Design of PI Controller for Angular Velocity Control of Brushed DC Motor plus Neuro Adaptive Control

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Abstrak

Recently, the DC motor has been widely used in industry even though its maintenance costs are higher than the induction motor. Although control theory has made great advance in the last few decades, which has led to many sophisticated control schemes, PID (Proportional Integral Derivative) control still remains the most popular type of control being useed in industries today. This popularity is partly due to the fact that PID controllers have simple structures and very well understood principles. PI controller is essentially PID controller with the derivative (D) coefficient set equal to zero. A derivative coefficient is not essential and may have a detrimental effect on system response characteristics especially in a first order system.Generally, the DC servo motor systems have uncertain and nonlinear characteristics which degradeperformance of controllers.To alleviate the problems,we added a robustifying adaptive gain. Based on aLyapunov synthesis method, it was shown that the proposed adaptive gain guaranteed the convergence of the tracking error to zero and the global boundedness of all signals in the closed-loop system.

Keywords : brushed dc motor, PI controller, neural adaptive control, simulation.

1. Introduction

Due to its excellent speed control characteristics, the DC motor has been widely used in industry even though its maintenance costs are higher than the induction motor (Fisher, et. al., 1996). As a result, speed control of DC motor has attracted considerable research and several methods have evolved. Although many advanced control techniques such as robust control (Kucukdemiral, et. al., 1999), model reference adaptive control (Anwari and Kusumah, 2006), robust model reference adaptive control (Anwari, 2006), ADALINE Based Adaptive Control (He and Xu, 2008), Fuzzy Logic Control (Akbarzadeh, et. al., 1997; Bulut, et. al., 2000), neural network (Weerasoorya and Al-Sharkawi, 1991), and sliding mode control (Jovanovic and Golo, 1997) have been proposed to improve system performances, the conventional PI/PID controllers are still dominant in majority of real-world servo systems (Nandam and Sen, 1987).

PID controllers are simple in structure, easy for implementation, have low cost, inexpensive maintenance, and effectiveness. However, they may produce large overshoots and over-oscillatory responses. Combining PID control with other control techniques often results in advanced hybrid schemes that are able to improve pure PID controllers. For example, Kim et al proposed a fuzzy pre-compensated PID controller for a DC motor position servo system (Kim, et. al., 1994); Matsunaga et al presented a switching-type hybrid scheme in which the system controller switches between a fuzzy controller and a PID controller (Matsunaga, 1991); Li and Tsang proposed concurrent relay-PID controller (Li and Tsang, 2007).

PI controller is essentially an adaptation of PID controller with the derivative (D) coefficient set equal to zero. PI controller is simpler than PID controller. Although the formulae for the PI coefficients are simple, they are applicable to DC motor control system. One of the first applications of a PI controller to DC motor control system was carried out by Zein et al (Zein, et. al., 1990). They used a switchable dead-band PI controller to control DC motor. Ji and Sul proposed self tuning PI controller using digital signal processor (Ji and Sul, 1995). Iracleous and Alexandridis used fuzzy logic to tune PI controller (Iracleous and Alexandridis, 1995). Sundareswaran and Vasu proposed genetic algorithm to tune PI controller for speed control of DC motor (Sundareswaran and Vasu, 2000). Dobra proposed a robust PI synthesis, via LMI optimization (Dobra, 2002).

In this paper, we propose a combination of a classical PI controller and neural adaptive control to improve transient response. The first is to design PI controller. The second is to add a robustifying adaptive neural controller to maintain stability in the face of modeling imprecision and uncertainty.

In the simulation experiments of this paper, the proposed controller was applied on brushed DC motor control to prove its effectiveness.

2. Modelling

A theory is a general statement of principle abstracted from observation. A model is a representation of a theory that can be used for control and prediction. For a model to be useful, it must be realistic and yet simple enough to understand and manipulate. These requirements are not easily fulfilled as realistic models are seldom simple and simple models are seldom realistic.

The scope of a model is defined by what is considered relevant. Features or behaviour that is relevant must be included in the model and those that are not can be ignored. Modelling refers to the process of analysis and synthesis to arrive at a mathematical description that contains the relevant dynamic charac-teristics of the particular model (Ong, 1998).

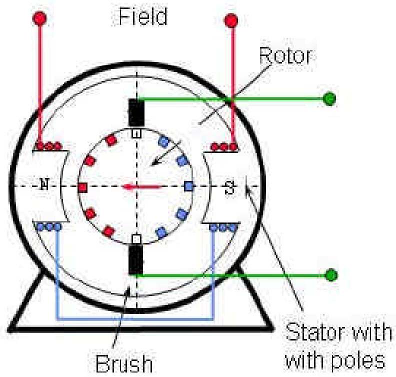

Figure 1. Brushed DC Motor Construction

The stator of the DC motor has poles, which are excited by DC current to produce magnetic fields. The rotor has a ring-shaped laminated iron-core with slots. Coils with several turns are placed in the slots. The distance between the two legs of the coil is about 180 electric degrees. The coils are connected in series. To keep the torque on a DC motor from reversing every time the coil moves through the plane perpendicular to the magnetic field, a split-ring device called a commutator is used to reverse the current at that point. The commutator shown in Figure 2 consists of insulated copper segments mounted in a cylinder. The electrical contacts to the rotating ring are called "brushes" since copper brush contacts were used in early motors. Modern motors normally use spring-loaded carbon contacts, but the historical name for the contacts has persisted. Two brushes are pressed to the commutator to permit current flow. The brushes are placed in the neutral zone (magnetic field is close to zero) to reduce arcing (Krause, 1989).

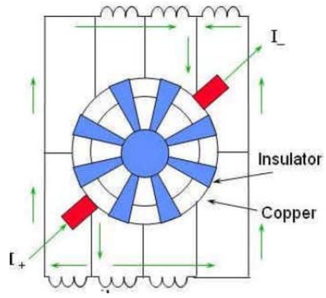

Figure 2. Concept of The Commutator

In any electric motor, operation is based on simple electromagnetism. A current carrying conductor generates a magnetic field which when placed in an external magnetic field, it will experience a force proportional to the current in the conductor and to the strength of the external magnetic field. The internal configuration of a DC motor is designed to harness the magnetic interaction between a current-carrying conductor and an external magnetic field to generate rotational motion.

The geometry of the brushes, commutator contacts, and rotor windings are such that when power is applied, the polarities of the energized winding and the stator magnet(s) are misaligned, and the rotor will rotate until it is almost aligned with the stator's field magnets. As the rotor reaches alignment, the brushes move to the next commutator contacts, and energize the next winding.

To perform the simulation of a system, an appropriate model needs to be established. For this paper, the system contains a DC motor. Therefore, a model based on the motor specifications needs to be obtained.

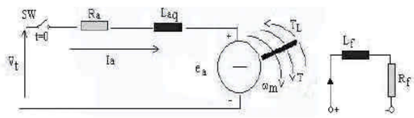

Figure 3. Schematic Diagram of a Brushed DC Motor

Applying a constant stator current and assuming magnetic linearity, the basic motor equations are

$$T_m = K_m I_a \quad (1)$$

$$e_a = K_m \omega \quad (2)$$

Let the switch SW be closed at t = 0. After the switch is closed,

$$V_t = e_a + I_a R_a + L_{aq} \frac{di_a}{dt} \quad (3)$$

From Equation (2) and (3)

$$V_t = K_m \omega + I_a R_a + L_{aq} \frac{di_a}{dt} \quad (4)$$

$$T_m = K_m I_a = J \frac{d\omega}{dt} + B\omega + T_L \quad (5)$$

The term $B\omega$ represents the rotational loss torque of the system.

The state-space representation is given by the equations:

$$\dot{\bar{x}} = A\bar{x} + Bu$$
$$y = C\bar{x} + Du \quad (6)$$

where $\bar{x} = \begin{bmatrix} \omega \\ i_a \end{bmatrix}$, $A = \begin{bmatrix} -B/J & K_m/J \\ -R_a/L_{aq} & -K_m/L_{aq} \end{bmatrix}$,

$B = \begin{bmatrix} 0 & -1/J \\ 1/L_{aq} & 0 \end{bmatrix}$, $u = \begin{bmatrix} V_t \\ T_L \end{bmatrix}$, $y = \omega$, $C = [1 \quad 0]$,

and $D = [0]$.

The transfer function between angular velocity and voltage is

$$\frac{\omega}{V_t} = \frac{A_v}{s^2 + B_1 s + B_0} \quad (7)$$

The transfer function between angular velocity and Load Torque is

$$\frac{\omega}{T_L} = \frac{A_L}{s^2 + B_1 s + B_0} \quad (8)$$

where $A_v = \frac{K_m}{J L_{aq}}$, $A_L = -\frac{K_m}{J^2}$,

$B_1 = \frac{B}{J} + \frac{K_m}{L_{aq}}$, ,and $B_0 = \frac{(B + R_a)K_m}{J L_{aq}}$.

In most DC motors, the rotor inductance and the value of B are small that can be neglected to lead to reduced order.

If is neglected then Eq. (4) becomes

$$V_t = K_m \omega + I_a R_a \quad (9)$$

If B is neglected then Eq. (5) becomes

$$T_m = K_m I_a = J \frac{d\omega}{dt} + T_L \quad (10)$$

The current-voltage relationship for the left hand side of the equation can be written and manipulated to relate between voltage and angular velocity.

$$I_a = \frac{V_t - e_a}{R_a} \quad (11)$$

$$\frac{T_m}{K_m} = \frac{V_t - K_m \omega}{R_a} \quad (12$$

$$\frac{J \frac{d\omega}{dt} + T_L}{K_m} = \frac{V_t - K_m \omega}{R_a} \quad (13)$$

$$\frac{d\omega}{dt} + \left(\frac{K_m^2}{J R_a} \right) \omega = \left(\frac{K_m}{J R_a} \right) V_t - \frac{T_L}{J} \quad (14)$$

The nominal first order linear model of a motor is shown in (15).

$$\frac{d\omega}{dt} + 2\omega = 50.3\, V_t - \frac{T_L}{19.8 \times 10^{-6}} \quad (15)$$

Since T_L is unknown, it can be included in the plant perturbation, hence transfer function of the nominal plant is

$$G_0(s) = \frac{50.3}{s + 2} \quad (16)$$

The exact model and the nominal model can be related as

$$G = G_0(1 + \Delta_m) \quad (17)$$

where G is exact model, G_0 is nominal model, and Δ_m is a multiplicative perturbation.

3. Controller Design

The PI controller can be written in this form:

$$C(s) = \frac{K_p(s + K_i/K_p)}{s} \quad (18)$$

If we set $\frac{K_i}{K_p} = 2$ and $K_p = \frac{10}{50.3}$ then the closed loop transfer function is

$$CLTF(s) = \frac{10}{s + 10} \quad (19)$$

Uncertanty in processes dynamics of the plant lead to poor control performances if controller parameters are not properly adapted. To overcome this problem we propose control input as follows

$$C(s) = PI(controller) - C_r \quad (20)$$

where C_r is a robustifying control term.

The first stage in designing an adaptive controller scheme for DC motor, the goal of the control is to drive speed of a dc motor ω to the desired speed ω_d. Let us define $e = \omega - \omega_d$ as a tracking error. We firstly define the error metric as follows:

$$s = e + \lambda \int_0^t e \, d\tau \quad (21)$$

where λ is a positive real constant. The goal is an ideal condition i.e.

$$s = 0 \quad (22)$$
$$\dot{s} = 0 \quad (23)$$

If system states remain on the ideal condition, the tracking error e will governed after such finite amount of time by the first-order differential equation $\dot{e} + \lambda e = 0$. Thus the tracking error e will converge asymptotically to 0 as $t \to \infty$ because λ is a positive constant (Slotine and Li, 1991).

Based on PI controller applied in nominal model (model without uncertainty) will imply

$$\dot{e} + K e = 0 \quad (24)$$

Based on PI controller applied in real model with uncertainty will imply

$$\dot{e} + K\,e = \Delta \qquad (25)$$

where Δ is the error between application of real model and nominal model.

Let the control input u can be chosen as Eqs. 20 will imply

$$\dot{e} + K\,e = \Delta - C_r \qquad (26)$$

The robustifying control term C_r can be constructed as a function of error metric variable s, as follows

$$C_r = W_1 s + W_2 \qquad (27)$$

where W_1 and W_2 are adaptable weights that are updated during the operation. The goal is to push s to zero in finite time.

To achieve this requirement, the following Lyapunov function is selected

$$V = \frac{1}{2} s^2 \qquad (28)$$

The function selected is positive definite and its vanishes only when $s = 0$. A global reaching condition is its time derivative be negative definite. Choosing its time derivative as

$$\dot{V} = -\gamma s^2 \qquad (29)$$

where, γ is a positive constant, restricts the derivative to be negative definite. Substituting (28) into (29), the following equation is obtained

$$\dot{s}\,s = -\gamma s^2 \qquad (30)$$

Going one step further,

$$s\,(\dot{s} + \gamma\, s) = 0 \qquad (31)$$

Hence, for the Lyapunov stability criteria to be held,

$$\dot{s} + \gamma\, s = 0 \qquad (32)$$

must be satisfied for $s \neq 0$.

A solution to this problem can be using an adaptive weight that minimizes the function $\dot{s} + \gamma\, s$ and eventually makes it zero, without need for the information about the system dynamics.

As stated above, the goal is to push the function $\dot{s} + \gamma\, s$ to zero. To achieve this goal the error function

$$E = \frac{1}{2} (\dot{s} + \gamma\, s)^2 \qquad (33)$$

is introduced to the system and weights are updated accordingly.

Weights are updated using simple gradient descent approach –in continuous form– or back propagation :

$$\dot{w}_j = -\eta \frac{\partial E}{\partial w_j} \qquad (34)$$

where, η is the learning constant, generally chosen between 0 and 1. To compute the weight updates, the derivative of the error function E w.r.t. w_j should be found. Using the chain rule, the derivative can be written as

$$\frac{\partial E}{\partial w_j} = \frac{\partial E}{\partial k} \frac{\partial k}{\partial w_j} \qquad (35)$$

Substituting (33) into (35) and taking the derivatives, the following equation is obtained:

$$\frac{\partial E}{\partial w_j} = \begin{cases} (\dot{s} + \gamma\, s) \dfrac{\partial(\dot{s} + \gamma\, s)}{\partial k_i} \dot{\sigma} \\ \quad \text{for } j = 1 \\ (\dot{s} + \gamma\, s) \dfrac{\partial(\dot{s} + \gamma\, s)}{\partial k} \\ \quad \text{for } j = 2 \end{cases} \qquad (36)$$

4. Simulation Results

Load torque was unknown but in the simulation it was assumed as follows

$$T_L = 19.8 \text{ x } 10^{-6} \text{ Nm} \qquad (37)$$

In order to validate the control strategies as described above, digital simulation was carried out on a DC motor drive system. A comparative study of the results obtained with Proportional Integral Derivative (PID) controller and PI Controller plus Neuro Adaptive Control (PINAC) was presented in this section. The control simulation results are shown in Figure 4 and Figure 5.

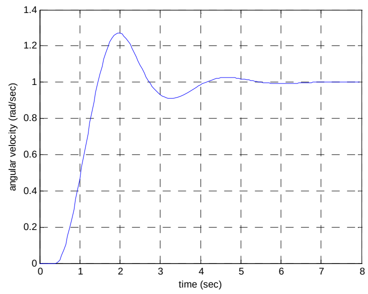

Figure 4. Simulation Result of DC Motorwith a PID Controller

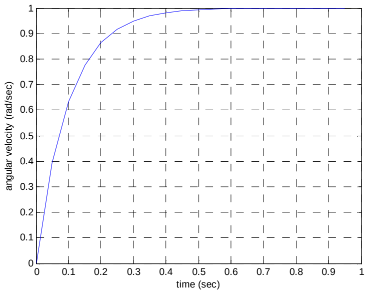

Figure 5. Simulation Result of DC Motorwith a PINAC Controller

Figure 4 shows that the response of the DC motor with Proportional Integral Derivative (PID) controller while Figure 5 depicts the speed of the DC motor using PI Controller plus Neuro Adaptive Control (PINAC) controller. Applying a PID controller to the system, the speed achieved is slightly lower than the PINAC. There is no overshoot when PINAC is applied while using PID there is a little overshoot. Results in Fig. 4 and Fig. 5 show that the PINAC performs slightly better than the PID controller.

5. Conclusions

The control of DC motor was investigated in this research work with PI Controller plus Neuro Adaptive Control (PINAC) controller. The conclusion is that PINAC was found to be superior, more robust, faster, flexible, and less sensitive to the parameter variations as compared with conventional PID controllers. Simulation results are presented to demonstrate the potential of the proposed scheme. It had been shown that the proposed scheme has several advantages such as, small steady state error, fast response, and small overshoot.

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