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Variable Structure Adaptive Pole Placement Control

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Abstract— **In this paper, a variable structure adaptive pole placement control (VS-APPC) for plants of arbitrary order is proposed. Due to its flexibility in choosing the controller design methodology (state feedback, compensator design, linear quadratic, etc.) and the adaptive law (least squares, gradient, etc.), the APPC is the most general type of adaptive control. Traditionally, it has been developed in an indirect approach and, as an advantage, it may be applied to nonminimum phase plants. The combination of this strategy with the variable structure systems allows to aggregate fast transient and robustness to parametric uncertainties and disturbances. Therefore, new switching laws are proposed, instead of using the traditional integral adaptive laws. Additionally, preliminary simulation results for a nonminimum phase unstable system are shown.**

I. Introduction

A class of control schemes that are popular in the known parameter case are those that change the poles and do not involve plant zero-pole cancellations. These schemes are referred as pole placement schemes and are applicable to both minimum and nonminimum phase linear time invariant (LTI) plants. The combination of a pole placement control law with a parameter estimator or an adaptive law leads to an adaptive pole placement control (APPC) scheme that can be used to control a wide class of LTI plants with unknown parameters. Such technique was developed based on the indirect adaptive control schemes, where the control signal is a function of the plant parameters estimates.

On the other hand, the variable structure control (VSC) approach has its roots in relay control, and consists of using a switching control law as a function of system state variables, and, in its common configuration, in order to restrict the system dynamics to a surface referred as a sliding surface. The variable structure systems have as main characteristics the fast transient and robustness to parameter uncertainties and disturbances (in a range stipulated on project), although measurements of all state variables be necessary, what may be undesirable or even not possible in some cases [1].

Thereby, a control technique that inherit the VSC qualities was developed, but with only input/output measurements, which was named VS-MRAC (Variable Structure Model Reference Adaptive Control) [2,3], where the MRAC (Model Reference Adaptive Control) integral adaptation laws [4] were replaced by switching laws. This algorithm was based on the direct approach of MRAC, where the desired I/O properties of the closed-loop system are given by a reference model. This approach is restricted to minimum phase plants. In order to simplify the controller design, a new controller was proposed, named indirect VS-MRAC [5], which makes use of the plant nominal parameters for the relays amplitude calculation, since they are related with physical parameters, such as resistances, capacitances, inertia moments, etc. The VS-MRAC controller, in its direct and indirect approaches, have been successfully applied on control of DC machines [6] as well as on control of three-phase induction motors [5,7,8].

In a recent work was presented a controller that aggregates the characteristics of both techniques, namely, APPC and VSC [9], to first order plants. Thus, it's expected applicability to nonminimum phase plants, fast transient and robustness to parameter uncertainties and disturbances. This controller was named VS-APPC, where, likewise VS-MRAC, the integral adaptive laws were replaced by switching laws. In this paper, the switching laws for plants of arbitrary order are shown.

II. Pole Placement Control (PPC)

Considering the single input/single output (SISO) LTI plant

$$y = G(s)u,\ G(s) = \frac{Z(s)}{R(s)} = \frac{b_{n-1}s^{n-1} + \ldots + b_1 s + b_0}{s^n + a_{n-1}s^{n-1} + \ldots + a_1 s + a_0} \qquad (1)$$

there are, as plant parameters, $2n$ elements, which are the coefficients of the numerator and denominator of $G(s)$.

S1. *R(s)* is a monic polynomial whose degree n is known.

S2. *Z(s)* and *R(s)* are coprime and degree(Z) < n.

Assumptions (S1) and (S2) allow Z and R to be non-Hurwitz in contrast to the MRC (Model Reference Control) case where Z is required to be Hurwitz.

We can also extend the PPC objective to include tracking, where y is required to follow a certain class of reference signals r, by using the internal model principle [10]. The uniformly bounded reference signal is assumed to satisfy

$$Q_m(s)r = 0 \qquad (2)$$

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where $Q_m(s)$, the internal model of r, is a known monic polynomial of degree q with non-repeated roots on the $j\omega$-axis and satisfies

S3. $Q_m(s)$ and $Z(s)$ are coprime.

We consider the control law

$$Q_m(s)L'(s)u = -P(s)y + M(s)r \tag{3}$$

where $P(s)$, $M(s)$ and $L'(s)$ are polynomials (with $L'(s)$ monic) of degree $q + n - 1$, $q + n - 1$ and $n - 1$, respectively, to be found and $Q_m(s)$ satisfies (2) and assumption (S3).

Applying (3) to the plant (1), we obtain the closed-loop plant equation

$$y = \frac{Z(s)M(s)}{Q_m(s)L'(s)R(s) + P(s)Z(s)} r \tag{4}$$

whose characteristic equation

$$Q_m(s)L'(s)R(s) + P(s)Z(s) = 0 \tag{5}$$

has order $2n+q-1$. The objective now is to choose P and L' such that

$$Q_m(s)L'(s)R(s) + P(s)Z(s) = A^*(s) \tag{6}$$

is satisfied for a given monic Hurwitz polynomial $A^*(s)$ of degree $2n+q-1$. Because of assumptions S2 and S3 guarantee that Q_m, R and Z are coprime, there is a solution so that L' and P satisfy (6) and this solution is unique [10].

Using (6), the closed-loop is described by

$$y = \frac{ZM}{A^*} r \tag{7}$$

Similarly, from the plant in (1) and the control law in (3) and (6), we obtain

$$u = \frac{RM}{A^*} r \tag{8}$$

Because r is uniformly bounded and $\frac{ZM}{A^*}, \frac{RM}{A^*}$ are proper with stable poles, y and u remain bounded whenever $t \to \infty$ for any polynomial $M(s)$ of degree $n + q - 1$ [10]. Therefore, the pole placement objective is achieved by the control law (3) without having to put any additional restrictions on $M(s)$ and $Q_m(s)$. When $r = 0$, (7), (8) imply that y and u converge to zero exponentially fast.

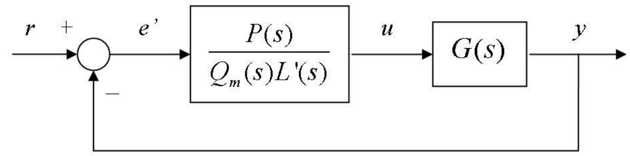

Fig. 1. Block diagram of pole placement control.

When $r \neq 0$, the tracking error $e' = r - y$ is given by

$$e' = \frac{A^* - ZM}{A^*} r = \frac{L'R}{A^*} Q_m r - \frac{Z}{A^*}(M - P) r \tag{9}$$

In order to obtain zero tracking error, (9) suggests the choice of $M(s) = P(s)$ to cancel the second term in (9). The first term in (9) is canceled by using $Q_m r = 0$. Therefore, the pole placement and tracking objective are achieved by using the control law

$$Q_m L' u = -P(y - r) \tag{10}$$

which is implemented as shown in Fig. 1 using $n + q - 1$ integrators to realize $C(s) = \frac{P(s)}{Q_m(s)L'(s)}$. Because $L'(s)$ is not necessarily Hurwitz, the realization of (10) with $n + q - 1$ integrators may have a transfer function, namely $C(s)$ unstable. An alternative realization of (10) is obtained by rewriting (10) as

$$u = \frac{F - Q_m L'}{F} u - \frac{P}{F}(y - r) \tag{11}$$

where F is any monic Hurwitz polynomial of degree $n + q - 1$.

III. Variable Structure Adaptative Pole Placement Control

Let us rewrite the plant in (1) as

$$\begin{cases} \dot{x} = Ax + Bu \\ y = h^T x \end{cases} \tag{12}$$

where $x \in R^n$, $x(0) = 0$. Let us suppose that only y and u are available for measurement and the plant parameters are unknown. Equation (12) may also be written as

$$y(s) = h^T (sI - A)^{-1} B u(s)$$

The plant equation (1) may be expressed as

$$(s^n + a_{n-1}s^{n-1} + \cdots + a_1 s + a_0)y(s) = (b_{n-1}s^{n-1} + \cdots + b_1 s + b_0)u(s)$$

Therefore,

$$s^n y(s) = (b_{n-1}s^{n-1} + \cdots + b_1 s + b_0)u(s) - (a_{n-1}s^{n-1} + \cdots + a_1 s + a_0)y(s) \tag{13}$$

If we write all the parameters in (13) in the parameter vector

$$\theta^* = [b_{n-1}, \ldots, b_1, b_0, a_{n-1}, \ldots, a_1, a_0]^T = [\theta_1^*, \ldots, \theta_n^*, \ldots, \theta_{2n}^*]^T$$

and all I/O signals and their derivatives in the signal vector

$$Y = [s^{n-1}u(s), \ldots, su(s), u(s), -s^{n-1}y(s), \ldots, -sy(s), -y(s)]^T$$

we have

$$s^n y(s) = \theta^{*T} Y(s) \tag{14}$$

Because in most applications the only signals available for measurement is the input u and output y and the use of differentiation is not desirable, the use of the signals $y^{(n)}$ and Y should be avoided. One way to avoid them is to filter each side of (14) with an nth-order stable filter $\frac{1}{\Lambda(s)}$ with

$$\Lambda(s) = s^n + \lambda_{n-1}s^{n-1} + \cdots + \lambda_0$$

Therefore,

$$\frac{s^n y(s)}{\Lambda(s)} = \frac{\theta^{*T} Y(s)}{\Lambda(s)} = \theta^{*T} \frac{1}{s^n} \frac{s^n Y(s)}{\Lambda(s)}$$

Defining $\frac{s^n}{\Lambda(s)} = W(s)$, we obtain

$$W(s)y(s) = \theta^{*T} \frac{1}{s^n} W(s)Y(s) = W(s)\frac{\theta^{*T} Y(s)}{s^n}$$

Considering $z(s) = W(s)y(s)$ we have

$$z(s) = W(s)\theta^{*T}\psi(s) \tag{15}$$

where

$$\psi(s) = \frac{Y(s)}{s^n}$$

Usually, $W(s)$ is not SPR (strictly positive real). Then, a polynomial $L(s)$ is chosen so that WL is strictly proper and SPR. Therefore

$$z(s) = W(s)L(s)\theta^{*T} \frac{\psi(s)}{L(s)}$$

Defining $\phi(s) = \frac{\psi(s)}{L(s)} = L^{-1}(s)\psi(s)$, we have

$$z(s) = W(s)L(s)\theta^{*T}\phi(s) \tag{16}$$

Let $\theta(t)$ be the estimate of θ^* at time t. Then the estimate $\hat{z}$ of z at time t is constructed as

$$\hat{z} = W(s)L(s)\theta^T\phi(s) \tag{17}$$

The estimation error e_0 is generated as

$$e_0 = z - \hat{z} \tag{18}$$

and the normalized estimation error as

$$\varepsilon_0 = z - \hat{z} - W(s)L(s)\varepsilon_0 n_s^2 \tag{19}$$

where n_s is the normalized signal which we design to satisfy

$$\frac{\phi}{m} \in L_\infty, \; m^2 = 1 + n_s^2 \tag{20}$$

Typical choices for n_s that satisfy (20) are $n_s^2 = \phi^T\phi$, $n_s^2 = \phi^T P\phi$ for any $P = P^T > 0$, etc. When $\phi \in L_\infty$, (20) is satisfied with $m = 1$, i.e., $n_s = 0$, and in this case, $\varepsilon_0 = e_0$.

If we use $\tilde{\theta} = \theta - \theta^*$ in (19), we obtain

$$\varepsilon_0 = W(s)L(s)(-\tilde{\theta}^T\phi - \varepsilon_0 n_s^2) \tag{21}$$

Let us consider the following state space representation of (21):

$$\begin{cases} \dot{e} = A_c e + b_c(-\tilde{\theta}^T\phi - \varepsilon_0 n_s^2) \\ \varepsilon_0 = h_c^T e \end{cases} \tag{22}$$

where A_c, b_c and h_c are the matrices associated with a state space representation for the transfer function $W(s)L(s) = h_c^T(sI - A_c)^{-1}b_c$.

By the SPR property of $W(s)L(s) = h_c^T (sI - A_c)^{-1} b_c$ and the Lefschetz-Kalman-Yakubovitch Lemma $\exists P = P^T > 0$ for some matrix $Q = Q^T > 0$ such that

$$\begin{cases} A_c^T P + PA_c = -2Q \\ Pb_c = h_c \end{cases} \tag{23}$$

In the traditional algorithms (with integral estimation laws) we use

$$\dot{\theta} = \Gamma \varepsilon_0 \phi \tag{24}$$

Let us now consider the following Lyapunov-like function

$$V(e,\tilde{\theta}) = \frac{1}{2} e^T Pe + \frac{1}{2} \tilde{\theta}^T \Gamma^{-1} \tilde{\theta} > 0,\ \Gamma = \Gamma^T > 0 \tag{25}$$

The time derivative $\dot{V}$, using (22) and (23), is given by

$$\dot{V}(e,\tilde{\theta}) = \frac{1}{2}\{e^T (A_c^T P + PA_c) e$$
$$+ 2e^T Pb_c (-\tilde{\theta}^T \phi - \varepsilon_0 n_s^2) + 2\tilde{\theta}^T \Gamma^{-1} \dot{\tilde{\theta}}\}$$
$$\dot{V}(e,\tilde{\theta}) = \frac{1}{2}\{e^T (-2Q) e + 2\varepsilon_0 (-\tilde{\theta}^T \phi - \varepsilon_0 n_s^2) + 2\tilde{\theta}^T \Gamma^{-1} \dot{\tilde{\theta}}\}$$
$$\dot{V}(e,\tilde{\theta}) = -e^T Qe - \varepsilon_0^2 n_s^2 - \varepsilon_0 \tilde{\theta}^T \phi + \tilde{\theta}^T \Gamma^{-1} \dot{\tilde{\theta}}$$

Since $\dot{\tilde{\theta}} = \dot{\theta} = \Gamma \varepsilon_0 \phi$, we have

$$\dot{V}(e,\tilde{\theta}) = -e^T Qe - \varepsilon_0^2 n_s^2 - \varepsilon_0 \tilde{\theta}^T \phi + \tilde{\theta}^T \Gamma^{-1} \Gamma \varepsilon_0 \phi$$
$$\dot{V}(e,\tilde{\theta}) = -e^T Qe - \varepsilon_0^2 n_s^2 \le 0$$

which guarantees that $[e^T,\ \tilde{\theta}^T]^{\mathrm{T}} = [0,\ 0]^{\mathrm{T}}$ is a stable equilibrium point.

Now, we consider the following switching laws for θ

$$\theta_i = \bar{\theta}_i \operatorname{sgn}(\varepsilon_0 \phi_i),\ \bar{\theta}_i > |\theta_i^*| \tag{26}$$

Let us choose the Lyapunov function

$$V(e) = \frac{1}{2} e^T Pe$$

Then,

$$\dot{V}(e) = -e^T Qe - \varepsilon_0^2 n_s^2 - \varepsilon_0 \tilde{\theta}^T \phi$$
$$\dot{V}(e) = -e^T Qe - \varepsilon_0^2 n_s^2 - \varepsilon_0 (\theta - \theta^*)^T \phi$$
$$\dot{V}(e) = -e^T Qe - \varepsilon_0^2 n_s^2 - \varepsilon_0 (\theta_1 - \theta_1^*)\phi_1 - \cdots - \varepsilon_0 (\theta_{2n} - \theta_{2n}^*)\phi_{2n}$$
$$\dot{V}(e) = -e^T Qe - \varepsilon_0^2 n_s^2 - [\bar{\theta}_1 \operatorname{sgn}(\varepsilon_0 \phi_1) - \theta_1^*]\varepsilon_0 \phi_1$$
$$- \cdots - [\bar{\theta}_{2n} \operatorname{sgn}(\varepsilon_0 \phi_{2n}) - \theta_{2n}^*]\varepsilon_0 \phi_{2n}$$
$$\dot{V}(e) = -e^T Qe - \varepsilon_0^2 n_s^2 - (\bar{\theta}_1 |\varepsilon_0 \phi_1| - \theta_1^* \varepsilon_0 \phi_1)$$
$$- \cdots - (\bar{\theta}_{2n} |\varepsilon_0 \phi_{2n}| - \theta_{2n}^* \varepsilon_0 \phi_{2n})$$

Since $\bar{\theta}_i > |\theta_i^*|$, we have

$$\dot{V}(e) \le -e^T Qe - \varepsilon_0^2 n_s^2 < 0$$

which guarantees that $e = 0$ is a globally asymptotically stable equilibrium point.

IV. A Controller Application

Let us illustrate the design of a VS-APPC scheme using a second order plant model.

Consider the plant

$$y = \frac{b_1 s + b_0}{s^2 + a_1 s + a_0} u \tag{27}$$

where b_1, b_0, a_1 and a_0 are unknown constants and the plant input u is chosen so that the poles of the closed-loop plant are the roots of $A^*(s) = (s+1)^4 = 0$ and y tracks a constant reference signal, $\forall t \ge 0$.

A. Calculation of Controller Parameters

As shown in Section II, the control law in (11) can be used to achieve the control objective, where $F(s) = s^2 + f_1 s + f_0$, $L'(s) = s + l_0$, $P(s) = p_2 s^2 + p_1 s + p_0$ and the coefficients l_0, p_2, p_1 and p_0 of $P(s)$ satisfy the Diophantine equation

$$\begin{aligned} &s(s + l_0)(s^2 + a_1 s + a_0) \\ &+ (p_2 s^2 + p_1 s + p_0)(b_1 s + b_0) = (s+1)^4 \end{aligned} \tag{28}$$

Equation (28) may be also written in the form of the Sylvester matrix of $Q_m R$ and Z:

$$\begin{bmatrix} 1 & 0 & 0 & 0 & 0 \\ a_1 & 1 & b_1 & 0 & 0 \\ a_0 & a_1 & b_0 & b_1 & 0 \\ 0 & a_0 & 0 & b_0 & b_1 \\ 0 & 0 & 0 & 0 & b_0 \end{bmatrix} \begin{bmatrix} 1 \\ l_0 \\ p_2 \\ p_1 \\ p_0 \end{bmatrix} = \begin{bmatrix} 1 \\ 4 \\ 6 \\ 4 \\ 1 \end{bmatrix}$$

whose solution is

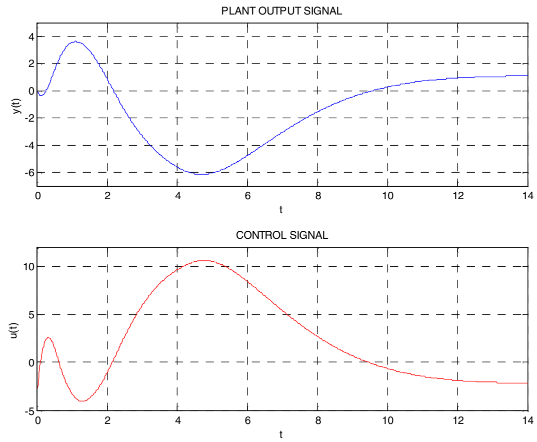

Fig. 2. Indirect APPC using gradient method.

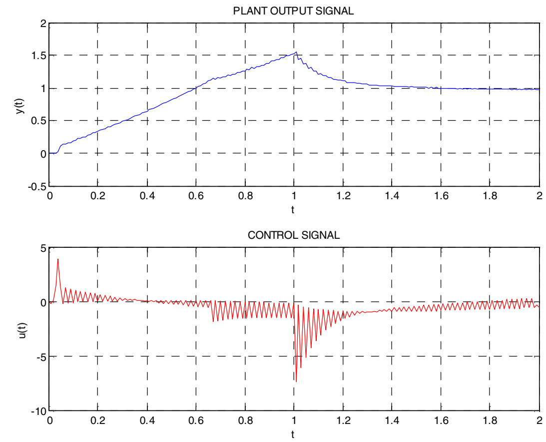

Fig. 3. VS-APPC for a second order system.

$$
\begin{aligned}
l_0 &= \frac{b_0^2 b_1 (a_0 - 6) + b_0^3 (4 - a_1) + 4 b_1^2 b_0 - b_1^3}{b_0^3 - a_1 b_1 b_0^2 + a_0 b_1^2 b_0} \\
p_2 &= \frac{4 - a_1 - l_0}{b_1} \\
p_1 &= \frac{b_0 (4 - a_0 l_0) - b_1}{b_0^2} \\
p_0 &= \frac{1}{b_0}
\end{aligned}
\tag{29}
$$

When the plant parameters are known with uncertainties, the certainty equivalence principle suggests the using of the same control law, but with the controller polynomials $P(s) = p_2 s^2 + p_1 s + p_0$ and $L'(s) = s + l_0$ calculated by using the estimates of the parameters, and, therefore, we have

$$
\begin{aligned}
\hat{l}_0 &= \frac{\hat{b}_0^2 \hat{b}_1 (\hat{a}_0 - 6) + \hat{b}_0^3 (4 - \hat{a}_1) + 4 \hat{b}_1^2 \hat{b}_0 - \hat{b}_1^3}{\hat{b}_0^3 - \hat{a}_1 \hat{b}_1 \hat{b}_0^2 + \hat{a}_0 \hat{b}_1^2 \hat{b}_0} \\
\hat{p}_2 &= \frac{4 - \hat{a}_1 - \hat{l}_0}{\hat{b}_1} \\
\hat{p}_1 &= \frac{\hat{b}_0 (4 - \hat{a}_0 \hat{l}_0) - \hat{b}_1}{\hat{b}_0^2} \\
\hat{p}_0 &= \frac{1}{\hat{b}_0}
\end{aligned}
\tag{30}
$$

where $\hat{l}_0$, $\hat{p}_2$, $\hat{p}_1$ and $\hat{p}_0$ are the controller parameters estimates that must be generated on-line. In the traditional indirect APPC scheme, adaptive laws driven by the error ε_0 are used. To achieve this, it may be used the gradient method, the least squares method, etc. The parameter estimation by gradient method is given by (24). For the VS-APPC scheme the adaptive laws will be replaced by switching laws as in (26).

Since the controller parameters can be functions of more than one plant parameter simultaneously, the signal may come undefined, due to high frequency switching signals. Thus, it's necessary the introduction of a nominal value of all the parameters, in order to maintain the value with a defined signal. Rewriting the switching laws with nominal values, we have

$$
\theta_i = \bar{\theta}_i \operatorname{sgn}(\varepsilon_0 \phi_i) + \theta_{i,nom}, \quad \bar{\theta}_i > | \theta_i^* - \theta_{i,nom} | \tag{31}
$$

where $\theta_{i,nom}$ is a nominal value for the parameter θ_i^*.

The control signal u is generated from (11).

B. Simulations and results

Let us consider, for the simulations, the nonminimum phase unstable plant

$$
G(s) = \frac{2s - 1}{s^2 - 3s + 2}
$$

The simulations were carried out with an integration step of h = 0.01s, the polynomials $\Lambda(s) = F(s) = s^2 + 3.2s + 2.5$, and a reference signal r = 1.

Initially, a simulation was done using the gradient method in the adaptive laws. In this simulation (Fig. 2), the plant output matches the reference model in 12.32s.

In the next simulation (Fig. 3), the conventional adaptive laws were replaced by the switching laws proposed in (31).

We used the constants $\theta_{1,nom} = 3$, $\bar{\theta}_1 = 2$, $\theta_{2,nom} = 3$, $\bar{\theta}_2 = 4.5$, $\theta_{3,nom} = -1$, $\bar{\theta}_3 = 4.5$, $\theta_{4,nom} = 1.5$, $\bar{\theta}_4 = 2$. It was verified that the plant output matches the reference model in 1.31s. VS-APPC presented a fast transient and a less oscillatory output.

The indirect VS-APPC scheme is easy to design, with the relays amplitude calculation directly related to the plant physical parameters.

V. Conclusion

In this paper, a new design technique was proposed for variable structure adaptive pole placement controllers in an indirect approach. It was presented the simulation results for a second order plant.

The proposed technique has presented a very fast transient.

As suggestions to the practical application of this work we could mention areas like robotics, process control, motor speed control, etc.

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