terms of state space representation and the modified equations as,
From the above equation we can define the following matrices for simplification,

$$\begin{bmatrix} l_{qs} & 0 & l_{aq} & 0 \\ 0 & l_{ds} & 0 & l_{ad} \\ l_{aq} & 0 & l_{qr} & 0 \\ 0 & l_{ad} & 0 & l_{dr} \end{bmatrix} \begin{bmatrix} p_{l_{qs}} \\ 0 \\ p_{l_{ds}} \\ p_{l_{qr}} \end{bmatrix} = \begin{bmatrix} r_a & \omega_a l_{ds} & 0 & \omega_a l_{ad} \\ -\omega_a l_{qs} & r_a & -\omega_a l_{aq} & 0 \\ 0 & 0 & r_{qr} & 0 \\ 0 & 0 & 0 & r_{dr} \end{bmatrix} \begin{bmatrix} i_{qs} \\ i_{ds} \\ i_{qr} \\ i_{dr} \end{bmatrix} + \begin{bmatrix} v_{qs} \\ v_{ds} \\ v_{qr} \\ v_{dr} \end{bmatrix}$$  \hfill (8)

Thus, equation (8) can be written in the form,

$$A_x = \begin{bmatrix} r_a & \omega_a l_{ds} & 0 & \omega_a l_{ad} \\ -\omega_a l_{qs} & r_a & -\omega_a l_{aq} & 0 \\ 0 & 0 & r_{qr} & 0 \\ 0 & 0 & 0 & r_{dr} \end{bmatrix}$$  \hfill (9)

$$A_y = \begin{bmatrix} l_{qs} & 0 & l_{aq} & 0 \\ 0 & l_{ds} & 0 & l_{ad} \\ l_{aq} & 0 & l_{qr} & 0 \\ 0 & l_{ad} & 0 & l_{dr} \end{bmatrix}$$  \hfill (10)

$$x = \begin{bmatrix} i_{qs} \\ i_{ds} \\ i_{qr} \\ i_{dr} \end{bmatrix}$$  \hfill (11)

$$B_x = \begin{bmatrix} 1 & 0 \\ 0 & 1 \\ 0 & 0 \\ 0 & 0 \end{bmatrix}$$  \hfill (12)

Thus, equation (8) can be written in the form,

$$A_y \dot{x} = A_x x + B_x u$$  \hfill (13)

or it will be modified as,

$$\dot{x} = Ax + Bu$$  \hfill (14)

with $$A = (A_y^{-1} A_x)$$ & $$B = (A_y^{-1} B_x)$$

3. Design of the Speed Controller (PI Controller)

The basic assumption in separating the speed loop from the overall closed loop system (figure 1) is that the dynamics of the current controller is sufficiently fast, so that no appreciable change in the speed takes place during its transient phase. This has to be ensured by design and implies that the closed loop band-widths of these two loops must differ by at least a factor of ten. A proportional-cum-integral (PI) controller is used for this loop. The output of the PI controller is the reference torque $$T_{e^*}$$, from which the reference currents, $$i_{qs}^*$$ and $$i_{ds}^*$$ can be generated. The design of the gain constants of this controller is as follows:

Considering the torque balance equation (6) involving speed (i.e., mechanical part),

$$p \omega_e = \frac{P}{2J} \left[ T_e - T_i - \frac{2}{P} J \omega_e \right]$$  \hfill (15)

And the torque balance equation for no. of poles, P=4 is taken as
\[ p \omega_e = \frac{2}{J}[T_e - T_i - \frac{\beta \omega_e}{2}] \] (16)

The equation of PI controller is

\[ T_e^* = k_p e + k_i \int_0^t e \, dt \] (17)

Where,

\[ e = (\omega_e - \omega_r) \] (18)

Here \( \omega_e \) is the set speed, \( \omega_r \) is the reference speed and \( k_p \) and \( k_i \) are the proportional and integral gains of the PI controller respectively.

Substituting (17) and (18) in (16) and taking Laplace transform, we get

\[ (s \omega_e - \omega_{e0}) = \frac{2}{J} \left[ \left( k_p + \frac{k_i}{s} \right) (\omega_e - \omega_r) - T_i - \left( \frac{3}{2} \right) \omega_r \right] \] (19)

For \( T_i = 0 \) and \( \omega_{e0} = \omega_e \) rearranging the terms in equation (19),

\[ \left[ s + \frac{\beta}{J} + \frac{2}{J} \left( \frac{k_p}{s} + \frac{k_i}{s} \right) \right] \omega_e = \left[ \frac{2}{J} \left( \frac{k_p}{s} + \frac{k_i}{s} \right) + 1 \right] \omega_r \] (20)

From which the ratio, \( \left( \frac{\omega_r}{\omega_e} \right) \) is obtained as

\[ \frac{\omega_r}{\omega_e} = \frac{2}{J} \left( \frac{k_p}{s} + \frac{k_i}{s} \right) + 1 \left[ s + \frac{\beta}{J} + \frac{2}{J} \left( \frac{k_p}{s} + \frac{k_i}{s} \right) \right] = s^2 + \left( \frac{\beta}{J} + \frac{2}{J} k_p \right) s + \frac{2}{J} k_i \] (21)

This is the standard form of transfer function for a second order system and the denominator can be represented in the form

\[ s^2 + 2 \xi \omega_n s + \omega_n^2 = 0 \] (22)

where

\[ \xi = \text{desired value of damping ratio, and} \]

\[ \omega_n = \text{desired value of natural frequency} \]

The characteristics of the above system is

\[ s^2 + \left( \frac{\beta}{J} + \frac{2}{J} k_p \right) s + \frac{2}{J} k_i = 0 \] (23)

Therefore, equating the corresponding terms in equations (22) and (23)

\[ \omega_n^2 = \frac{2 k_i}{J} \] (24)
\[ 2\xi\omega_n = \frac{\beta}{2} + \frac{2k_p}{J} \]  

(25)

The value of \( \xi \) is usually determined from the requirement of permissible maximum overshoot and the un-damped natural frequency, \( \omega_n \), determines the time response. The controller gains, \( k_i \) and \( k_p \), are obtained as,

\[ k_i = \frac{J}{2} \omega_n^2 \]  

(26)

\[ k_p = J\xi\omega_n - \frac{\beta}{2} \]  

(27)

Assigning proper values of \( \xi \) and \( \omega_n \) and using the values of \( J \) and \( \beta \), the numerical values of proportional and integral gain constants can be computed.

4. State Feedback Linearization

It is evident from equations (1) - (4) that the system matrices representing the electrical subsystem of PMSM are functions of \( \omega_r \), which varies with the operating point and makes the system model coupled and non-linear. Thus, standard techniques of linear system theory cannot be applied directly to design the control system in this situation. To overcome this problem, feedback linearization has been suggested by Isidori [3]. The central idea of the approach is to transform a non-linear model into a linear one by state feedback to which linear control techniques can be applied.

The system model using only the voltage equations (i.e, the electrical subsystem) is expressed as

\[ \dot{x} = Ax + Bu \]  

(28)

Partitioning \( A \) into \( A_1 \) and \( A_2 \)

\[ \dot{x} = (A_1 + \omega_r A_2)x + Bu \]  

(29)

Thus, the system matrix \( A \) in equation (29) has a term proportional to \( \omega_r \). To cancel this, a feedback term is needed, which depends on the product \( \omega_r x \). Choose a feedback control law of the form,

\[ u = u_1 + u_2 \]  

(30)

where \( u_1 \) and \( u_2 \) are the input control vectors of the non-linear and linear parts respectively. The non-linear feedback control law is chosen as

\[ u_1 = \omega_r k_1 x \]  

(31)

where \( k_1 \) is the feedback gain matrix.

Substituting (30) and (31) in (29),

\[ \dot{x} = (A_1 + \omega_r A_2)x + B(u_1 + u_2) \]

Or \( \dot{x} = A_1 x + B u_2 + \omega_r (A_2 + B k_1) x \)  

(32)

In order to get exact cancelation of the non-linear term,
\[ A_2 + B k_1 = 0 \]  
(33)

Or \[ A_2 = -B k_1 \]  
(34)

If \( k_1 \) is taken as

\[
k_1 = \begin{bmatrix} 0 & l_{ds} & 0 & l_{ad} \\ -l_{qs} & 0 & -l_{aq} & 0 \end{bmatrix}
\]  
(35)

Then, equation (34) is satisfied.

Thus, equation (32) changes to the standard linear form

\[ \dot{x} = A_1 x + B u_2 \]  
(36)

Alternatively, one can choose,

\[ u = (\omega_r - \omega_{d1}) k_1 x + u_2 \]  
(37)

where \( \omega_{d1} \) is a design constant, which can be chosen for a trade off between the linear and non-linear components of the control signal. Substituting equation (37) in (32),

\[ \dot{x} = (A_1 + \omega_{d1} A_2)x + B u_2 = A_d x + B u_2 \]  
(38)

Where

\[ A_d = A_1 + \omega_{d1} A_2 \]  
(39)

Thus, the system non-linearity is exactly cancelled. This linearization is valid for all operating points.

5. Results and Discussions

Figure 3. Simulation results of the state feedback controller with and without feedback linearization
Figure 4. Simulation results of the drive system for different values of 
(i) $\delta = 8.735^0$ (unity pf)  (ii) $\delta = 5^0$ (lagging pf)  (iii) $\delta = 15^0$ (leading pf)

Figure 5. Simulation results of the drive system for different values of 
(i) $\psi = -19.10^0$ (unity pf) (ii) $\psi = 5^0$ (lagging pf) (iii) $\psi = -30^0$ (leading pf)

Figure 3 clearly shows that the transient responses have improved with feedback linearization. The initial overshoots in the currents are reduced and steady state values are achieved fastly with the non-linear controller. For a wider change in speed reference, the linear controller fails, but the proposed one continuous to work. The simulation results of the proposed controller as shown in Figure 4 for different values of $\delta$, the currents are settled at different steady state values. Figure 5 shows the simulation results of the proposed controller for different values of $\Psi$ resulting in variation of power factor from lagging to leading including unity. The currents are, however, not very sensitive to variation in $\psi$.

6. Conclusion

In this paper the design presupposes that the control system for the inner current loop acts much faster so that for all practical purposes, it can be considered to be instantaneous to the outer speed loop. A PI controller for speed loop has been designed by choosing suitable values of $\zeta$ and $\omega_n$ as specifications to obtain the desired speed response. The output of the PI controller is the reference torque, from which the reference currents are generated based on the specified values of the torque angle ($\delta$) and the internal angle ($\psi$) of the motor. Simulation results clearly indicates that many shoots without exact feedback linearization bounding the system to be oscillatory when compared with the exact feedback linearization, though the final steady state values remain the same.
Appendix-A: Machine Ratings and Parameters

Machine Ratings and Parameters of Permanent Magnet Synchronous Motor (PMSM):

<table>
<thead>
<tr>
<th>Motor specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated voltage</td>
<td>400V</td>
</tr>
<tr>
<td>Rated current</td>
<td>2.17A</td>
</tr>
<tr>
<td>Rated speed</td>
<td>1500rpm</td>
</tr>
<tr>
<td>Number of Poles</td>
<td>04</td>
</tr>
<tr>
<td>Rated power</td>
<td>1.2/1.5kW</td>
</tr>
<tr>
<td>Power factor</td>
<td>0.8/1.0</td>
</tr>
<tr>
<td>Viscous coefficient</td>
<td>0.0048N.m/sec/rad</td>
</tr>
<tr>
<td>Moment of Inertia</td>
<td>0.048kg.m²</td>
</tr>
</tbody>
</table>

References