Computation of Stabilizing PI and PID parameters for multivariable system with time delays

Nour El Houda Mansour, Sami Hafsi, Kaouther Laabidi

Laboratoire d'Analyse, Conception et Commande des Systèmes
Université de Tunis El Manar
BP 37, Le Belvedère 1002, Tunis, Tunisia
nourelhoudamansour@gmail.com
samihafsi@ymail.com
Kaouther.Laabidi@enit.rnu.tn

Abstract—**In this paper, a new approach to multivariable Smith control design is proposed for the computation of all stabilizing by Proportional-Integra (PI) and Proportional-Integral-derivative PID controllers for Multi-Inputs Multi-Outputs (MIMO) plants with multiple time delays. First, the Smith decoupler is used to eliminate interactions. Second the decoupled models are approximated by second-order plus time delay (SOPDT) models using the standard recursive least square approach. We seek finally, by using a generalization of Hermite-Biehler theorem, the set of complete stabilizing PI and PID parameters.**

Keywords— **MIMO with multiple time delays plants, SOPDT model, PID controller, Smith control, Hermite-Biehler theorem.**

I. INTRODUCTION

Since the pioneering research woks of the early sixties (see for example HOROWITZ, SKOGESTAD and POLETHWAITE [1, 2]), the synthesis of controllers for multivariable delayed processes has received more attention in the industrial field and more particularly in the domain of chemical engineering [3]. It is still more practical to control multivariable processes by using a SISO (Single-Input Single-Output) control structure (to avoid interactions between different loops), that is, decentralized control.

Despite the significant developments made in advanced control theory, PID controllers are still the most commonly adopted in industries [4, 5]. The main reasons for the popularity of such controllers is that their simple structure leads to easy and rapid designs. Based on these advantages, PID controllers seem to be an interesting choice for MIMO plants (require fewer parameters to adjust) [6].

Effective control of multivariable system with time delays processes is a difficult issue in the context of process control [2]. Input–output loops in a multivariable plant usually have different time delays, and for a particular loop its output could be affected by all the inputs through different time delays. Such a plant can be represented by a multivariable transfer function matrix with multiple time delays around an operating point. The researches of control method for this kind of multivariable plants with multiple time delays have received considerable attention [7].

In the case of multivariable coupling system, the controller was designed for each sub-loop by analyzing the dominant pole and amplitude ratio by LUYBEN [8], JIETAE and THOMAD [9]. This method was used to obtain a large number of controller designs, but the controller regulating of one loop affected the performance of the system, even the other loops. Decoupling control is an effective method for multivariable system to eliminate the interactions between the sub-loops. By decomposing the system to independent sub-loops using a decoupler, a decentralized PID controller is designed. The great advantage of this method is that it permits the use of single-input single-output (SISO) controller design. ASTROM et al and SAEED et al in [10, 11] introduced static decoupler and dynamic decoupler at the input port to construct unit feedback closed-loop decoupling control. But the above method can be used only in two-input two-output (TITO) process.

In this paper, a control design procedure to deal with the Smith predictor controller design is presented for decoupling and stabilizing of MIMO process with multiple time delays. Decoupling Smith control design is adopted in this work. With the help of the standard recursive least square approach the decoupled single-loop models are reduced to SOPDT models.

Based on these identified SOPDT models we will provide a complete analytical solution to the problem which is based on an extension of the Hermite–Biehler theorem. The advantages of the proposed methods are clearly presented in the numerical example.

The remainder of this paper is organized as follows. In the next section the use of the smith control's decoupler is discussed. In section 3 the standard recursive least square approach was proposed to approximate the decoupled system by SOPDT models. We present in section 4 and 5 the Hermite–Biehler theorem that is useful for PI and PID controllers synthesis. In section 6 the proposed method is applied to the Wood & Berry binary distillation column plant [7] to show its effectiveness and applicability. Finally conclusion remarks are made.

II. SMITH PREDICTOR FOR MIMO SYSTEMS WITH DELAYS

For a SISO delayed, Smith proposed a compensation scheme which can eliminate the delay of the characteristic equation closed loop which facilitates its control and greatly improves its impulse response [12]. This type of control can be applied to MIMO systems with multiple delays. To eliminate the coupling between the various loops, we will then present the design of decoupling control Smith proposed by Wang et al [13]. Therefore, the resulting decoupled system will be reduced to a SOPDT plant which seems to be an useful tool for solving the problem of control of MIMO systems with multiple delays.

The structure of multivariable control by Smith predictor [13] is illustrated in Fig. 1:

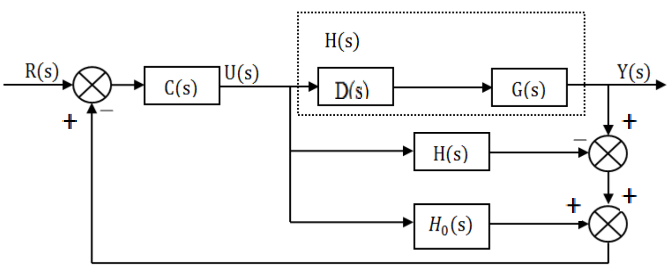

Fig.1 Multivariable Smith predictor structure.

With R, Y, C, D and G represent respectively the input, output, controller, decoupler, and the process which is a stable nonsingular matrix (det(G(0)≠0)). And $H_0(s)$ is the same as H(s) except with no delay.

Consider the multivariable system with the transfer matrix:

$$G(s) = \begin{bmatrix} g_{11}(s) & \dots & g_{1n}(s) \\ \vdots & \ddots & \vdots \\ g_{n1}(s) & \dots & g_{nn}(s) \end{bmatrix} \quad (1)$$

Where $g_{ij}(s) = g_{ij0}(s)\ e^{-\tau_{ij}s}$.

The transfer function of the closed loop system is given by:

$$\Phi(s) = \frac{H(s)C(s)}{1+H_0(s)C(s)} \quad (2)$$

In the case of multivariable control by Smith Predictor, the decoupler D (s) is determined in a way that the matrix of the transfer function H(s) is diagonal and is expressed as follows:

$$D(s) = \frac{K_D(s)K'(s)}{\det(G(0))} \quad (3)$$

Where

$$K' = \mathrm{adj}(G)$$

$$K_D(s) = \mathrm{diag}\,(\exp(\tau_{ri}s));\ i = 1, \dots, n$$

And τ_{ri} is the smallest time delay in each column of K'. The decoupled process can be then written as follows:

$$H(s) = G(s)D(s) = \mathrm{diag}(h_1, h_2, \dots\dots, h_n) \quad (4)$$

III. MODEL REDUCTION

After decoupling a MIMO system interactions are eliminated but the resulting decoupled process given by equation (4) is complex. In order to simplify the system and ensure an efficient control, the model is approximated by a SOPDT plant in this work.

In order to find an approximation by a SOPDT model for $h_i(s)$, these unknown parameters, namely K_i, τ_{di} , $a_{1i}$and a_{0i} (i=1, 2, …, n) should be determined. The standard recursive least square approach [14] is adopted here. The resulting L (s) then obtained can be written as follows:

$$L(s) = \mathrm{diag}\,(l_1, l_2, \dots, l_n)$$

$$L(s) = \mathrm{diag}\left(\frac{K_1\ e^{-\tau_{d1}s}}{s^2 + a_{11}\,s + a_{01}}, \frac{K_2\ e^{-\tau_{d2}s}}{s^2 + a_{12}\,s + a_{02}}, \dots, \frac{K_n\ e^{-\tau_{dn}s}}{s^2 + a_{1n}\,s + a_{0n}}\right) \quad \mathbf{(5)}$$

IV. PRELIMINARY RESULTS FOR ANALYZING TIME DELAY SYSTEM

Several problems in process control engineering are related to the presence of delays. These delays intervene in dynamic models whose characteristic equations are of the following form [5, 25]:

$$\delta(s)=d(s)+ e^{-L_1 s}n_1(s)+ e^{-L_2 s}n_2(s)+\dots+\ e^{-L_m s}n_m(s) \quad (6)$$

Where: d(s) and $n_i(s)$ are polynomials with real coefficients and L_i represent time delays. These characteristic equations are recognized as quasi-polynomials. Under the following assumptions:

(A_1) Deg (d(s))= n and deg ($n_i(s)$) < n for i=1,2,…,m

(A_2) $L_1 < L_2 < \dots < L_m$ (7)

One can consider the quasi-polynomials $\delta^*(s)$ described by:

$$\delta^*(s)=\ e^{L_m s}\ \delta(s)$$

$$\delta^*(s) =\ e^{L_m s}\ d(s) +\ e^{s(L_m-L_1)}\ n_1(s) +\ e^{s(L_m-L_2)}\ n_2(s) + \dots +\ n_m(s) \quad (8)$$

The zeros of $\delta(s)$ are identical to those of $\delta^*(s)$ since $e^{L_m s}$ does not have any finite zeros in the complex plan. However, the quasi-polynomial $\delta^*(s)$ has a principal term since the coefficient of the term containing the highest powers of s and e^s is nonzero. If $\delta^*(s)$ does not have a principal term, then it has an infinity roots with positive real parts [5].

The stability of the system with characteristic equation (6) is equivalent to the condition that all the zeros of $\delta^*(s)$ must be in the open left half of the complex plan. We said that $\delta^*(s)$ is Hurwitz or is stable. The following theorem gives a necessary and sufficient condition for the stability of $\delta^*(s)$.

Theorem 1 (Hermite–Biehler) [5]

Let $\delta^*(s)$ be given by (8), and write:

$$\delta^*(j\omega) = \delta_r(\omega) + j\delta_i(\omega) \tag{9}$$

where

$\delta_r(\omega)$ and $\delta_i(\omega)$ represent respectively the real and imaginary parts of $\delta^*(j\omega)$. Under conditions (A_1) and (A_2), $\delta^*(s)$ is stable if and only if:

1: $\delta_r(\omega)$ and $\delta_i(\omega)$ have only simple, real roots and these interlace,

2: $\delta_i'(\omega_0)\,\delta_r(\omega_0) - \delta_i(\omega_0)\,\delta_r'(\omega_0) > 0$ for some ω_0 in $[-\infty,+\infty]$.

Where $\delta_r'(\omega)$ and $\delta_i'(\omega)$ denote the first derivative with respect to ω of $\delta_r(\omega)$ and $\delta_i(\omega)$, respectively.

A crucial stage in the application of the precedent theorem is to verify that and have only real roots. Such a property can be checked while using the following theorem.

Theorem 2 [5]

Let M and N designate the highest powers of s and e^s which appear in $\delta^*(s)$. Let η be an appropriate constant such that the coefficient of terms of highest degree in $\delta_r(\omega)$ and $\delta_i(\omega)$ do not vanish at $\omega = \eta$. Then a necessary and sufficient condition that $\delta_r(\omega)$ and $\delta_i(\omega)$ have only real roots is that in each of the intervals $-2l\pi+\eta < \omega < 2l\pi+\eta$, $l = l_0, l_0+1, l_0+2...$ $\delta_r(\omega)$ or $\delta_i(\omega)$ have exactly $4lN + M$ real roots for a sufficiently large l_0.

V. PI CONTROL FOR SECOND ORDER DELAY SYSTEM

A second order system with delay can be mathematically expressed by a transfer function having the following form:

$$G(s) = \frac{K}{s^2+a_1 s+a_0} e^{-Ls} \tag{10}$$

Where K is the static gain of the plant, L is the time delay and a_0 and a_1 are the plant parameters. The parameters are always positive.

In this section, the stabilization of the plant is assured by the PI controller designed as follow:

$$C(s) = K_p + \frac{K_i}{s} \tag{11}$$

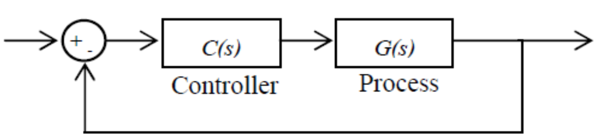

Fig.2 The closed-loop system with controller

The proposed method leads to an efficient calculation of the proportional and integral gains K_p and K_i achieving stability.

The characteristic equation of the closed-loop system is given by:

$$\delta(s) = K(K_i + K_p s)\, e^{-Ls} + (s^2 + a_1 s + a_0)s \tag{12}$$

We deduce the quasi-polynomial:

$$\delta^*(s) = e^{Ls}\delta(s) = K(K_i + K_p s) + s(s^2 + a_1 s + a_0)e^{Ls} \tag{13}$$

by replacing s by $j\omega$, we get:

$$\delta^*(j\omega) = \delta_r(\omega) + j\delta_i(\omega)$$

With:

$$\begin{cases} \delta_r(\omega) = K K_i + (\omega^3 - a_0\omega)\sin(L\omega) - a_1\omega^2\cos(L\omega) \\ \delta_i(\omega) = \omega[K K_p + (a_0 - \omega^2)\cos(L\omega) - a_1\omega\sin(L\omega)] \end{cases}; \tag{14}$$

Clearly, the parameter K_i only affects the real part of $\delta^*(j\omega)$ whereas the parameter K_p affects the imaginary part.

Let's put $z = L\omega$, we get:

$$\begin{cases} \delta_r(z) = K K_i + \sin(z)\left(\frac{z^3}{L^3} - a_0\frac{z}{L}\right) - a_1\frac{z^2}{L^2}\cos(z) \\ \delta_i(z) = \frac{z}{L}[K K_p + \cos(z)\,(a_0 - \frac{z^2}{L^2}) - a_1\frac{z}{L}\sin(z)] \end{cases}; \tag{15}$$

From the previous expressions, the real part $\delta_r(z)$ depends on the controller parameter K_i whereas the imaginary part $\delta_i(z)$ depends on K_p. In order to determine the range of K_p, Silva et al. (2005) and Farkh et al. (2009a) use the following theorem [17, 18]:

$$-\frac{a_0}{K} < K_p < \frac{1}{K}\left(a_1\frac{\alpha}{L}\sin(\alpha) - \cos(\alpha)\left(a_0 - \frac{\alpha^2}{L^2}\right)\right); \tag{16}$$

Where α is the solution of the equation $\tan(\alpha) = \frac{\alpha(2+a_1 L)}{(\alpha^2 - a_1 L - a_0 L^2)}$ in the interval $[0,\pi]$.

For K_p values outside this range, there are no stabilizing PID controllers.

All of the values of the parameter K_p given in (16) verify the first condition of Hermite–Biehler theorem, which required that the roots of $\delta_r(z)$ and $\delta_i(z)$ are simple and interlaced. Applying the condition, $\delta_i'(z)\delta_r(z) - \delta_i(z)\delta_r'(z) > 0$, we can compute the set of parameter controller K_i gains that verifies the interlace property of the roots.

Thus we can rewrite $\delta_r(z)$ as follows:

$$\begin{aligned} \delta_r(z) &= K K_i + \sin(z)\left(\frac{z^3}{L^3} - a_0\frac{z}{L}\right) - a_1\frac{z^2}{L^2}\cos(z) \\ &= K[K_i - a(z)] \end{aligned} \tag{17}$$

Where

$$a(z) = \frac{z}{KL}\left[\sin(z)\left(a_0 - \frac{z^2}{L^2}\right) + a_1\frac{z}{L}\cos(z)\right] \tag{18}$$

Let's put z_j, $j = 1, 2, 3...$ the roots of $\delta_r(z)$ and $a(z_j) = a_j$. Interlacing the roots of $\delta_r(z)$ and $\delta_i(z)$ is equivalent to $\delta_r(z_0) > 0$ (since $K_i > 0$), $\delta_r(z_1) < 0$, $\delta_r(z_2) > 0$... We can use the interlacing property and the fact that and $\delta_i(z)$ has only real roots to establish that $\delta_r(z)$ possess real roots too. From the previous equations we get the following inequalities:

$$\begin{cases}\delta_r(z_0) > 0\\ \delta_r(z_1) < 0\\ \delta_r(z_2) > 0\\ \delta_r(z_3) < 0\\ \delta_r(z_4) > 0\\ \vdots\end{cases} \Rightarrow \begin{cases}K_i > 0\\ K_i < a_1\\ K_i > a_2\\ K_i < a_3\\ K_i > a_4\\ \vdots\end{cases}; \qquad (19)$$

From these inequalities, it is clear that the a_j odd bounds must be strictly positive; however the a_j even bounds are negative in order to find a feasible range of K_i. From which we have:

$$0 < K_i < \min_{j=1,3,5,\ldots}\{a_j\} \qquad (20)$$

Algorithm for determining PI parameters [17]

1) Choose K_p in the interval suggested in (16) and initialize j = 1,
2) find the roots z_j of $\delta_i(z)$,
3) Compute the parameter a_j associated with the z_j previously founded,
4) Determine the lower and the upper bounds for K_i as follows:

$$0 < K_i < \min_{j=1,3,5,\ldots}\{a_j\}$$

5) Go to step 1.

VI. PID CONTROL FOR SECOND ORDER DELAY SYSTEM

Considering the same system expressed by equation (10) shown in Fig.2, we attempt to achieve stabilization with PID controller presented by:

$$C(s) = K_p + \frac{K_i}{s} + K_d s \qquad (21)$$

The characteristic equation of the closed-loop system is given by:

$$\delta(s)=K(K_i + K_p s+K_d s^2)\, e^{-Ls}+ (s^2 + a_1 s + a_0)s \quad (22)$$

we deduce the quasi-polynomial:

$$\delta^*(s) =K(K_i + K_p s+K_d s^2)+s(s^2 + a_1 s + a_0)e^{Ls} \quad (23)$$

by replacing s by jω , we get:

$$\delta^*(j\omega) = \delta_r(\omega) + j\delta_i(\omega)$$

Where

$$\begin{cases}\delta_r(\omega) = K K_i - K K_d + (\omega^3 - a_0\omega)\sin(L\omega) - a_1\omega^2\cos(L\omega)\\ \delta_i(\omega) = \omega\left[K K_p + (a_0 - \omega^2)\cos(L\omega) - a_1\omega\sin(L\omega)\right]\end{cases} \quad (24)$$

And by putting z = Lω, we get:

$$\begin{cases}\delta_r(z) = K K_i - K K_d \frac{z^2}{L^2} + \sin(z)\left(\frac{z^3}{L^3} - a_0\frac{z}{L}\right) - a_1\frac{z^2}{L^2}\cos(z)\\ \delta_i(z) = \frac{z}{L}\left[K K_p + \cos(z)\left(a_0 - \frac{z^2}{L^2}\right) - a_1\frac{z}{L}\sin(z)\right]\end{cases}; \quad (25)$$

We notice that $\delta_i(z)$ have the same expression as in (15) then to determine the range of K_p we will use the same theorem.
All of the values of the parameter K_p given in (16) verify the first condition of Hermite–Biehler theorem, which required that the roots of $\delta_r(z)$ and $\delta_i(z)$ are simple and interlaced. Applying the condition, $\delta_i'(z)\,\delta_r(z) - \delta_i(z)\,\delta_r'(z) > 0$, we can compute the set of K_i and K_d gains that verifies the interlace property of the roots.
Thus we can rewrite $\delta_r(z)$ as follows:

$$\delta_r(z) = K\frac{z^2}{L^2}\left[-K_d + m(z)\,K_i + b(z)\right] \qquad (26)$$

Where

$$\begin{cases}m(z) = \frac{z^2}{L^2}\\ b(z) = \frac{L}{Kz}\left[-a_{11}\frac{z}{L}\cos(z) + \sin(z)\left(\frac{z^2}{L^2} - a_{00}\right)\right]\end{cases}; \quad (27)$$

Let's put z_j, j = 1, 2, 3... the roots of $\delta_r(z)$, $m(z_j) = m_j$ and $b(z_j) = b_j$.
By using the interlacing property and the fact that and $\delta_i(z)$ has only real roots to establish that $\delta_r(z)$ possess real roots too we get the following inequalities:

$$\begin{cases}\delta_r(z_0) > 0\\ \delta_r(z_1) < 0\\ \delta_r(z_2) > 0\\ \delta_r(z_3) < 0\\ \delta_r(z_4) > 0\\ \vdots\end{cases} \Rightarrow \begin{cases}K_i > 0\\ K_d < m_1K_i + b_1\\ K_d > m_2K_i + b_2\\ K_d < m_3K_i + b_3\\ K_d > m_4K_i + b_4\\ \vdots\end{cases}; \qquad (28)$$

In order to determine the cross-section of the stabilizing region in the (K_i, K_d) space for each value of the parameter K_p given in (16), the theorem of Farkh et al. [19] is used. It is defined as:
1) A trapezoid T if $b_3 > b_1$,
2) A triangle Δ if $b_1 > b_3$ [19].

Algorithm for determining PID parameters [19]

1) Choose K_p in the interval suggested in (16) and initialize j = 1,
2) Find the roots z_j of $\delta_i(z)$,
3) Compute the parameter b_j, m_j associated with the z_j for j = 1, 2, 3 founded,
4) Determine the stability region in the plane (K_i, K_d) using Farkh et al. Theorem [19].
5) Go to step 1.

VII. SIMULATION RESULTS

To show the effectiveness of the proposed design methods, we consider the following binary distillation column:

$$G(s) = \begin{bmatrix} \frac{12.8e^{-s}}{16.7s+1} & \frac{-18.9e^{-3s}}{21s+1} \\ \frac{6.6e^{-7s}}{10.9s+1} & \frac{-19.4e^{-s}}{14.4s+1} \end{bmatrix}$$

The decoupler is designed using Eq. (3):

$$D(s) = \begin{bmatrix} \frac{0.157}{14.4s+1} & \frac{-0.153e^{-2s}}{21s+1} \\ \frac{0.053e^{-4s}}{10.9s+1} & \frac{-0.104}{16.7s+1} \end{bmatrix}$$

The resulting diagonal decoupled system is:

$$H(s) = \mathrm{diag}\big(h_1(s), h_2(s)\big)$$

$$\begin{cases} h_1(s) = \frac{2.01e^{-s}}{(16.7s+1)(14.4s+1)} - \frac{1.01e^{-7s}}{(21s+1)(10.9s+1)} \\ h_2(s) = -\frac{1.01e^{-9s}}{(21s+1)(10.9s+1)} + \frac{2.01e^{-3s}}{(16.7s+1)(14.4s+1)} \end{cases}$$

In order to determine PI and PID controllers, $h_1(s)$ and $h_2(s)$ should be expressed as SOPDT processes. Using the standard recursive least square, SOPDT models for $h_1(s)$ and $h_2(s)$ are determined as follow:

$$\begin{cases} l_1(s) = \frac{0.004125\ e^{-13.11s}}{s^2 + 0.13\,s + 0.004} \\ l_2(s) = \frac{0.04\ e^{-29.59s}}{s^2 + 0.127\,s + 0.004} \end{cases}$$

In order to determine K_p values stabilizing $l_1(s)$ and $l_2(s)$, we look for α in interval [0, π] satisfying Eq. (16).we find respectively the following results:

$\tan(\alpha_1) = \frac{3.7043\alpha_1}{\alpha_1{}^2 - 2.3918}$ → α_1 = 1.562 → $K_p$range is given by: $-0.97 < K_p < 3.776$.

$\tan(\alpha_2) = \frac{5.75\alpha_2}{\alpha_2{}^2 - 7.26}$ → α_2= 1.89 → $K_p$range is given by: $-1 < K_p < 1.919$.

The systems stability regions, obtained in (K_p, K_i) plane are presented in Fig.3 and Fig.4.

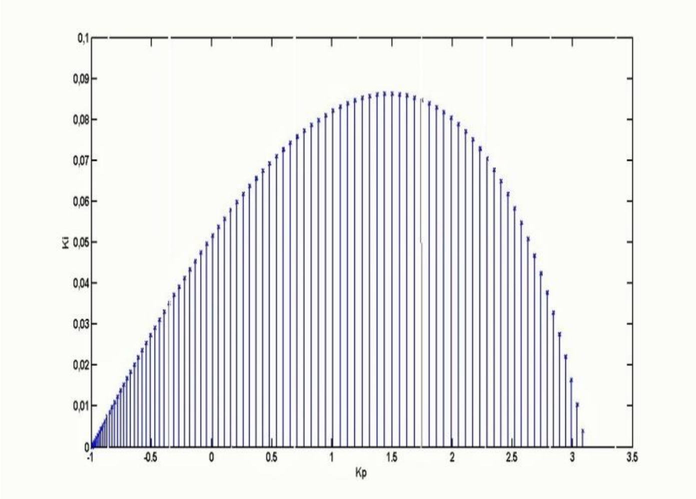

Fig.3 PI controller stability region for l_1

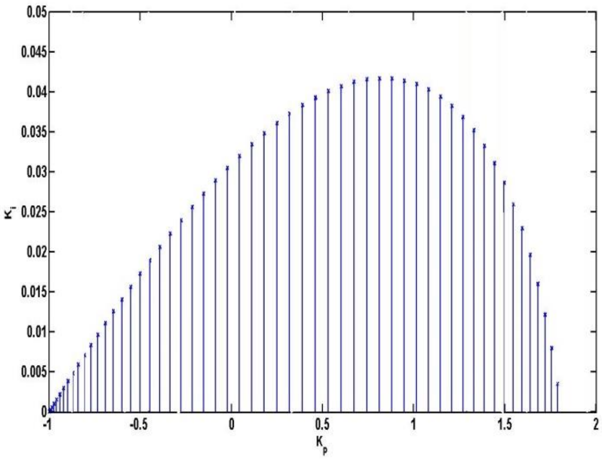

Fig.4 PI controller stability region for l_2

By sweeping over K_p, a stability region is defined in the (K_i, K_d) plane. A three dimensional curve is then obtained as shown in Fig.5 and Fig.6.

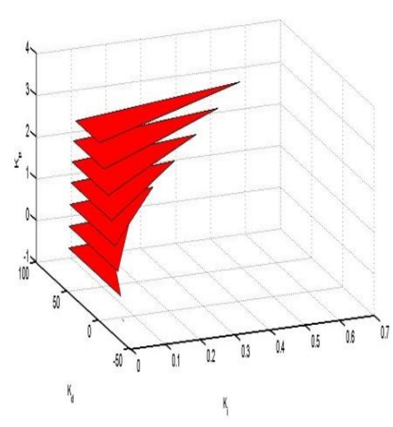

Fig.5 PID controller stability domain for l_1

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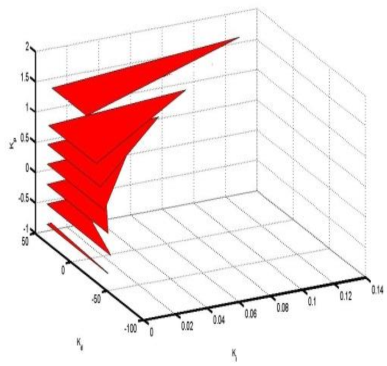

Fig.6 PID controller stability domain for l_2

By using Fig. 3 and Fig. 4, we start with the following PI controllers:

$$C_{PI}(s) = \begin{bmatrix} 0.408 + \dfrac{0.020581}{s} & 0 \\ 0 & 1.5 + \dfrac{0.03}{s} \end{bmatrix}$$

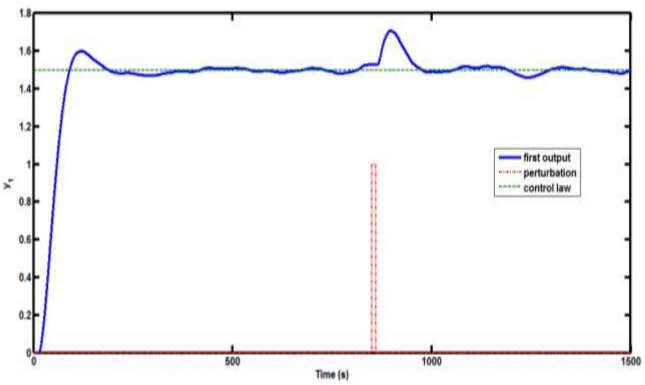

Fig.7 First output Step responses with PI Controller (K_p = 0.40814 and K_i= 0.020581).

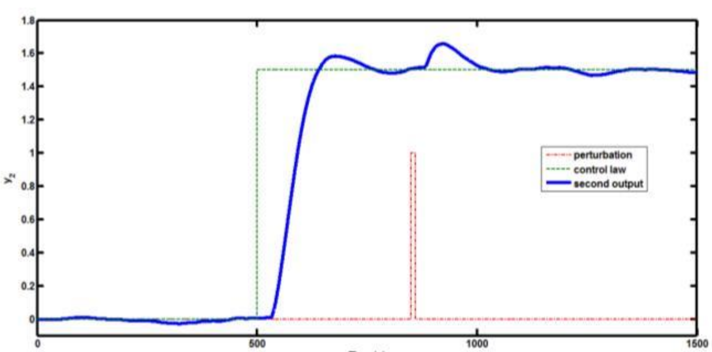

Fig.8 Second output Step responses with PI Controller (K_p = 1.5 and K_i= 0.03).

Output responses to unit step function in the first input r_1 and the second input r_2 are shown in Fig. 7 and Fig. 8 for the simulation, an unit step disturbance has been applied to the process input at t=850 s. It is clearly seen that perturbations of the process static gains do not affect decoupling regulation of the output responses.

By using Fig. 5 and Fig. 6, we choose with the following PID controllers:

$$C_{PID}(s) = \begin{bmatrix} 0.80822 + \dfrac{0.022993}{s} - 5.8542\, s & 0 \\ 0 & 0.12 + \dfrac{0.010856}{s} \end{bmatrix}$$

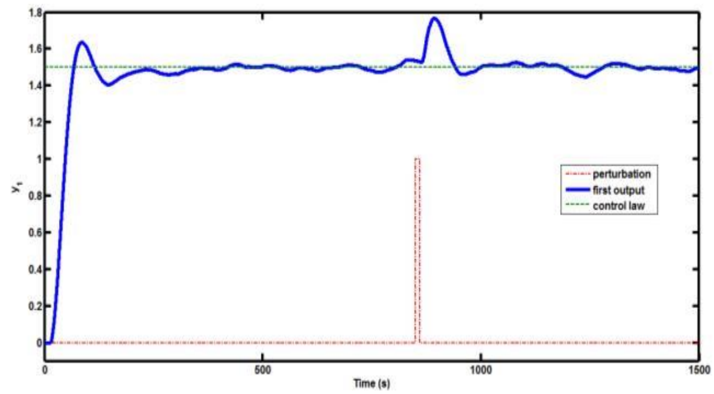

Fig.9 First output Step responses with PID Controller (K_p = 0. 80822 ; K_i= 0.022993 and K_d=-5.8542).

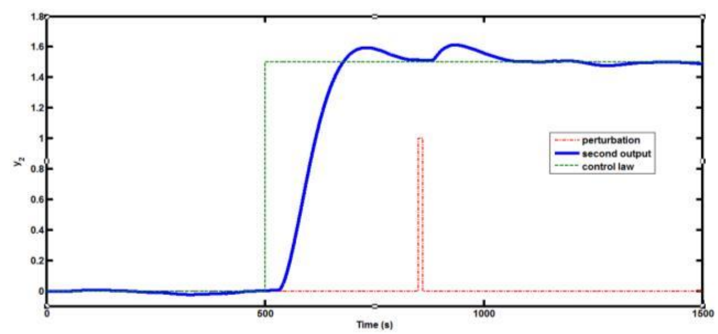

Fig.10 Second output Step responses with PID Controller (K_p = 0.12; K_i= 0.010856 and K_d=0).

A unit step disturbance has been applied to the process input at t=850s. The simulation results shown in Fig. 9 and Fig. 10 reveal that the proposed control strategy gives robust system performance.

The obtained results show the effectiveness of the proposed approach.

VIII. CONCLUSION

This paper proposed a simple method to tune decentralized PID controllers for MIMO plants with multiple delays. The MIMO was first decoupled using the design of decoupling Smith control. A model reduction was proposed to find a suitable (SOPDT) model for each element of the resulting diagonal process. An extension of Hermit-Biehler theorem is used to find stabilizing PI and PID sets for the reduced system This approach is finally investigated through its application to the Wood & Berry binary distillation column plant.

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