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Robust adaptive control of DC motor system fed by Buck converter

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Abstract

For the Buck-fed DC motor system with parameter uncertainties and external disturbances, an adaptive robust controller is proposed based on dynamic surface control and slide mode control method. The derived adaptive robust control guarantees the closed-loop errors system is uniform ultimately bounded, and the parameters update laws are designed based on the uncertainty equivalence principle. The theoretical analysis and simulation results all reveal that the developed adaptive nonlinear robust controller is not sensitive to uncertainties of system, and owns simpler configuration and more accurate tracking performance than that of conventional backstepping control or PID control method.

Keywords: *buck converter, parameter uncertainty, external disturbances, dynamic surface control, slide mode control*

1. Introduction

DC-DC Buck converter is one of the basic power electronic circuits, and it has been widely used in the fields of DC power supplies and DC motor speed regulating systems.

Generally speaking, different approaches for switching control of Buck converter have the following three control objectives. The first one is concentrated on proper control of the switching for DC-DC converter to accomplish a zero steady-state error between the output voltage and the desired setpoint. In the second one, satisfactory transient responses can be achieved when the system is subject to perturbations on load currents or line input voltages. Thirdly, much of the work in the control of DC-DC converters has been focused on the applications of linear control methods.

So far, both linear and nonlinear control strategy have been used to improve the performance of Buck converters (see [1-3], and the references therein). The disadvantage of linear control method, such as feedback linearization, is that the feedback control is valid only in the neighborhood of the operating point, so linear controller is not easy to achieve global accurate response because of the non-linearity of Buck converters. In recent years, with the application of differential geometric theories and nonlinear control technologies, DC-DC converters can be transformed into linear ones firstly, and then linear control methods can be used. But it must be noted that, some nonlinear control methods, such as differential geometric theories, need the exact dynamic models of DC-DC converters. At the same time, to the authors' best knowledge, little attention, to date, has been focused on nonlinear control of Buck converters with parametric uncertainties and exogenous disturbances.

In order to overcome the above limitations, robust and adaptive nonlinear control (RANC) has been widely studied, and there are fruitful results in a considerable amount of literature [4-6]. Backstepping approach presented in [6-8] may be the most

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successful one. The basic thought of this approach is the so-called certainty equivalence principle. In this approach, model errors and external disturbances are taken as a special kind of uncertain parameters. The common procedure of backstepping is to transform the controlled nonlinear systems into linear feedback form firstly, then find a parameter adaptive law such that a quadratic function of the states and the parameter estimation errors becomes a Lyapunov function for the closed-loop system. In other words, the parameter-dependent terms will be canceled out from the derivative of the Lyapunov function, thus the global stability and boundedness of closed-loop signals can be guaranteed. A shortcoming of this approach is that the controlled systems must meet the so-called matching condition, i.e. the control input and the unknown parameter must enter the same integrator of the system state. However, most nonlinear systems do not satisfy this condition. In order to achieve RANC of these systems, the above cancellation must be carried out for many times. As a result, over-parameterized problem and complex controller are inevitable. In order to solve this problem, dynamic surface control (DSC) method was proposed. Due to the addition of first-order low-pass filters in the DSC approach [9-11], there is no need for repetitive differentiation of the nonlinear terms of the system model. Compared to the traditional backstepping method, the explosion of differential terms can be avoided. As a benefit, the complexity of the designed controller is reduced and the design procedure is much simpler than that of traditional backstepping control.

In this paper, a new improved DSC approach is proposed for the nonlinear DC motor system fed by Buck converter which can be feedback linearized but need not satisfy matching condition. Compared with conventional backstepping or DSC methods, apart from preserving the above mentioned advantages, the parameters update laws are designed on uncertainty equivalence principle.

The rest of this paper is organized as follows. In Section 2, the problem description is given along with the estimator design and controller design. The stability analysis of closed-loop system is presented in Section 3. In Section 4, the simulation is performed on the buck converter. Finally, this paper is concluded in Section 5 with a brief discussion of the results obtained.

2. Description of System Model

The DC motor system fed by a Buck converter under consideration is depicted in Figure 1. The system model is described as follows [12].

$$\begin{cases} L\dfrac{di}{dt} = -v + Eu \\ C\dfrac{dv}{dt} = i - i_a \\ L_m\dfrac{di_a}{dt} = v - R_m i_a - k_e\omega \\ J\dfrac{d\omega}{dt} = k_m i_a - f\omega - T_L \end{cases} \tag{1}$$

where i is the converter input current, i_a is the DC motor armature current, v is the converter output voltage, ω is the motor angular velocity, T_L is the load torque, u is the control input, k_m is the torque contant, k_e is the EMF constant, J is the moment of inertia, and f is the coefficient of friction.

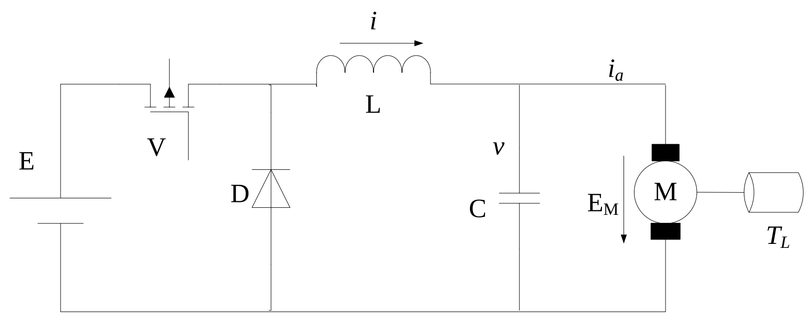

Figure 1. DC Motor System Fed by Buck Chopper

We define the state variable as $\boldsymbol{x}=\{x_1, x_2, x_3, x_4\}$, where x_1, x_2, x_3, x_4 denote the average values of the angular velocity ω, the DC motor armature current i_a, the converter output voltage v and the converter input current i, respectively. Then (1) can be rewritten as

$$\begin{cases} \dot{x}_1 = -\dfrac{f}{J}x_1 + \dfrac{k_m}{J}x_2 - \dfrac{T_L}{J} \\ \dot{x}_2 = -\dfrac{k_e}{L_m}x_1 - \dfrac{R_m}{L_m}x_2 + \dfrac{1}{L_m}x_3 \\ \dot{x}_3 = \dfrac{x_4}{C} - \dfrac{x_2}{C} \\ \dot{x}_4 = -\dfrac{x_3}{L} + \dfrac{uE}{L} \end{cases} \tag{2}$$

For (2), the some uncertainties should be considered. We set $\theta_{11} = -\dfrac{f}{J}, \theta_{12} = -\dfrac{T_L}{J}, \theta_{21} = -\dfrac{k_e}{L_m}$, $\theta_{22} = -\dfrac{R_m}{L_m}$, and supposed they are constant parameter uncertainties, and d_i (i=1,…,4) is defined as the broad-sense bounded uncertainty, then system (2) is transformed as follows.

$$\begin{cases} \dot{x}_1 = \theta_{11}x_1 + \theta_{12} + \dfrac{k_m}{J}x_2 + d_1 \\ \dot{x}_2 = \theta_{21}x_1 + \theta_{22}x_2 + \dfrac{1}{L_m}x_3 + d_2 \\ \dot{x}_3 = -\dfrac{x_2}{C} + \dfrac{x_4}{C} + d_3 \\ \dot{x}_4 = -\dfrac{x_3}{L} + \dfrac{uE}{L} + d_4 \end{cases} \tag{3}$$

For system (3), we have the following assumptions.

Assumption 1. $x_i, (i = 1, \cdots, 4)$ are both measurable, and reference signal $x_{1d} \in C^1$.

Assumption 2. $d_i \in R$ and satisfies $|d_i| \le a_i$, where a_i is a known positive constant, i=1,…,4.

3. Controller Design

For (3), we define the surface error as follows.

$$e_i = x_i - x_{id} \,, \tag{4}$$

where x_{1d} is the reference trajectory, x_{id} (i=2,…,4) will be given later on by the first order filter.

Define the boundary layer errors as

$$y_{i+1} = x_{(i+1)d} - x_{i+1}{}^{*} \,, \tag{5}$$

where $x_{i+1}{}^{*}$(i=1,…,3) is the stabilizing function which will also be designed later on.

Now we will show a new dynamic surface control procedure for the robust adaptive controller of the system defined in (3).

Step1: For the first subsystem of (3), viewing x_2 as the virtual control, we have

$$\dot{e}_1 = \theta_{11} x_1 + \theta_{12} + \frac{k_m}{J} x_2 + d_1 - \dot{x}_{1d} \tag{6}$$

Consider the first Lyapunov function V_1 candidate as follows.

$$V_1 = \frac{1}{2} e_1^2 \,. \tag{7}$$

Then the time derivative of V_1 is:

$$\begin{aligned} \dot{V}_1 &= e_1[\theta_{11} x_1 + \theta_{12} + \frac{k_m}{J} x_2 + d_1 - \dot{x}_{1d}] \\ &= e_1[\theta_{11} x_1 + \theta_{12} + \frac{k_m}{J}(e_2 + y_2 + x_2^{*}) + d_1 - \dot{x}_{1d}] \end{aligned} \,. \tag{8}$$

Select the first stabilizing function as

$$x_2^{*} = \frac{J}{k_m}\{-c_1 e_1 - \frac{e_1}{(2\gamma_1)^2} - \varphi(\mathbf{x}_1)[\hat{\boldsymbol{\theta}}_1 + \boldsymbol{\zeta}_1(\mathbf{x}_1)] + \dot{x}_{1d}\} \tag{9}$$

where c_1 and γ_1 are all larger than zero, $\boldsymbol{\theta}_1=[\theta_{11} \;\; \theta_{12}]^T$, $\boldsymbol{\varphi}(\mathbf{x}_1)=[x_1 \;\; 1]$, $\boldsymbol{\zeta}_1(\mathbf{x}_1) = \begin{bmatrix} \frac{1}{2} x_1^2 \\ x_1 \end{bmatrix}$.

The first parameter update law is taken as follows based on the uncertainty equivalence principle.

$$\dot{\hat{\boldsymbol{\theta}}}_1 = \begin{bmatrix} -x_1[\frac{k}{J} x_2 + x_1(\hat{\theta}_{11} + \frac{1}{2} x_1^2)] \\ -(\frac{k}{J} x_2 + \hat{\theta}_{12} + x_1) \end{bmatrix} \tag{10}$$

Therefore the dynamics of the estimation errors are

$$\mathbf{z}_1 = \hat{\boldsymbol{\theta}}_1 - \boldsymbol{\theta} + \boldsymbol{\zeta}_1(\mathbf{x}_1) \tag{11}$$

$$\dot{\mathbf{z}}_1 = \begin{bmatrix} -x_1^2 z_{11} \\ -z_{12} \end{bmatrix} + \begin{bmatrix} x_1 \\ 1 \end{bmatrix} d_1 \tag{12}$$

Then we have

$$\dot{V}_1 = e_1[\frac{k_m}{J}(e_2 + y_2) - c_1 e_1 - \frac{e_1}{(2\gamma_1)^2} - \varphi(\mathbf{x}_1)\mathbf{z}_1 + d_1] \cdot \tag{13}$$

Step2: Let x_2^* be an input and passed through a first-order filter as follows.

$$\tau_2 \dot{x}_{2d} + x_{2d} = x_2^* \tag{14}$$

The second Lyapunov function V_2 candidate is taken as

$$V_2 = \frac{1}{2} e_2^2 \cdot \tag{15}$$

We can get from (4)

$$\dot{e}_2 = \theta_{21} x_1 + \theta_{22} x_2 + \frac{1}{L_m} x_2 + d_2 - \dot{x}_{2d} \tag{16}$$

Then we have

$$\dot{V}_2 = e_2(\theta_{21} x_1 + \theta_{22} x_2 + \frac{1}{L_m} x_3 + d_2 - \dot{x}_{2d}) \cdot$$

Select the second stabilizing function as

$$x_3^* = L_m\{-c_2 e_2 - \frac{e_2}{(2\gamma_2)^2} - \varphi(\mathbf{x}_1, \mathbf{x}_2)[\hat{\boldsymbol{\theta}}_2 + \boldsymbol{\zeta}_2(\mathbf{x}_1, \mathbf{x}_2)] + \dot{x}_{1d}\} \tag{17}$$

where design constants $c_2>0$ and $\gamma_2>0$, $\boldsymbol{\theta}_2=[\theta_{21} \;\; \theta_{22}]^T$, $\boldsymbol{\varphi}(\mathbf{x_1}, \mathbf{x_2})=[x_1 \;\; x_2]$, $\boldsymbol{\zeta}_2(\mathbf{x}_1, \mathbf{x}_2) = \begin{bmatrix} x_1 x_2 \\ \frac{1}{2} x_2^2 \end{bmatrix}$.

The second parameter update law is taken as

$$\dot{\hat{\boldsymbol{\theta}}}_2 = \begin{bmatrix} -x_1[\frac{1}{L_m} x_3 + x_1(\hat{\theta}_{21} + x_1 x_2)] \\ -x_2[\frac{1}{L_m} x_3 + x_2(\hat{\theta}_{22} + \frac{1}{2} x_2^2)] \end{bmatrix} \tag{18}$$

The dynamics of the second estimation errors are

$$\mathbf{z}_2 = \hat{\boldsymbol{\theta}}_2 - \boldsymbol{\theta} + \boldsymbol{\zeta}_2(\mathbf{x}_1, \mathbf{x}_2) \tag{19}$$

$$\dot{\mathbf{z}}_2 = \begin{bmatrix} -x_1^2 z_{21} \\ -x_2^2 z_{22} \end{bmatrix} + \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} d_2 \tag{20}$$

We can deduce

$$\dot{V}_2 = e_2[\frac{1}{L_m}(e_3 + y_3) - c_2 e_2 - \frac{e_2}{(2\gamma_2)^2} - \varphi(\mathbf{x}_1, \mathbf{x}_2)\mathbf{z}_2 + d_2] \cdot \tag{21}$$

Step3：Let x_3^* be an input and passed through a first-order filter as follows.

$$\tau_3 \dot{x}_{3d} + x_{3d} = x_3^* \tag{22}$$

The third Lyapunov function V_2 candidate is taken as

$$V_2 = \frac{1}{2} e_3^2 \cdot \tag{23}$$

We can get

$$\dot{e}_3 = -\frac{x_2}{C} + \frac{x_4}{C} + d_3 - \dot{x}_{3d} \tag{24}$$

Select the third stabilizing function as

$$x_4^* = C[-c_3 e_3 - \frac{e_2}{(2\gamma 3)^2} + \dot{x}_{3d}], \tag{25}$$

We deduce

$$\dot{V}_3 = e_3[\frac{1}{C}(e_4 + y_4) - c_3 e_3 - \frac{e_3}{(2\gamma_3)^2} + d_3]. \tag{26}$$

Step4: Let x_4^* be an input and passed through a first-order filter as follows.

$$\tau_4 \dot{x}_{4d} + x_{4d} = x_4^* \tag{27}$$

We have

$$\dot{e}_4 = -\frac{x_3}{L} + \frac{uE}{L} + d_4 - \dot{x}_{4d} \tag{28}$$

Define the sliding mode $s=b_1e_1+b_2e_2+b_3e_3+e_4=0$ which satisfies asymptotic reached condition, where b_i are positive design constants, and define the Lyapunov function of whole system as follow.

$$V = \sum_{i=1}^{3} V_I + \frac{1}{2}\sum_{i=2}^{4} y_i^2 + \frac{1}{2}s^2 + \frac{\Gamma}{2}\sum_{I=1}^{2} \mathbf{z}_i^T \mathbf{z}_i \tag{29}$$

where $\Gamma=diag\{\lambda_1, \lambda_2\}>0$ is the gain metric.

The time derivative of V is

$$\dot{V} = \sum_{i=1}^{3} \dot{V}_i + \sum_{i=2}^{4} y_i \dot{y}_i + s\dot{s} + \Gamma\sum_{i=1}^{2} \mathbf{z}_i^T \dot{\mathbf{z}}_i \tag{30}$$

Noting that (5) and $\dot{x}_{(i+1)d} = \dfrac{x_{i+1}^* - x_{(i+1)d}}{\tau_{i+1}}$, we can get

$$\dot{y}_{i+1} = \dot{x}_{(i+1)d} - \dot{x}_{i+1}^* = \frac{x_{i+1}^* - x_{(i+1)d}}{\tau_{i+1}} - \dot{x}_{i+1}^* = \frac{-y_{i+1}}{\tau_{i+1}} - \dot{x}_{i+1}^* \tag{31}$$

Let $B_{i+1} = \dot{x}_{i+1}^*$, and assume $\sup|B_i|=D_i$, then we get

$$\begin{aligned}\dot{V} = {} & e_1[\frac{k_m}{J}(e_2 + y_2) - c_1e_1 - \frac{e_1}{(2\gamma_1)^2} - \varphi(\mathbf{x}_1)\mathbf{z}_1 + d_1] + e_2[\frac{1}{L_m}(e_3 + y_3) - c_2e_2 - \frac{e_2}{(2\gamma_2)^2} \\ & - \varphi(\mathbf{x}_1, \mathbf{x}_2)\mathbf{z}_2 + d_2] + e_3[\frac{1}{C}(e_4 + y_4) - c_3e_3 - \frac{e_3}{(2\gamma_3)^2} + d_3] + \sum_{i=2}^{4} y_i(\frac{-y_i}{\tau_i} + D_i) \\ & + s[b_1\dot{e}_1 + b_2\dot{e}_2 + b_3\dot{e}_3 - \frac{x_3}{L} + \frac{uE}{L} + d_4 - \dot{x}_{4d}] - \lambda_1 x_1^2 z_{11}^2 + \lambda_1 x_1 z_{11} d_1 - \lambda_1 z_{12}^2 + \lambda_1 z_{12} d_1 \\ & - \lambda_2 x_1^2 z_{21}^2 + \lambda_2 x_1 z_{21} d_2 - \lambda_2 x_2^2 z_{22}^2 + \lambda_2 x_2 z_{22} d_2\end{aligned} \tag{32}$$

Select the real real control input u_s as

$$u = \frac{L}{E}[-\beta_1 s - \beta_2 \operatorname{sgn}(s) - b_1\dot{e}_1 - b_2\dot{e}_2 - b_3\dot{e}_3 - \frac{s}{(2\gamma_4)^2} + \frac{x_3}{L} + \dot{x}_{4d}] \tag{33}$$

where β_1, β_2, γ_4 are positive design constants. Substituting (33) into (28) yields

$$\dot{e}_4 = -\beta_1 s - \beta_2 \operatorname{sgn}(s) - b_1\dot{e}_1 - b_2\dot{e}_2 - b_3\dot{e}_3 - \frac{s}{(2\gamma_4)^2} + d_4 \tag{34}$$

In the new coordinate defined by (4)-(34), we have an important theorem as follows.

Theorem 1. For nonlinear systems (3) in parameter feedback form with parameter uncertainty and external disturbances, the closed-loop errors system will be uniformly ultimately bounded if we apply the robust adaptive control law (33), the stabilizing function(9), (17), (25) and the parameter adaptive laws (10) and (18).

Proof. The time derivative of storage function V gives

$$\begin{aligned}\dot{V} =& -\sum_{i=1}^{3} c_i e_i^2 - \frac{e_1^2}{(2\gamma_1)^2} - \frac{e_2^2}{(2\gamma_2)^2} - \frac{e_3^2}{(2\gamma 3)^2} - \lambda_1 x_1^2 z_{11}^2 - \lambda_1 z_{12}^2 - \lambda_2 x_1^2 z_{21}^2 - \lambda_2 x_2^2 z_{22}^2 \\ & - e_1 \varphi(\mathbf{x}_1)\mathbf{z}_1 - e_2 \varphi(\mathbf{x}_1, \mathbf{x}_2)\mathbf{z}_2 + e_1 \frac{k_m}{J}(e_2 + y_2) + e_2 \frac{1}{L_m}(e_3 + y_3) + + e_3 \frac{1}{C}(e_4 + y_4) \\ & + e_1 d_1 + e_2 d_2 + e_3 d_3 + \lambda_1 x_1 z_{11} d_1 + \lambda_1 z_{12} d_1 + \lambda_2 x_1 z_{21} d_2 + \lambda_2 x_2 z_{22} d_2 - \beta_1 s^2 - \beta_2 |s| \\ & - \frac{s^2}{(2\gamma_4)^2} + s d_4 + \sum_{i=2}^{4} y_i \left(\frac{-y_i}{\tau_i} + D_i\right)\end{aligned} \tag{35}$$

It follows

$$\begin{aligned}\dot{V} \le& -\sum_{i=1}^{3} c_i e_i^2 - \beta_1 s^2 - \beta_2 |s| - \lambda_1 x_1^2 z_{11}^2 - \lambda_1 z_{12}^2 - \lambda_2 x_1^2 z_{21}^2 - \lambda_2 x_2^2 z_{22}^2 + e_1^2 + \frac{1}{2}(x_1 z_{11})^2 \\ & + \frac{1}{2} z_{12}{}^2 + \varepsilon e_2^2 + \frac{1}{2}(x_1 z_{21})^2 + \frac{1}{2}(x_2 z_{22})^2 + \lambda_1 x_1 z_{11} d_1 + \lambda_1 z_{12} d_1 + \lambda_2 x_1 z_{21} d_2 \\ & + \lambda_2 x_2 z_{22} d_2 - \frac{e_1^2}{(2\gamma_1)^2} + e_1 d_1 - \frac{e_2^2}{(2\gamma_2)^2} + e_2 d_2 - \frac{e_3^2}{(2\gamma_3)^2} + e_3 d_3 - \frac{s^2}{(2\gamma_4)^2} + s d_4 \\ & + e_1 \frac{k_m}{J}(e_2 + y_2) + e_2 \frac{1}{L_m}(e_3 + y_3) + e_3 \frac{1}{C}(e_4 + y_4) - \sum_{i=2}^{3} \frac{y_i^2}{\tau_i} + \sum_{i=2}^{3} |y_i||D_i|\end{aligned}$$

It deduce

$$\begin{aligned}\dot{V} \le& -(c_1 - 1 - \frac{k_m}{J}) e_1^2 - (c_2 - 1 - \frac{k_m}{2J} - \frac{1}{L_m}) e_2^2 - (c_3 - \frac{1}{2L_m} - \frac{1}{C}) e_3^2 - \beta_2 |s| \\ & - (\frac{\lambda_1}{2} - \frac{1}{2}) x_1^2 z_{11}^2 - (\frac{\lambda_1}{2} - \frac{1}{2}) z_{12}^2 - (\frac{\lambda_2}{2} - \frac{1}{2}) x_1^2 z_{21}^2 - (\frac{\lambda_2}{2} - \frac{1}{2}) x_2^2 z_{22}^2 \\ & - \beta_1 [(b_1 e_1 + b_2 e_2 + b_3 e_3)^2 + e_4^2 + 2(b_1 e_1 + b_2 e_2 + b_3 e_3) e_4] + \sum_{i=1}^{4} \gamma_i^2 d_i^2 \\ & + \lambda_1 d_1^2 + \lambda_2 d_2^2 + \frac{k_m}{2J} y_2^2 + \frac{1}{2L_m} y_3^2 + \frac{1}{2C} y_4^2 + \frac{1}{2C} e_4^2 - \sum_{i=2}^{4} \frac{y_i^2}{\tau_i} + \sum_{i=2}^{4} \frac{\sigma y_i^2}{2} + \sum_{i=2}^{4} \frac{D_i^2}{2\sigma}\end{aligned} \tag{36}$$

where σ>0 is a design constant.

It follows

$$\begin{aligned}\dot{V} \le & -(c_1 - 1 - \frac{k_m}{J})e_1^2 - (c_2 - 1 - \frac{k_m}{2J} - \frac{1}{L_m})e_2^2 - (c_3 - \frac{1}{2L_m} - \frac{1}{C})e_3^2 - \beta_2|s| \\ & -(\frac{\lambda_1}{2} - \frac{1}{2})x_1^2 z_{11}^2 - (\frac{\lambda_1}{2} - \frac{1}{2})z_{12}^2 - (\frac{\lambda_2}{2} - \frac{1}{2})x_1^2 z_{21}^2 - (\frac{\lambda_2}{2} - \frac{1}{2})x_2^2 z_{22}^2 \\ & + m_1 e_1^2 + m_2 e_2^2 + m_3 e_3^2 + m_4 e_4^2 + \frac{1}{2C}e_4^2 - (\frac{1}{\tau_2} - \frac{\sigma}{2} - \frac{k_m}{2J})y_2^2 - [\frac{1}{\tau_3} - \frac{\sigma}{2} - \frac{1}{2L_m}]y_3^2 \\ & -[\frac{1}{\tau_4} - \frac{\sigma}{2} - \frac{1}{2C}]y_4^2 + (\gamma_1^2 + \lambda_1)d_1^2 + (\gamma_1^2 + \lambda_2)d_2^2 + \gamma_3^2 d_3^2 + \gamma_4^2 d_4^2 + \sum_{i=2}^{4}\frac{D_i^2}{2\sigma}\end{aligned} \tag{37}$$

where $m_1 = -\beta_1(b_1^2 - 2b_1b_2 - 2b_1b_3 - 2b_1)$, $m_2 = -\beta_1(b_2^2 - 2b_1b_2 - 2b_2b_3 - 2b_2)$,

$m_3 = -\beta_1(-2b_1b_3 - 2b_2b_3 - 3b_3)$, $m_4 = -\beta_1(1 - 2b_1 - 2b_2 - 2b_3)$ 。

Suppose $|d_i| \le d_{\max}$, and select $\gamma = \max\{\sqrt{\gamma_1^2 + \lambda_1}, \sqrt{\gamma_2^2 + \lambda_2}, \gamma_3, \gamma_4\}$, we get

$$\begin{aligned}\dot{V} \le & -(c_1 - 1 - \frac{k_m}{J} - m_1)e_1^2 - (c_2 - 1 - \frac{k_m}{2J} - \frac{1}{L_m} - m_2)e_2^2 - (c_3 - \frac{1}{2L_m} - \frac{1}{C} - m_3)e_3^2 \\ & -(-m_4 - \frac{1}{2C})e_4^2 - \beta_2|s| - (\frac{\lambda_1}{2} - \frac{1}{2})x_1^2 z_{11}^2 - (\frac{\lambda_1}{2} - \frac{1}{2})z_{12}^2 - (\frac{\lambda_2}{2} - \frac{1}{2})x_1^2 z_{21}^2 \\ & -(\frac{\lambda_2}{2} - \frac{1}{2})x_2^2 z_{22}^2 - (\frac{1}{\tau_2} - \frac{\sigma}{2} - \frac{k_m}{2J})y_2^2 - [\frac{1}{\tau_3} - \frac{\sigma}{2} - \frac{1}{2L_m}]y_3^2 - [\frac{1}{\tau_4} - \frac{\sigma}{2} - \frac{1}{2C}]y_4^2 \\ & + 4\gamma^2 d_{\max}^2 + \sum_{i=2}^{4}\frac{D_i^2}{2\sigma}\end{aligned} \tag{38}$$

Select $c_1 > 1 + k_m/J + m_1$, $c_2 > 1 + k_m/(2J) + m_2 + 1/L_m$, $c_3 > 1/(2L_m) + m_3 + 1/C$, $m_4 + 1/(2C) < 0$, $\beta_2 > 0$, $\lambda_1 > 1$, $\lambda_2 > 1$, $1/\tau_2 > \sigma/2 + k_m/(2J)$, $1/\tau_3 > \sigma/2 + 1/(2L_m)$, $1/\tau_4 > \sigma/2 + /2C$, $a_0 = \min\{c_i, \beta_i, \lambda_i, \frac{1}{\tau_{i+1}}\}$,

$b_0 = 4\gamma^2 d_{\max}^2 + \sum_{i=2}^{n}\frac{D_i^2}{2ó}$, then

$$\dot{V} \le -a_0(\sum_{i=1}^{4} e_i^2 + \sum_{i=2}^{4} y_i^2 + |s| + \sum_{i=1}^{2}\varphi(\bar{\mathbf{x}}_\mathbf{i})\varphi(\bar{\mathbf{x}}_\mathbf{i})^\mathbf{T}\mathbf{z}_\mathbf{i}^\mathbf{T}\mathbf{z}_\mathbf{i}) + b_0 \tag{39}$$

If $\|\mathbf{e}\| > \sqrt{b_0/a_0}$, $\|\mathbf{Y}\| > \sqrt{b_0/a_0}$, then $\dot{V} \le 0$, where $\mathbf{e} = [e_1, e_2, e_3, e_4]]^\mathrm{T}$, $\mathbf{Y} = [y_2, y_3, y_4]^\mathrm{T}$. The system errors are uniformly ultimately bounded. □

Remark 1. When the external disturbances d_i disappear, the whole system wtill be global uniformly ultimately bounded.

Remark 2. For the selection of designed constants, it seems that the value of $G_{\max}$ needs to be decided. But in fact, this is not completely such. When control gain $g(x_1)$ is constant, we can know the real value of $G_{\max}$ definitely. When $g(x_1)$ is a bounded function, we only set c_i, $1/\tau_i$ as large enough to guarantee the stability and some performance of system by trial and errors.

4. Simulation Results

In This Section, we simulated the closed-loop system under the designed controller.

The system parameters are in the following: L=20×10^{-3}H, C=400×10^{-6}F, k_m=0.046, L_m=2.63×10^{-3}H, J=7.06×10^{-5}kg*m^2, f=8.42×10^{-4}N-m/rad, ω_d=50rad/sec, R_m=2Ω, k_e=0.05, T_L=0.05N.m, E=12V, $\theta_{11} = -11.92$, $\theta_{12} = -708.21$, $\theta_{21} = -19.01$, $\theta_{22} = -760.45$.

Set d_1=0.01sinx_1, d_2=0.05cosx_2, d_3=0.2sinx_3, d_4=0.1cosx_4. The relevant design parameters are taken as follows. c_1=5000, c_2=5000, c_3=8000, γ=0.1, β_1=20000, β_2=10, b_1=0.2, b_2=0.05, b_3=0.05, m_4=–8000, m_3=3450, m_2=2200, m_1=1600, λ_1=λ_2=1.05, σ=100, τ_2=0.002=τ_3, τ_4=0.0005, $x(0)=\hat{\theta}_i(0)=0$.

When the reference signals is set as x_{1d}=[120 120 160 160 120 120] at time=[0 0.5 1.0 1.5 2.0 2.5], the simulation results are shown in Figs. 2-3.

As Fig. 2 shows, although the bounded disturbances have been imposed on the system states, but under the designed controller, the motor speed can track the reference signal, and reach the steady state in a very short period of time, it has the very good convergence and robustness.

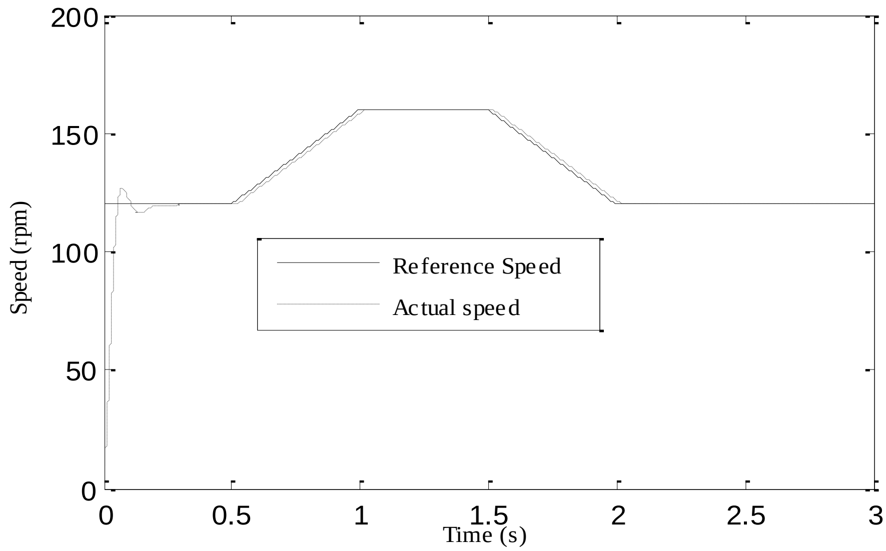

Figure 2 The Actual Speed Responding and Reference Speed Curves

We also compare the control performance of the proposed controller with that of PID control. From Fig. 3, the responding speed and adjusting time of the motor speed under the proposed controller is superior to that of PID control. The proposed controller has the advantages of high accuracy, fast response and little overshoot.

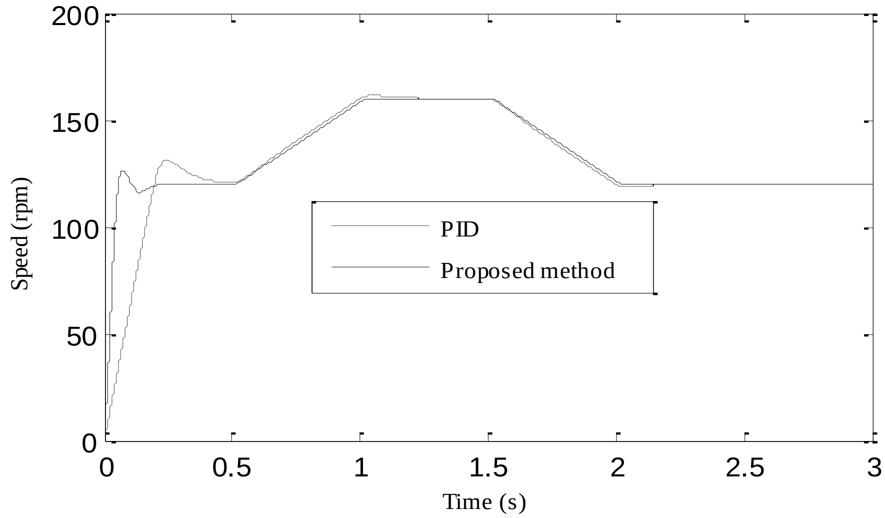

Figure 3. The Speed Responding Curves under the Proposed Controller and under the PID Control

5. Conclusions

Based on the nonlinear model of DC motor system, a novel adaptive robust DC motor controller is designed by combining the sliding mode control with the dynamic surface control design. The uncertainties of system parameters and disturbances are taken into consideration, and the parameters updated law is based on the uncertainty principle of equivalence. At the same time, dynamic surface control avoids the explosion of differential terms by using a low pass filter, thus the design method and process are further simplified. The simulation results verified the effectiveness of the proposed control method.

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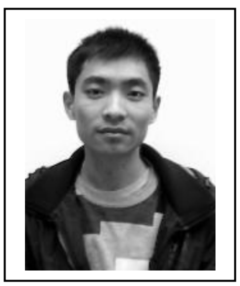

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