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Application of a PID Controller using MRAC Techniques for Control of the DC Electromotor Drive

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Abstract **– In DC motor drives, device aging and environmental factors can degrade the performance of the control system and make it difficult to redesign controller parameters in real time. In this paper, an algorithm that combines PID control scheme and Model Reference Adaptive Control (MRAC) to autotune the controller parameters in real time when system performances change is proposed and applied to control of the DC electromotor drive. The algorithm modifies the traditional PID control scheme by replacing the error between the reference and the system output with the direct system output in the derivative part. Compared to the traditional MRAC approach, the algorithm is normalized so that a reference model can be easily generated given bandwidth requirement. To be practical for implementation, the algorithm also simplifies the transfer function of the speed open-loop as an integrator by adding a current inner loop. The results of simulation on a two-loop servo control system show that the adjusted controller parameters meet the system design very well. The algorithm was implemented on a test platform and the performance of the system was evaluated with a dynamic signal analyzer. The effectiveness of the proposed method is shown through the experimental results.**

Index Terms – Adaptive control, PID control, Parameter estimation, Normalization, DC Electromotor

I. INTRODUCTION

With advances in digital technology, the science of automatic control now offers a wide spectrum of choices for control schemes, among which the proportional-integral-derivative (PID) controllers with its three-term functionality covering treatment to both transient and steady-state responses are mostly used in a wide range of applications including electromotor drives, magnetic and optic memories, automotive, flight control, instrumentation, etc. In industrial applications, more than 90% of control loops are PID type, particularly at lowest levels, as no other controllers match the simplicity, clear functionality, applicability, and ease of use offered by the PID controller. Its wide application has stimulated and sustained the development of various PID tuning techniques, sophisticated software packages, and hardware modules [1], [2].

Over the past 50 years, several methods for determining PID controller parameters have been developed for stable processes that are suitable for auto-tuning and adaptive control, such as Coon-Cohen reaction curve method, Ziegler-Nichols frequency-response method, and so on. However, these tuning methods use only a small amount of information about the dynamic behavior of the system, and often do not provide good tuning. Performances of controllers play important roles in process industry, or electromechanical systems. However, the time varying and nonlinear properties inherent in system frequently result in the widely used PID controller performance degradation. This means that the fixed gain PID controller is not robust enough and therefore further improvement is necessary because of the high demands for accurate and robust systems [3], [4]. It has been shown that the adaptive controller can significantly improve system behavior [5], [6], [7], [8].

Although the adaptive controllers improve responses of the nonlinear systems and systems with variable parameters, they are not yet used very often. The obvious reason is their complexity [9], [10], [11]. The signal adaptation ensures fast adaptation process, but the signal in such case contains high frequency oscillations, so that the signal adaptation system is more likely to be realized in analogous technique rather than in digital technique [12]. In parameter model reference adaptation, the problem of high frequency does not exist. Therefore, the model reference adaptive control with parameter adaptation is more convenient for microprocessor implementation. The idea of the model reference adaptive control (MRAC) is based on forcing the plant to follow the reference model, i.e. the adaptive controller has to decrease the error vector between the reference model and plant to zero [13], [14], [15].

In this paper, a technique of designing a PID controller using MRAC techniques with parameter adaptation is proposed. As in the DC electromotor control system, system parameters variation which is the result of device aging and environmental factors, make the system performance degrade. In this condition, the adaptation loop of such an application is simply switched for a period of time and perturbation signals are normally added to improve parameter estimation. The adaptive controller runs until the system performance is satisfied. Then the adaptation loop is disconnected, and the system is left running with fixed controller parameters.

The rest of the paper is organized as follows. In Section II, an overview of the PID control scheme and the MRAC approach is given. In Section III, we present an adaptation law

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which combines the MRAC concepts with the MIT rule and the PID controller, and the normalization of the algorithm is also given. Section IV introduces the simplified speed open-loop of a nominal two-loop servo control system after adding a current inner loop. In addition, the simulation of the proposed algorithm on the speed loop is provided. Also, results of the presented algorithm implemented on the experimental platform are given. Finally, Section V concludes with further discussion on the results and the algorithm presented.

II. PID CONTROL AND MRAC

A. PID Control

The PID algorithm remains the most popular approach for industrial process control, despite continual advances in control theory. This is not only because the PID algorithm has a simple structure which is conceptually easy to understand and implement in practice but also because the algorithm provides adequate performance in the vast majority of applications. A PID controller may be considered as an extreme form of a phase lead-lag compensator with one pole at the origin and the other at infinity. Similarly, its cousin, the PI and the PD controllers, can also be regarded as extreme forms of phase-lag and phase-lead compensators, respectively. A standard PID controller is also known as the "three-term" controller, whose transfer function is generally written in the "parallel form" given (1) by or the "ideal form" given by (2)

$$G_{PID}(s) = U(s)/E(s) = K_p + K_i/s + K_d s \tag{1}$$

$$= K_p(1 + 1/T_i s + T_d s) \tag{2}$$

where $U(s)$ is the control signal acting on the error signal $E(s)$, K_P is the proportional gain, K_i the integral gain, K_d the derivative gain, T_i the integral time constant, T_d the derivative time constant. In the time domain, (1) can be transformed to:

$$u(t) = K_p e(t) + K_i \int_0^t e(\tau)d\tau + K_d de(t)/dt \tag{3}$$

The "three-term" functionalities are highlighted by the following.

- The proportional term—providing an overall control action proportional to the error signal through the all-pass gain factor.
- The integral term—reducing steady-state errors through low-frequency compensation by an integrator.
- The derivative term—improving transient response through high-frequency compensation by a differentiator.

The message that increasing the derivative gain K_d, will lead to improved stability is commonly conveyed from academia to industry. However, practitioners have often found that the derivative term can behave against such anticipation particularly when there exists a transport delay. Frustration in tuning K_d has hence made many practitioners switch off or even exclude the derivative term. For optimum performance, K_p, K_i (or T_i) and K_d (or T_d) must be tune jointly.

In cascade control applications, the inner-loop often needs to be less sensitive to set-point changes than the outer-loop. For the inner-loop, a variant to the standard PID structure may be adopted, which uses the process variable instead of the error signal, for the derivative term, i.e.

$$u(t) = K_p e(t) + K_i \int_0^t e(\tau)d\tau - K_d dy(t)/dt \tag{4}$$

Where $y(t)$ is the process variable, $e(t)$ is the output error.

For set-point tracking applications, however, one alternative to using the type above is perhaps a set-point filter that has a critically-damped dynamics so as to achieve soft-start and smooth control [1].

B. MRAC

The model reference adaptive control (MRAC) was originally proposed to solve a problem in which the specifications are given in terms of a reference model that tells how the process output should responds to the command signal. The MRAC which was proposed by Whitaker in 1958 is an important adaptive controller [5], [13]. A block diagram of MRAC is illustrated in Fig. 1.

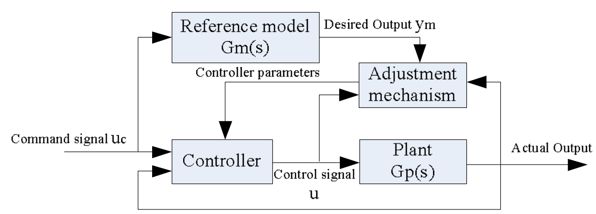


Fig. 1 Block diagram of a model-reference adaptive control (MRAC) system.

The controller presented above can be thought of as consisting of two loops. The ordinary feedback loop which is called the inner loop composed of the process and controller. The parameters of the controller are adjusted by the adaptation loop on the basis of feedback from the difference between the process output y and the model output y_m. The adaptation loop which is also called the outer loop adjusted the parameter in such a way that makes the difference becomes small.

An important problem associated with the MRAC system is to determine the adjustment mechanism so that a stable system that brings the error to zero is obtained. The following parameter adjustment mechanism, called the MIT rule, was originally used in MRAC:

$$d\theta/dt = -\gamma e(\partial e/\partial\theta) \tag{5}$$

In (5), e $(e = y - y_m)$ denotes the model error and θ is the controller parameter vector. The components of $\partial e/\partial\theta$ are the sensitivity derivatives of the error with respect to θ. The parameter γ is known as the adaptation gain. The MIT rule is a gradient scheme that aims to minimize the squared model error e^2 [3].

III. DESIGN OF A PID CONTROLLER USING MRAC TECHNIQUES

A. A PID Controller using MRAC Techniques

Consider a process with a first order transfer function

$$G(s) = b/(s+a) \tag{6}$$

Where a is assumed to be positive and b nonnegative.

In the time domain, using the differential operator $p = d/dt$, (4) is rewritten as

$$u(t) = K_p(u_c(t) - y(t)) + K_i(u_c(t) - y(t))/p - K_d py(t) \tag{7}$$

and (6) is rewritten as

$$G(p) = y(t)/u(t) = b/(p+a) \tag{8}$$

It follows from (7) and (8) that

$$\begin{aligned} y(t) &= (bK_p p/(1+bK_d) + bK_i/(1+bK_d))u_c(t) \\ &/(p^2 + (a+bK_p)p/(1+bK_d) + bK_i/(1+bK_d)) \end{aligned} \tag{9}$$

In (9), the numerator and the denominator have the same constant $bK_i/(1+bK_d)$, so the system will has unit stability gain and a real zero $-K_i/K_p$ after design.

According to the closed form transfer function in (9), we select a second order oscillating system as the reference model:

$$G_m(s) = Y_m(s)/U_m(s) = (\alpha s + \omega_n^2)/(s^2 + 2\zeta\omega_n s + \omega_n^2) \tag{10}$$

which has unit steady state gain, natural frequency ω_n, and damping ratio ζ. Notice the presence of a real zero at $-\omega_n^2/\alpha$ in the reference model. This zero is introduced to match the structure of (9).

Recall that the model error is defined as the difference between the process output y and the reference model output y_m $(e = y - y_m)$. It is then possible to derive adaptation rules for the controller parameter vector $\theta = \{K_p, K_i, K_d\}$ of control law (7) using MIT rule $d\theta/dt = -\gamma' e(\partial e/\partial\theta)$ with θ as follows:

$$\partial e/\partial\theta = \partial(y - y_m)/\partial\theta = \partial y/\partial\theta \tag{11}$$

$$d\theta/dt = p\cdot\theta = -\gamma' e\partial e/\partial\theta \tag{12}$$

It follows from (11) and (12) that

$$\theta = (-\gamma'/p)\cdot e\cdot\partial e/\partial\theta \tag{13}$$

By replacing θ with K_p, K_i and K_d respectively in (13), we get following equations:

$$K_p = (-\gamma_p'/p)\cdot b/(1+bK_d)\cdot e\cdot p\cdot(u_c - y)/M(p) \tag{14}$$

$$K_i = (-\gamma_i'/p)\cdot b/(1+bK_d)\cdot e\cdot(u_c - y)/M(p) \tag{15}$$

$$K_d = (-\gamma_d'/p)\cdot b/(1+bK_d)\cdot e\cdot p^2\cdot y/M(p) \tag{16}$$

where $M(p) = p^2 + (a+bK_p)\cdot p/(1+bK_d) + bK_i/(1+bK_d)$

Given that the process parameters a and b are not known, the exact formulas that derived using the MIT rule (5) can not be used. Instead, some approximations are required. If we think the system is perfect model-following [5], comparing the input-output relations of the system with the reference model, the following approximations will be used:

$$\begin{cases} 2\zeta\omega \approx (a+bK_p)/(1+bK_d) \\ \omega_n^2 \approx bK_i/(1+bK_d) \end{cases} \tag{17}$$

At the same time, the original adaptation gain γ' is replaced by $\gamma_j'' = b\gamma_j'/(1+bK_d), (j = k,i,d)$ and by taking (17) into (14) (15) (16) we get following equations:

$$K_p = (-\gamma_p''/p)\cdot e\cdot p(u_c - y)/(p^2 + 2\zeta\omega_n p + \omega_n^2) \tag{18}$$

$$Ki = (-\gamma_i''/p)\cdot e\cdot(u_c - y)/(p^2 + 2\zeta\omega_n p + \omega_n^2) \tag{19}$$

$$K_d = (-\gamma_d''/p)\cdot e\cdot p^2\cdot y/(p^2 + 2\zeta\omega_n p + \omega_n^2) \tag{20}$$

B. Normalization

To relate the adaptation parameters vector θ with the system design demand, θ should be normalized to meet different system bandwidth desire. Assume that we want to obtain a closed-loop system described by

$$G_{mb}(s) = (\alpha(s/\omega_b) + \omega_n^2)/((s/\omega_b)^2 + 2\zeta\omega_n(s/\omega_b) + \omega_n^2) \tag{21}$$

where ω_b is the expected system bandwidth frequency. If $\omega_n = 1(rad/s)$ is chosen to be the unit frequency, (21) will be transformed into as follows:

$$G_{mb}(s) = (\alpha\omega_b s + \omega_b^2)/(s^2 + 2\zeta\omega_b s + \omega_b^2) \tag{22}$$

In the same way, (18) (19) (20) can be rewritten as follows:

$$K_p = (-\omega_b^2\gamma_p''/p)\cdot e\cdot p\cdot(u_c - y)/(p^2 + 2\zeta\omega_b p + \omega_b^2) \tag{23}$$

$$Ki = (-\omega_b^3\gamma_i''/p)\cdot e\cdot(u_c - y)/(p^2 + 2\zeta\omega_b p + \omega_b^2) \tag{24}$$

$$K_d = (-\omega_b\gamma_d''/p)\cdot e\cdot p^2\cdot y/(p^2 + 2\zeta\omega_b p + \omega_b^2) \tag{25}$$

Replacing $\omega_b\gamma_p'', \omega_b\gamma_i'', \omega_b\gamma_d''$ with $\gamma_p, \gamma_i, \gamma_d$ respectively, it follows from (22) (23) (24) that

$$K_p = (-\gamma_p/p)\cdot e\cdot p\cdot(u_c - y)\cdot\omega_b/(p^2 + 2\zeta\omega_b p + \omega_b^2) \tag{26}$$

$$Ki = (-\gamma_i/p)\cdot e\cdot(u_c - y)\cdot\omega_b^2/(p^2 + 2\zeta\omega_b p + \omega_b^2) \tag{27}$$

$$K_d = (-\gamma_d/p)\cdot e\cdot p^2\cdot y/(p^2 + 2\zeta\omega_b p + \omega_b^2) \tag{28}$$

Normalized MRAC algorithm according to (26) (27) (28) can control the plant successfully even though the reference signal is different from a signal used during optimization and with presence of disturbance in plant as well. The normalization ensures algorithm insensitivity to variations in signal and the disturbance level, and makes the system design more convenient. The diagrammatic representation of the three equations above is depicted in Fig.2.

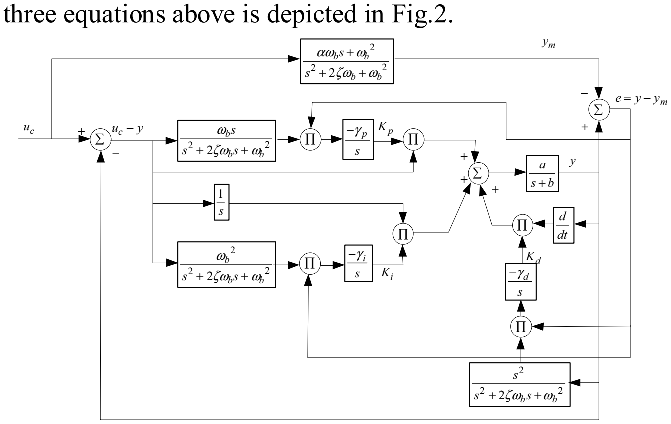


Fig. 2 Block diagram of a PID controller using the MIT rule.

IV. Application

PID controller parameters autotuning using MRAC techniques can be carried out in two steps. At first, the adaptation is started when there exists system performance changes and the process is excited adequately in order to enable the parameter adaptation to take place. Then, when the parameters have adapted after a period time, the adaptation is stopped and the PID controller operates again with fixed parameters.

A. A Two-Loop Servo Control System

A nominal two-loop servo control system is shown in Fig. 3. It consists of a current inner loop and a speed loop with two controllers.

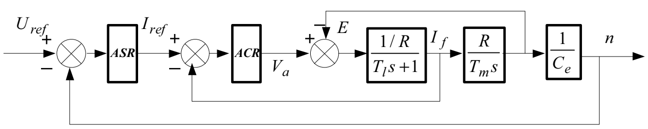


Fig. 3 Block diagram of a two-loop servo control system.

where T_l the electromagnet time constant of the armature loop, T_m the electromechanical time constant, R the armature resistance, C_e the constant of counter electromotive force, ACR the automatic current regulator, ASR the automatic speed regulator, E the counter electromotive force, V_a the control voltage, I_f the armature current, I_{ref} the reference current, U_{ref} the reference voltage, and n the electromotor speed.

Firstly, we consider the current open-loop, as shown in Fig. 4. The transfer function relating the output E to the input V_a is

$$G_c(s) = E(s)/V_a(s) = (T_m/R)s/(T_m T_l s^2 + T_m s + 1) \tag{29}$$

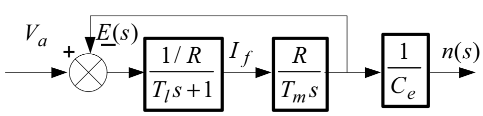


Fig. 4 Block diagram of the current open-loop.

In most servo control applications, T_l and T_m are constants if the electromotor is selected and the two transition frequencies of the system are also constants. This can be shown by changing (29) to

$$G_c(s) = (R/T_l)s/(s-T_1)(s-T_2) \tag{30}$$

where $T_{1,2} = (-1 \pm \sqrt{1-4T_l/T_m})/2T_l$ and $T_{1,2} < 0$ ($T_l/T_m < 0.1$ is normally the case in the real applications).

It is shown from (30) that the two poles of $G_c(s)$ are on the left plane of the complex plane and the system is stable. However, the system bandwidth seldom can meet the design need. To expand the system bandwidth of the current open-loop, a PI controller with the transfer function

$$G_{acr}(s) = K(T_e s + 1)/T_e s \tag{34}$$

is designed as shown in Fig. 5.

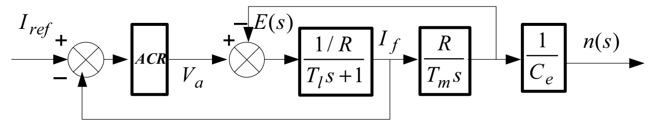


Fig. 5 Block diagram of the current closed-loop using a PI controller.

Therefore, the transfer function relating I_f to I_{ref} is

$$\begin{aligned} &I_f(s)/I_{ref}(s) = (KT_m T_e s + KT_m) \\ &/(RT_m T_l T_e s^2 + (T_e T_m R + T_e T_m R)s + (RT_e + KT_m)) \end{aligned} \tag{32}$$

If the PI controller and electromotor parameters are selected to satisfy $KT_m >> RT_e$ and $T_e < T_l$, (32) is simplified as following:

$$I_f(s)/I_{ref}(s) \approx (KT_e s + K)/(RT_l T_e s^2 + (T_e R + T_e R)s + KT_m) \tag{33}$$

It is shown from (33) that the system is not influenced by the electromechanical load and the performance (mainly the system bandwidth) of the system will not change if the control system doesn't saturate. Now, the transfer function relating the output $n(s)$ and the input $I_{ref}(s)$ is

$$\begin{aligned} &n(s)/I_{ref}(s) = (KRT_e s + KR) \\ &/(T_m T_l T_e C_e R s^3 + C_e T_e T_m (R+K)s^2 + C_e(RT_e + KT_m)s) \end{aligned} \tag{34}$$

To simplify (34), we use the following assumptions

$$\begin{cases} T_m T_l T_e C_e R << 1 \\ K >> R \\ KT_m >> RT_e \end{cases} \tag{35}$$

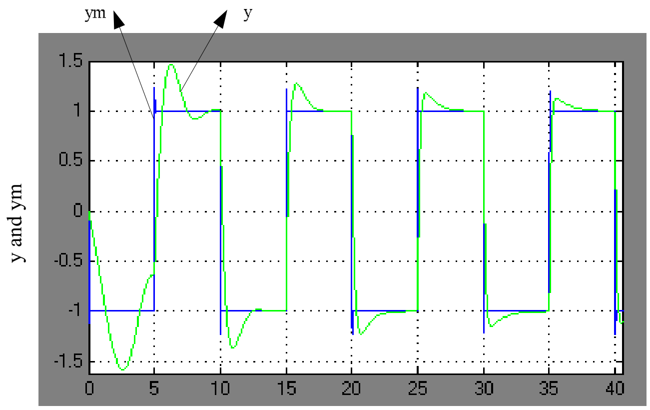
and get the equation

$$n(s)/I_{ref}(s) \approx R(T_e s + 1)/C_e T_m (T_e s + 1)s \approx (R/C_e T_m)(1/s) \tag{36}$$

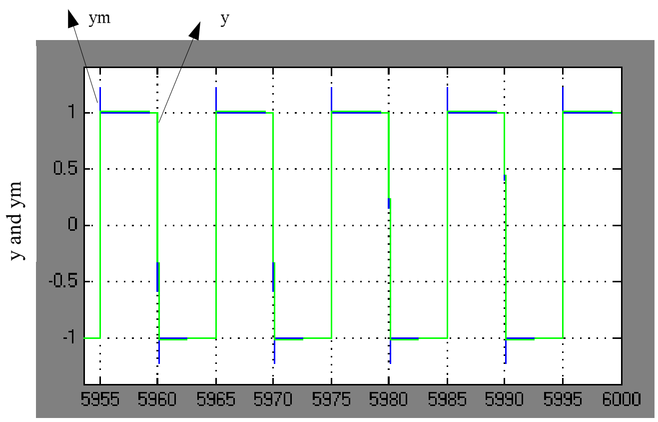
which shows the transfer function of the speed open-loop is nearly an integrator. Comparing (36) with (6), it is shown that the simplified speed open-loop satisfies the demand of the algorithm proposed.

B. Simulation

To illustrate, a simulation has been carried out with the following parameter values: $b=0$, $a=1$, $\zeta=0.7$, $\alpha=1$ $\omega_b = 100(rad/s)$, $\gamma_p = 100$, $\gamma_i = 2.5$ and $\gamma_d = 0.1$. The external signal was chosen to be a square wave with amplitude 1 and frequency 0.1 Hz, and the simulating time is 6000(s). Fig. 6 shows the reference model and process output during and after the adaptive tuning with the switching time 5000(s). Notice the behavior of the system during and after the tuning period although the oscillation process is omitted. Notice that the system follows closely the reference model output after the adaptation is carried out.



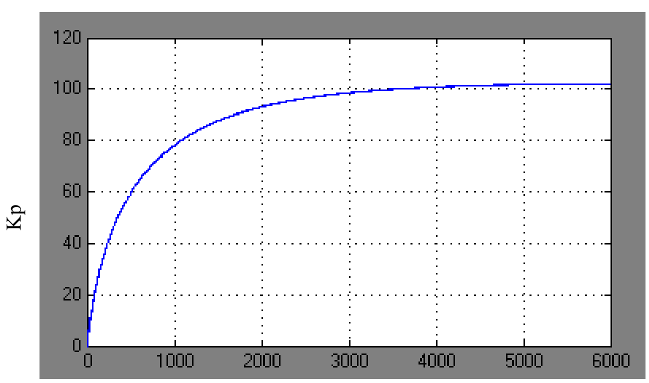
Time(s)
(a) During adjustment



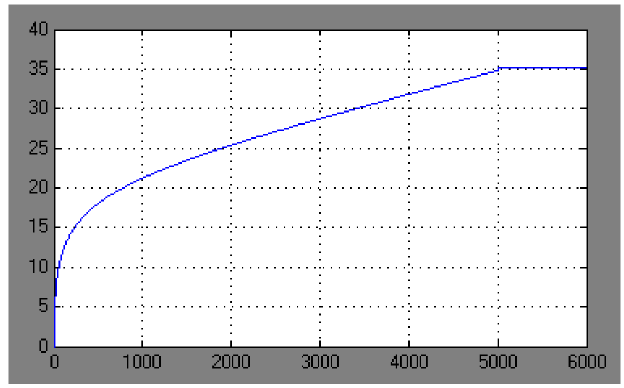
Time(s)
(b) After adjustment

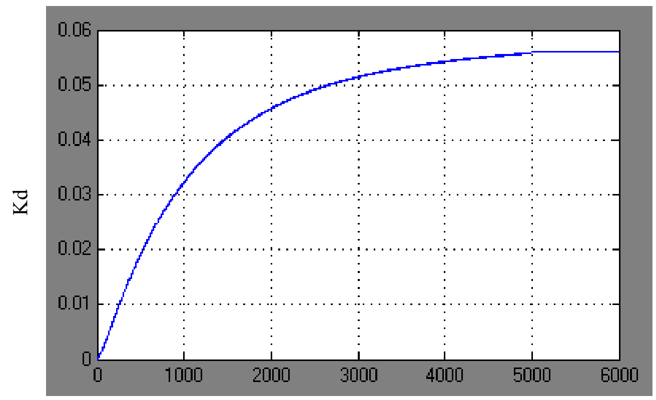
Fig. 6 Process and model response during and after adjustment. (a)during adjustment and (b)after adjustment

Figures 7 to 9 show the behavior of the PID controller parameters during and after adaptive tuning. Notice that the results are good despite of the adaptation not having converged yet when the adaptation mode was switched off at 5000(s).



Time(s)
Fig. 7 Variation of parameter K_p .


Time(s)
Fig. 8 Variation of parameter K_i .



Time(s)
Fig. 9 Variation of parameter K_d .

C. Experiment

The test platform which is used to evaluate the algorithm proposed is specified in Fig. 10. PowerPC based embedded operating system which is controlled by the PC through the Ethernet communicates with Texas Instrument Corporation DSP TMS320F2812 using CPCI (Compacted PCI) bus. The DC electromotor J320LYX08E (R=2.1Ω , $T_l = 0.0052s$, $C_e = 15V/(rad/s)$, and $T_m = 0.2s$) is driven at constant frequency by the DSP TMS320F2812 through IGBT (Insulated Gate Bipolar Transistor) module. The dynamical signal analyzer SR785 of Stanford Research System Inc. which is connected to the DSP TMS320F2812 by a special connector is used to evaluate the time and frequency domain performance of the control system.

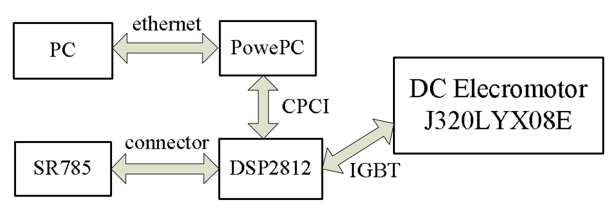


Fig. 10 Block diagram of the infrastructure of the test platform.

Ki

Using the PI controller to the current loop and selecting the suitable parameters which satisfy the assumption in (35), the speed open-loop of the control system is adjusted as an integrator which is really the case in (36). Also, the system bandwidth of the current loop is expanded to about 1000 rad/s to satisfy the system design need. The only difference is that the numerator of the transfer function is $c(c \in (0.69, 0.7))$ which is obtained by SR785.

Now, through discretization the block scheme of the MRAC algorithm with parameter adaptation according to Fig.2 is transformed from the Matlab/Simulink model to the C code of DSP TMS320F2812. The exciting square wave is generated by the PowerPC with amplitude 1 and frequency 0.1Hz which can be configured by the PC. The sampling frequencies of the speed-loop and current-loop are $400Hz$ and $4kHz$ respectively. Adaptation gains : $\gamma_p = 200$, $\gamma_i = 0.5$ and $\gamma_d = 0.1$ are selected empirically and have good practical effects. The reference model is set by $\alpha = 1$, $\zeta = 0.75$ and $\omega_b = 150(rad/s)$. The control system is autotuned adaptively for a fixed time five times and PID controller parameters of the speed loop are obtained. Performances of the control system with fixed PID controller parameters are evaluated through SR785 both in time domain and frequency domain. The results are shown in Table I.

TABLE I
TIME AND FREQUENCY DOMAIN PERFORMANCE O F THE ADAPTED SYSTEM

No.	K_p	K_i	K_d	$\omega_b(rad/s)$	$t_r(s)$	$\sigma\%$	$t_s(s)$
1	216	7.62	0.055	149	0.016	0	0.045
2	212	7.59	0.054	144	0.015	0.1	0.046
3	212	7.63	0.05	151	0.014	0	0.042
4	215	7.55	0.059	146	0.012	0.05	0.043
5	210	7.7	0.057	152	0.016	0.02	0.048

It is shown from Table I that the bandwidth of the servo system with the adapted PID controller fluctuates around the desire. Notice that there is nearly no overshoot and the reaching time is so small that meets the real system demand. Although with some disturbances to the control system of the test platform, the robustness of the PID controller with adapted controller parameters is really what we expected.

V. CONCLUSION

A PID controller for control of the DC electromotor drive using MRAC concepts and the MIT rule has been investigated. Simple adaptation laws for the controller parameters have been presented assuming that the process under control can be approximated by a first order transfer function. The developed adaptation rules with normalization have been applied to a first order system with a second order reference model. Furthermore, the proposed technique has been applied to a two-loop servo control system which has non-linear dynamics. The results obtained show the effectiveness and robustness of the technique. The performance could be improved by a better choice of adaptation gains.

The MRAC with the MIT rule is known to have some disadvantages. Firstly, the stability of the nominal system can not be decided by the MIT rule [5]. The Lyapunov approach can be used to guarantee nominal stability. Secondly, the speed of adaptation is influenced by the values and frequency of the command signal. Usually a normalized adaptation rule can be used to solve this problem.

A further limitation of the approach is the decision of the adaptation gains. In this paper empirical gains are used for the adaptation of the controller parameters, which could be suboptimal in some applications. In general, choosing an increased adaptation gain make the system follow faster, but the behavior of a system with exceedingly high gains is relatively unpredictable. Optimally choosing the adaptation gain is our future work.

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