Control of uncertain systems by feedback linearization with neural networks augmentation. Part II. Controller validation by numerical simulation

Ioan URSU*, Adrian TOADER*

*Corresponding author
INCAS - National Institute for Aerospace Research "Elie Carafoli"
Bdul Iuliu Maniu 220, Bucharest 061136, Romania
iursu@incas.ro
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Abstract: *The paper was conceived in two parts. Part I, previously published in this journal, highlighted the main steps of adaptive output feedback control for non-affine uncertain systems, having a known relative degree. The main paradigm of this approach was the feedback linearization (dynamic inversion) with neural network augmentation. Meanwhile, based on new contributions of the authors, a new paradigm, that of robust servomechanism problem solution, has been added to the controller architecture. The current Part II of the paper presents the validation of the controller hereby obtained by using the longitudinal channel of a hovering VTOL-type aircraft as mathematical model.*

Key Words: uncertain systems, relative degree, adaptive control, pseudo control, feedback linearization, dynamic inversion, neural network, robust servomechanism problem, Kalman synthesis, internal model, stabilizing compensator, VTOL-type aircraft, numerical simulation.

1. INTRODUCTION

The paper was conceived in two parts. Part I, previously published in this journal [1], highlighted the main steps of adaptive output feedback control for non-affine uncertain systems, having a known relative degree [2]. A main paradigm of the approach has been the feedback linearization (dynamic inversion) with neural network augmentation. Meanwhile, two contributions of the authors were published [3], [4], centred on the works [5], [6], thus proposing a new unitary approach on adaptive control synthesis. This approach supposes the addendum of another paradigm, namely that of designing a stabilizing compensator for a pair plant-internal model of exogeneous signals. The current Part II of the paper presents the validation of the controller hereby obtained by using the longitudinal channel of a hovering VTOL-type aircraft as mathematical model.

2. NEW UNITARY ADAPTIVE CONTROL DESIGN

The control design parameters are herein step by step presented. For the sake of friendliness and clarity of the Section, let's recall the basic architecture of the control system (Fig. 1) and the improved, versus the work [1], implementation block diagram (Fig. 2).

The plant dynamics (Fig. 2) refers to a service model – the longitudinal channel of a hovering VTOL-type aircraft [7] (see (1), τ is the actuator time constant. *It is worthy to note that this mathematical model is only a pretext in order to validate the controller by numerical simulations; in fact, our approach on controller design is a model free one*):

$$\begin{bmatrix}\dot{u}_g\\ \dot{u}\\ \dot{x}\\ \dot{q}\\ \dot{\theta}\\ \dot{\delta}\end{bmatrix}=\begin{bmatrix}-0.314 & 0 & 0 & 0 & 0 & 0\\ X_u & X_u & 0 & 0 & -g & 0\\ 0 & 1 & 0 & 0 & 0 & 0\\ M_u & M_u & 0 & M_q & 0 & M_\delta\\ 0 & 0 & 0 & 1 & 0 & \delta\\ 0 & 0 & 0 & 0 & 0 & -1/\tau\end{bmatrix}\begin{bmatrix}u_g\\ u\\ x\\ q\\ \theta\\ \delta\end{bmatrix}+\begin{bmatrix}0\\ 0\\ 0\\ 0\\ 0\\ 1/\tau\end{bmatrix}\delta_c+\begin{bmatrix}1\\ 0\\ 0\\ 0\\ 0\\ 0\end{bmatrix}w \qquad (1)$$

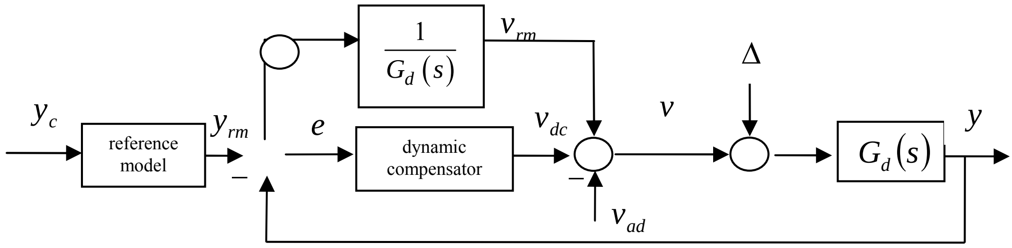


Fig. 1. Control system architecture

or, in matrix form and with added measurement

$$\dot{x}=Ax(t)+B\delta_c(t)+Ew(t),\ y=\theta \qquad (2)$$

with u_g − longitudinal component of the gust velocity [m/s], obtained as a filtered white noise *w*; u − velocity perturbation along the *x* axis [m/s]; θ − pitch attitude [rad]; $q=\dot{\theta}$ − pitch rate, rad/sec; δ − actuator state [m]; δ_c − control stick input (control variable) [m]; M_u − speed stability parameter [rad/m-sec]; M_q − pitch rate damping [1/sec]; M_δ − control sensitivity [rad/sec^2/m]; X_u − longitudinal drag parameter [1/sec]; g − gravitational constant, 9.81 [m/sec^2].

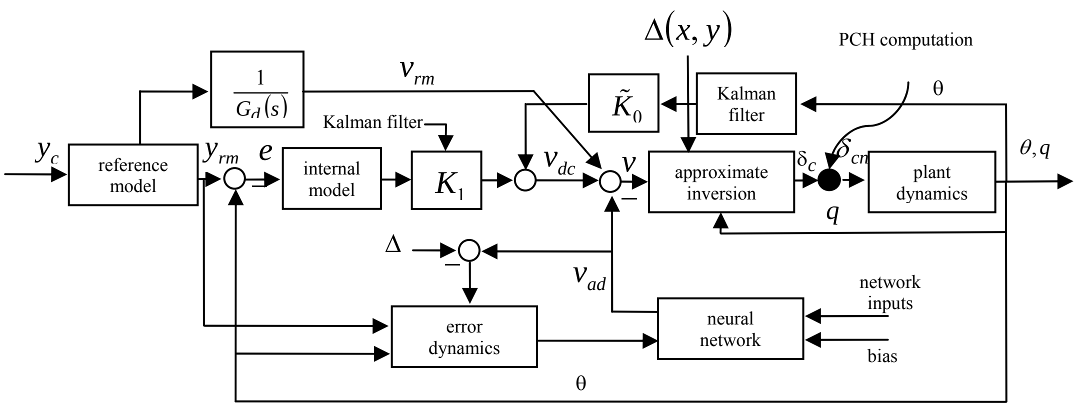


Fig. 2. Block diagram implementation

The control objective is the following: the system output y is required to track a known bounded input y_c. The main sources of uncertainty are the neglected dynamics and control delay time: $\delta_c(t)$ is really $\delta_c(t-T_d)$. As main knowledge about the system, *the relative degree was assumed*: one can see that in (1)-(2) *the controlled output* θ *has the relative degree* $r=3$.

The pseudo control v, Fig. 1, 2, is chosen to have the form [8]-[13]

$$v = v_{rm} + v_{dc} - v_{ad} \tag{3}$$

The three components are: v_{rm} – the output of a reference model, v_{dc} – the output of a stabilizing compensator for the linearized dynamics with $\Delta = 0$ and v_{ad} – the adaptive control signal designed to approximately cancel the dynamic inversion error Δ. Thus, the control objective is now the following: the system output y is required to track a known bounded input y_{rm}, rather than y_c.

The component v_{dc} requires a special attention. By virtue of assuming in synthesis only the minimal knowledge about the relative degree, the dynamics of the output, with a key parameter b_0 in design, should be written as follows (see Fig. 1), (where s is Laplace variable)

$$y = G_d(s)(v+\Delta),\ G_d(s) := b_0 / s^3 \tag{4}$$

Aiming to correlate the blocks in order to simplify the structure, the block in upper loop of Fig. 1 should be conceived as

$$v_{rm} = y_{rm} / G_d(s) \tag{5}$$

Substituting (5) in (3), taking into account the substratum of v_{ad} synthesis, one gets the error dynamics.

$$e + b_0(v_{dc} - v_{ad} + \Delta)/s^3 = 0, e := y_{rm} - y \text{ or } e + b_0 v_{dc}/s^3 \cong 0 \tag{6}$$

At this point, an ordinary reflection concerns the necessity of completing the error dynamics by introducing a stabilizing compensation by means of a v_{dc} control component. The treatment of the question in the quoted references (e.g., [10]) suffers of some lack of coherence and clarity concerning the theory and exemplification of compensation selection. In [3], [4], a unitary viewpoint of approach was proposed. Consider, thence, the output dynamics as object and framework of v_{dc} component synthesis. The procedure used in [5]-[6] is invoked: for the order three integrator- type plant (4) (with the state vector x), two compensators are designed, a) a servocompensator (internal model) (with the state vector η) and b) a stabilizing compensator (with the state vector $\hat{x}_E$), see Fig. 2. As stabilizing compensator for the system (x, η) is considered the well known Kalman filter [14]-[16]. Two pairs of matrices are thus introduced: the pair (Q_w, Q_ξ) of white noise matrices, parameters of the estimation Riccati equation, which provides filter gain K_f, and the pair of weighting matrices (Q_J, R_J), parameters of the control Riccati equation, providing the control gain K_R. Q_w stands for formal state noise and Q_ξ stands for formal measurement noise. Q_J weights the performance output $y_p := \eta = C_p x_E$, as a measure of e integral error, and R_J weights the control $v_{dc} := u$. Then, the open loop triplet system will have the form

$$\begin{gathered}
\dot{x} = Ax + Bv_{dc},\ A = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ 0 & 0 & 0 \end{bmatrix},\ B = \begin{bmatrix} 0 \\ 0 \\ b_0 \end{bmatrix},\ x = \begin{bmatrix} y \\ \dot{y} \\ \ddot{y} \end{bmatrix} \\
\dot{\eta} = A_c\eta + B_c\left(y_{rm} - C_o x\right) \\
\dot{\hat{x}}_E = A_E\hat{x}_E + B_E v_{dc} + K_f\left(y - C_E\hat{x}_E\right) + D_E y_{rm},\ x_E := \begin{bmatrix} x \\ \eta \end{bmatrix},\ \hat{x}_E := \begin{bmatrix} \hat{x} \\ \hat{\eta} \end{bmatrix} \\
C_E = \begin{bmatrix} C_o & 0 & 0 \end{bmatrix},\ A_E := \begin{bmatrix} A & 0 \\ -B_c C_o & A_c \end{bmatrix},\ B_E := \begin{bmatrix} B \\ 0 \end{bmatrix},\ D_E := \begin{bmatrix} 0 \\ B_c \end{bmatrix},\ C_o := \begin{bmatrix} 1 & 0 & 0 \end{bmatrix},\ y := C_o x
\end{gathered} \tag{7}$$

The state vector η must have in principle the dimension $r - 1 = 2$ [13], and the selection of matrices A_c, B_c aims to obey the property of internal model for the step input signals [6]: $\|A_c\| \ll \|B_c\|$. The Kalman filter output $K_R\hat{x}_E$ will be used as control variable v_{dc} in the following manner (available η is taken in calculation)

$$K_R = \left[K_0, K_1\right],\ v_{dc} = -K_0\hat{x} - K_1\eta = -\tilde{K}_0 x_E - K_1\eta,\ \tilde{K}_0 = \left[K_0 \quad 0 \quad 0\right] \tag{7'}$$

Thus, the closed loop system is given by

$$\begin{bmatrix} \dot{x} \\ \dot{\eta} \\ \dot{\hat{x}}_E \end{bmatrix} = \begin{bmatrix} A & -BK_1 & -B\tilde{K}_0 \\ -B_c C_o & A_c & 0 \\ K_f C_o & -B_E K_1 & A_E - B_E\tilde{K}_0 - K_f C_E \end{bmatrix} \begin{bmatrix} x \\ \eta \\ \hat{x}_E \end{bmatrix} + \begin{bmatrix} 0 \\ B_c \\ D_E \end{bmatrix} y_{rm} := A_{cl} \begin{bmatrix} x \\ \eta \\ \hat{x}_E \end{bmatrix} + \begin{bmatrix} 0 \\ B_c \\ D_E \end{bmatrix} y_{rm} \tag{8}$$

and the stability of matrix A_{cl} must be proved, due to the substitution of $\hat{\eta} \to \eta$.

Bringing now in attention the complete equation of error (6), let proceed therein to the substitution of the control component value $v_{dc} = -\tilde{K}_0 x_E - K_1\eta$. The *error dynamics* system will have the form ($e_1 := e$)

$$\begin{bmatrix} \dot{e}_1 \\ \dot{e}_2 \\ \dot{e}_3 \\ \dot{\eta} \\ \dot{\hat{x}}_E \end{bmatrix} = \begin{bmatrix} 0 & 1 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & b_0K_1 & b_0\tilde{K}_0 \\ B_c & 0 & 0 & A_c & 0 \\ 0 & 0 & 0 & -B_E K_1 & A_E - B_E\tilde{K}_0 - K_f C_E \end{bmatrix} \begin{bmatrix} e_1 \\ e_2 \\ e_3 \\ \eta \\ \hat{x}_E \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ K_f \end{bmatrix} y + \begin{bmatrix} 0 \\ 0 \\ b_0 \\ 0 \\ 0 \end{bmatrix} \left(v_{ad} - \Delta\right) + \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ D_E \end{bmatrix} y_{rm} \tag{9}$$

or, in the matrix-vector description

$$\dot{E} = \bar{A}E + \bar{B}\left(v_{\mathrm{ad}} - \Delta\right) + \bar{D}_1 y + \bar{D}_2, y_{rm} := \bar{A}E + \bar{B}\left(v_{\mathrm{ad}} - \Delta\right) + \bar{D}\bar{y},\ \bar{D} := \left[\bar{D}_1 \quad \bar{D}_2\right],\ \bar{y} := \begin{bmatrix} y \\ y_{rm} \end{bmatrix} \tag{9'}$$

Summarizing the proposed procedure for the control component v_{dc} synthesis, we have to run on computer the system (9) with the inputs $v_{ad} - \Delta$, $\bar{y}$, and providing the output E as input for the equations of the control component v_{ad} synthesis. Thus, it is important to underline the evading in this work, as not being strictly necessary, of the error estimation $\hat{E}$ [9]-[13], [17], [18]. Also, it is worthy to mention the following: the choice of the key parameters $b_0, Q_w, Q_\xi, Q_J, R_J, C_p$ is performed until this choice leads, by a *trial and error* process, to a *stable matrix* $\bar{A}$.

The procedures of designing the adaptive components v_{rm} and v_{ad} remain valid as described in Part I of the paper, but with the mention that a proof of the stable working of the controller (3) was done in [4]. Thus, the representation for the reference model is chosen as a third order filter of the input

$$y_{rm} = \frac{\omega_1 \omega_2^2}{\left(s+\omega_1\right)\left(s^2 + 2\zeta\omega_2 s + \omega_2^2\right)} y_c \tag{10}$$

where $\omega_1, \zeta, \omega_2$ stockpile some information – if this is available – about the basic, low modes, of the plant (in our case, represented by the system (2)). The structure of the adaptive component is more complicated, but for reasons of conformity it will be shortly presented furher on.

Given $z \in R^{n_1}$, a three layer-layer neural network (NN) (with a single hidden layer) has an output given by

$$f_k = b_W \theta_{Wk} + \sum_{j=1}^{n_2} w_{j,k}\sigma_j, k = 1,\ldots,n_3, \sigma_j = \sigma\left(b_V \theta_{Vj} + \sum_{i=1}^{n_1} v_{i,j} z_i\right), j = 1,\ldots,n_2 \tag{11}$$

where n_1, n_2, and n_3 are the numbers of input nodes, hidden layer nodes, and outputs, respectively. $\sigma(\cdot)$ is so called activation function, $v_{i,j}$ are the first-to-second layer interconnection weights, $w_{j,k}$ are the second to third layer interconnection weights, b_V, b_W are bias terms, θ_{Vj} acts as thresholds for each neuron, θ_{Wk} allows the bias term b_W to be weighted in each output channel. In fact, such a structure, $f = W^{\mathrm{T}}\sigma(z)$, acts as an universal approximator of piecewise continuous nonlinearities with squashing activation functions z [19]. Accordingly, the dynamic inversion error Δ (Figs. 1, 2) can be approximated by a NN. A contextual result was proven in [4] in the form: *Given* $\varepsilon^* > 0$, *there exists a set of bounded weights,* $W \in R^{(n_2+1)\times(n_3)}$, $V \in R^{(n_1+1)\times(n_2)}$ *such that*

$$\Delta = W^{\mathrm{T}}\sigma\left(V^{\mathrm{T}}\mu\right) + \varepsilon(\mu), \|\varepsilon\| < \varepsilon^* \tag{12}$$

where $\mu(t) = \left[1 \quad \bar{v}_d^{\mathrm{T}}(t) \quad \bar{y}_d^{\mathrm{T}}(t)\right]^{\mathrm{T}}, \mu \in R^{n_1+1}$ *is the past input/output history vector,* $\sigma(\cdot)$ *is given by*

$$\sigma_i = \sigma(z_i) = \frac{1}{1+e^{-a_i z_i}}, \sigma := \sigma\left(V^{\mathrm{T}}\mu\right),\ \sigma'(z) := \mathrm{diag}\left(\frac{\partial \sigma_i}{\partial z_i}\right) \tag{13}$$

$\varepsilon(z)$ *is the functional reconstruction error and* a *is an activation potential. Thus*

$$v_{ad} := \hat{W}^{\mathrm{T}}\sigma\left(\hat{V}^{\mathrm{T}}\mu\right) \tag{14}$$

$$\dot{\hat{V}} = -\Gamma_V \left[\mu E^{\mathrm{T}} P \bar{B} \hat{W}^{\mathrm{T}} \sigma' + k \left(\hat{V} - V_0 \right) \right], \dot{\hat{W}} = -\Gamma_W \left[\left(\sigma - \sigma' \hat{V}^{\mathrm{T}} \mu \right) E^T P \bar{B} + k \left(\hat{W} - W_0 \right) \right], \bar{A}^{\mathrm{T}} P + P \bar{A} = -Q \tag{14'}$$

(V_0, W_0) are initial guess of NN weights, $Q > 0$ is a suitable matrix, $k > 0$ is a sufficiently large constant adaptation gain, and Γ_V, $\Gamma_W > 0$, are matrices of appropriate dimensions. $\hat{V}$ and $\hat{W}$ are the inner (hidden) layer weight matrix and the outer layer weight vector which must be updated on line, respectively.

As concerning the approximate inversion law (Fig. 2), this will provide the real control $\delta_c \equiv u$. To be consequent in our approach, let's assume an enhanced level of uncertainty and evade the direct use of equations (1). A simple, heuristic approach of flight mechanics enables us a series of inferences on the dynamics of output $y = \theta$

$$y = \theta, \dot{y} := q, \ddot{y} \approx M_q q + M_\delta \delta, \dddot{y} := g_r = M_q \left(M_q q + M_\delta \delta \right) - M_\delta \delta / \tau + M_\delta \delta_c / \tau, \hat{g}_r = M_q^2 q + M_\delta \delta_c / \tau \tag{15}$$

Assume however, that above g_r is not the exact expression derived from applying feedback linearization mapping on (1). Now taking into account (4),

$$b_0 \nu \cong \dddot{y} = M_\delta \delta_c / \tau + M_q^2 q \tag{16}$$

therefore the inversion is performed

$$u := \delta_c = \tau \left[b_0 \nu - \hat{M}_q^2 q \right] / \hat{M}_\delta \tag{17}$$

where $\hat{M}_q, \hat{M}_\delta$ were introduced for parametric uncertainties in $\hat{M}_q, \hat{M}_\delta$, respectively. Note

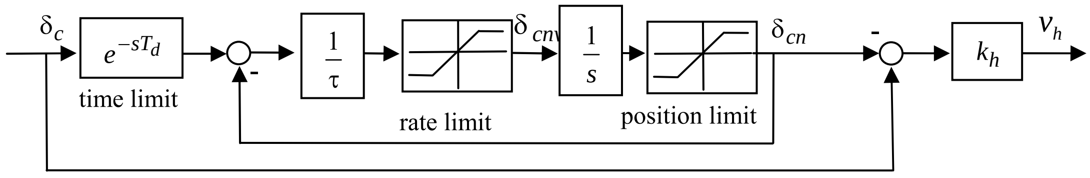


Fig. 3. Computation of the pseudo control hedge – PCH – ν_h

that with measured output $y = \theta$, the derivative $\dot{y} = q$ is available and considered in the above relations, but the variable δ is ignored.

At last, let's remember the Pseudo Control Hedge (PCH) procedure [8] of limiting the pseudo control to prevent shortcomings such as time delay, actuator position and rate limits. The idea is simple: an estimate of the actuator position is firstly obtained, and then this estimate is used to compute the difference between the commanded pseudo control ν and the estimated achievable pseudo control (see Figs. 3, 4, and Fig. 2)

$$\nu_h = h(\xi, \delta_c) - \hat{h}(\xi, \delta_{cn}) = \hat{M}_\delta / b_0 \tau \left(\delta_c - \delta_{cn} \right) := k_h \left(\delta_c - \delta_{cn} \right) \tag{18}$$

The pseudo control hedge is then substracted from the third derivative of the reference model, see Fig. 4.

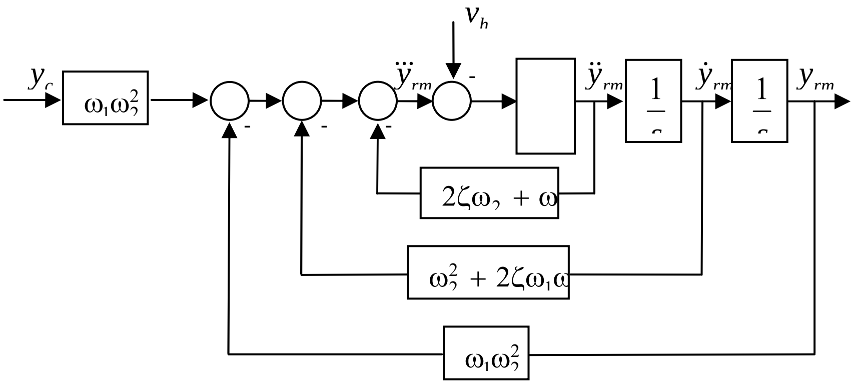


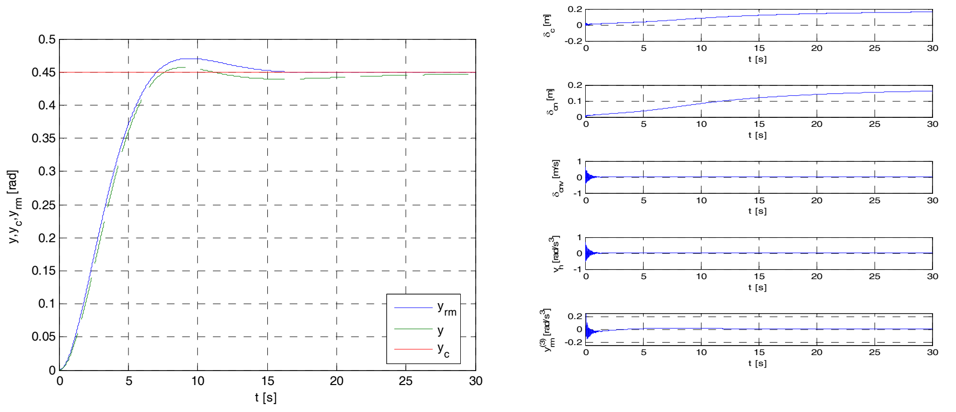
Fig. 4. Incorporation of the PCH ν_h in the reference model

3. CONTROLLER VALIDATION

An excellent working of the achieved control law, in conditions of increased parameter structural uncertainty concerning the system, was demonstrated in [3], considering a different mathematical model. Herein, and in [4], the simulations are performed on the mathematical model (1) and in the presence of the filtered white noise w having the intensity $Q_w = 0.0001$ m/s, with the filter pole at 0.314 rad/sec; the nonlinear effects were neglected. The maximum accepted level for the u_g state variable in (1) was set at 5.14 ft/s and the saturation level for the control δ_c is 0.17 m [7]. For the sake of rigor, canonical coordinates transformation [2] on system (1) was done and the zero dynamics were proved to be asymptotically stable. Further on, the system parameters were as follows: $M_\delta = 16.968\ (\text{rad}\times\text{s}^2)/\text{m}$, $M_q = -3\ \text{s}^{-1}$, $\hat{M}_\delta = 0.6 M_\delta$, $\hat{M}_q = 2 M_q$, $X_u = -0.1\ \text{s}^{-1}$, $n_2 = 5$, $n_1 = 15$, $M_u = 0.0679$ rad/(m×s), $b_0 = 10$, ω_1=33 rad/s, $\tau = 0.03\ s$, $\zeta = 0.7$, $\omega_2 = 0.466$ rad/s, $Q_J = 100$, $Q_\xi = 100$, $R_J = 10^{-6}$, $C_p = [0\ \ 0\ \ 0\ \ 100\ \ 100]$, $\Gamma_W = I_6$, $T_d = 0.005$ s; $\delta_{cnv} \in [-0.01, 0.01]$ m/s, $\delta_{cn} \in [-0.1684,\ 0.1684]$ m, $V_0 = 0.001 \times U \in R^{16\times 5}$, $W_0 = 0.001 \times [1\ \ \cdots\ \ 1]^{\mathrm{T}} \in R^{6\times 11}$, $A_c = \begin{bmatrix} -0.001 & 0 \\ 0 & -0.001 \end{bmatrix}$, $B_c = [100\ \ \ 10000]^{\mathrm{T}}$, $\Gamma_V = I_{16}$, $Q_w = [1\ \ 10\ \ 100\ \ 0.1\ \ 0.01]$. U is a matrix with all entries 1; $Q = I_{10}$, δ_{cnv}, δ_{cn} are rate limit and position limit of the saturation functions, respectively. With these choices, the stable matrices A_{cl} and $\bar{A}$ occur. The obtained stabilizing compensator is:

$$\tilde{K}_0 = 10^7 \times [4.6817\ \ \ 0.0109\ \ \ 0.00001473\ \ \ 0\ \ \ 0]^{\mathrm{T}}, K_1 = 10^5 \times [-10\ \ \ -10]^{\mathrm{T}},$$

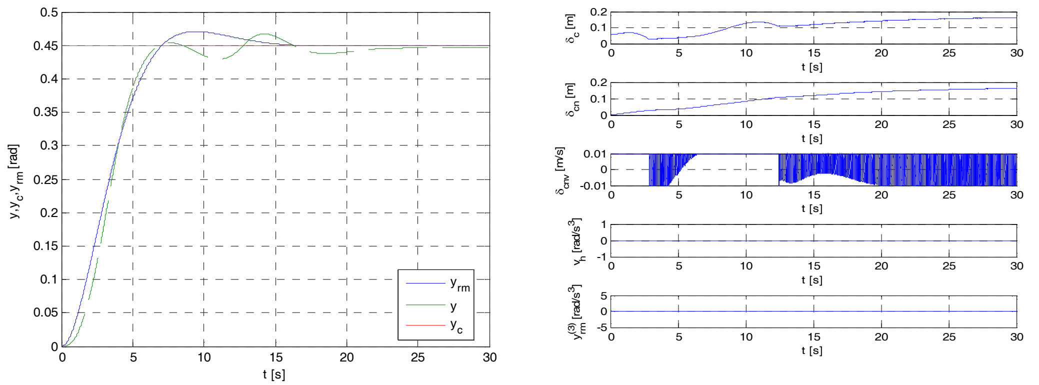
$$K_f = [2\ \ 2\ \ 1\ \ -100\ \ -10000]^{\mathrm{T}}$$



a) b)

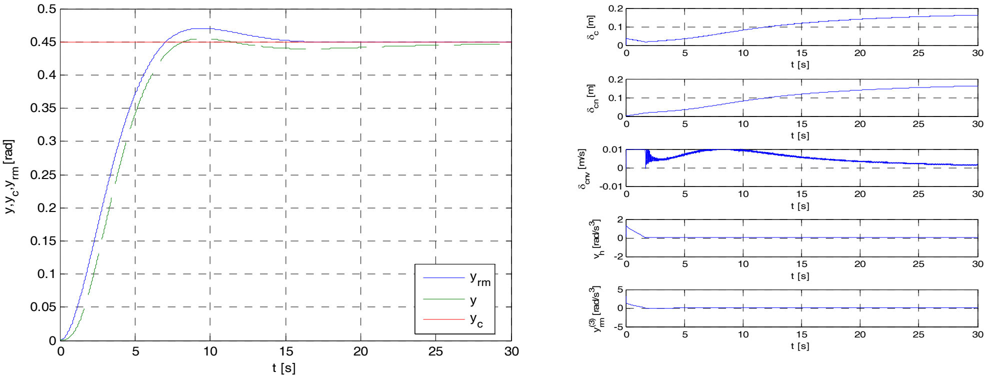
Fig. 5. The case of the system with time delay and counteracting PCH. a) Time history of controlled output y versus reference model y_{rm} representing the filtered input signal y_c. b) Time histories of control variables: control variable before saturations; δ_{cn} – control variable after saturations; $\delta_{cnv} := \dot{\delta}_{cn}$; ν_h – estimated achievable pseudo control; $y_{rm}^{(3)} := \dddot{y}_{rm}$

Fig. 5a) displays the tracking performance of the system (1) with the PCH signal counteracting harmful effects of a time delay $T_d = 0.005$ s representing, for example, a computation delay; histories of control variables are shown in Fig 5b).



a) b)

Fig. 6. The case of the system with time delay and both position and rate saturations, without ounteracting PCH. a) Time history of controlled output y versus reference model y_{rm} representing the filtered input signal y_c. b) Time histories of control variables: δ_c – control variable before saturations; δ_{cn} – control variable after saturations; $\delta_{cnv} := \dot{\delta}_{cn}$; ν_h – estimated achievable pseudo control; $y_{rm}^{(3)} := \dddot{y}_{rm}$. Last two variables are herein automatically canceled



a) b)

Fig. 7. The case of the system with time delay and both position and rate saturations, with counteracting PCH − let compare with both Figs. 5, 6. a) Time history of controlled output y versus reference model y_{rm} representing the filtered input signal y_c. b) Time histories of control variables: δ_c − control variable before saturations; δ_{cn} − control variable after saturations; $\delta_{cnv} := \dot{\delta}_{cn}$; ν_h − estimated achievable pseudo control; $y_{rm}^{(3)} := \dddot{y}_{rm}$.

Figs. 6 and 7 displays the tracking performance without and with PCH compensation of saturations. In Fig. 7, it can be seen that the PCH compensator removes the oscillation which was present in the response in Fig. 6. This exhibits that the proposed controller, improved with an antisaturation mechanism, which interacts with the actuator dynamics, achieves good tracking performance for a system with advanced level of uncertainty. From Fig. 7 it is evident that the control position and rate limits are partially diminished during most of the response, as a consequence of the PCH influence. Thus, the synthesized adaptive output feedback controller, augmented with the PCH compensator, ensures a superior tracking performance for uncertain systems, having a known relative degree.

A further improving of tracking properties y versus y_{rm} can be still reached by the *trail and error* procedure.

4. CONCLUSIONS

Output feedback adaptive control is a promising approach in order to achieve good tracking performances for uncertain systems. This approach relies on accurately accounting for the relative degree for the system, but does not require an accurate model for the plant dynamics. Achievable performance is ultimately limited by the actuator performance, but the effects of the actuator limits and delays can be considerably reduced by employing an anti-saturation compensator; herein, a pseudo-control hedging was considered, but other procedures are available, see for example [5], [6], [20].

But the main conclusion of this paper in two parts refers to the proposal and validation, by numerical simulations, of a new unitary conceived pseudo control, obtained in an uncertainty framework. The components of this so called pseudo control, thought on a superposition effects principle, are the following: 1) the output of the reference model, 2) the output of a Kalman type stabilizing compensator of the pair of systems composed by a) an output dynamics of a set of integrators of order tantamount to the assumed known relative degree r of the controlled system and b) an internal model, of order $r-1$, oriented to the tracking error decreasing in the presence of step input signals, and 3) the adaptive control designed to approximately cancel the error of approximate dynamic inversion by virtue of whom the real control is hereby determined from the pseudo control. A single hidden layer neural network is used to counteract this dynamic inversion error. The common approach of the pseudo control design based on tracking error dynamics estimation [10]-[13] is evaded.

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