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Automatic Tuning of Cascade Structure CNC Controllers

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Index terms : Automatic tuning, Cascade structure PI and P controllers.

Abstract : *This paper proposes a method for the automatic tuning of the cascade structure CNC controllers. The design takes into account constraints on the system such as delays induced by numerical control effects and elasticity of bodies. The robustness of the algorithm is ensured by requiring minimal stability margins.*

1. Introduction

The controllers of industrial CNC are usually based on cascade structure with a fast inner loop for torque control and outer loops for speed and position control. This structure improves the robustness of the controller with respect to the disturbances and modeling errors. The speed control loop is classically implemented with proportionnal and integral (PI) controllers (PI or IP structure) with anti-windup strategy. The position loop is designed with a P controller. In order to decrease the tracking errors, velocity and acceleration feedforward signals might be added to the control loops. For a given structure, the problem is to tune the controller gains considering the CNC components, the mechanical device and the drive chain structure. PI controllers have traditionally been tuned empirically e.g. by the method described by Ziegler and Nichols in [1]. In recent years, several methods have been proposed [2], [3], [4], [5], [6]. In [2], for instance, considering the emerging auto-tuners and tuning devices, the design problem is solved as an optimization problem based on sensitivity constraint and is fully adapted for an extended class of complex systems. In [5] and [6], the digital RST is designed using pole placement and/or sensitivity function shaping. The robustness indicator is based on the modulus margin or given in terms of time response and delay margin. Following recent works [7] dealing with the modeling and simulation of a machine tool axis, the objective is now to provide a complete simulation toolbox to predict the performances of the machine tool at the stage of its design and before its prototyping or manufacturing. The problem for the machine tool manufacturer is to assembly different existing components (numerical controllers, actuators,

sensors and mechanical devices) available on the market without knowing the global resulting performances. Considering the main problems of such a process that is the dynamics of rigid bodies, the elasticity of bodies and mechanical transmission, the dynamics and saturations of electric actuators, the effect of numerical control as delays and quantization, the last stage of the design is to define a procedure to tune the controllers taking into account all the previous components of the modeling. Therefore, the main contribution of this work is to propose an efficient procedure based on a frequency approach and stability margins, using a combination of very more simple but still efficient tuning methods. Basically, the digital control systems are redesigned in the continuous space. Then the gain values are tuned using a simplified rigid model. These values initialize an optimization procedure using the full model to perform an accurate tuning. The paper is organized as follows : Section 2 describes the proposed design tuning method. Section 3 briefly recalls the control tuning strategies developed in [2] and [5]. Section 4 exhibits comparative results. The simulation tests are performed on a simplified machine tools axis model.

2. The proposed design method

2.1 Digital controller design in case of a rigid model

Two basic approaches to the design of digital control systems can be considered : i) a direct discrete-time design or ii) a continuous controller redesign [8]. The second one is preferable because it allows continuous time techniques to be used, specially the design in the frequency domain which appears as the most convenient way to observe stability margin indicator. The proposed tuning procedure is based on the second one and is composed of two steps. Gain values are first tuned using a rigid model. More precisely a continuous time control law is first defined. Secondly it is discretized and finally it is approximated by continuous transfer function. Therefore the gains can be calculated with a continuous controller taking into account the effect of

digital control in the bandwidth of the closed loop. At the end, these values are introduced to start an optimization procedure using a complete model considering the mechanical flexibility to perform an accurate tuning.

2.1.1 Continuous time case

As a first approximation, the velocity loop and the rigid model can be represented as given on Figure 1 :

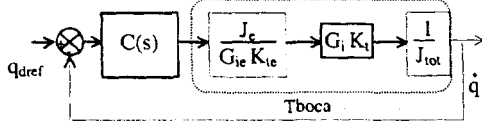


Figure 1: Velocity loop and rigid model

J_{tot} is the total inertia moment of the rigid model (motor, mechanical transmission, load),

G_i is the gain of the amplifier,

K_t is the torque constant.

J_e, G_{ie}, K_{ie} are the estimated values of J_{tot}, G_i, K_t respectively and $C(s)$ the transfer function of the velocity controller. Assuming that there is no model error i.e. $J_e = J_{tot}, G_{ie} = G_i, K_{ie} = K_t$, a general continuous model without any physical parameters (Figure 2) may be used for the tuning of the velocity and position controllers :

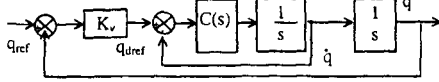


Figure 2 : Controllers with the rigid model

2.1.2 Numerical implementation

Among a few methods of discretizing continuous-time controllers, (Tustin's approximation, matched pole-zero, ...), the most popular in industry is to approximate the derivative by backward difference and the integral by a forward Euler rule. The gains of numeric elements are calculated using classical continuous approximations of the discrete transfer functions. Details of their computations are given in [7]. In the following only the PI structure tuning is developed. The digital control scheme is $-3dB$ approximated by the continuous one in the range of the Nyquist frequency $f_c/2$. Figure 3 shows the continuous digital control redesign :

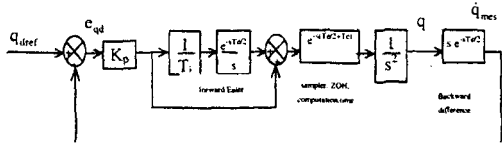


Figure 3: Continuous model of the digital controller

T_c is the computation delay.

$T_e = htr_{vel}$ is the sampling period of the velocity loop.

Time delays in this control scheme introduce irrational transfer function. But it is suitable to perform the calculation of the control gains in the frequency domain, using stability margins, without the need of using rational approximations as Padé approximants.

2.1.3 Performance index

The performance index is defined in the frequency domain and is based on optimization techniques to guarantee the desired phase and gain margins. In that way, the speed loop controller is first calculated in order to get a phase margin φ_{mv} at a given frequency ω_{vit} . This frequency is related to the first flexible mode (f_{meca}) of the process and is usually chosen between $[f_{meca}/10 \ f_{meca}/2]$. Then the controller of the position loop is calculated in order to get a phase margin φ_{mp} . Adding some test on the gain margin ensure satisfying stability margin.

2.1.4 Speed loop controller tuning

For a PI structure, T_i is first calculated to obtain the desired phase margin for the open loop transfer function given on Figure 3, i.e. :

$$\text{Arg}(T_{bov}(j\omega_{vit})) = -\pi + \varphi_{mv} \quad (1)$$

$$T_{bov}(s) = \frac{\dot{q}_{max}}{e_{sp}} \quad (2)$$

The gain K_p is then computed such as $|T_{bov}(j\omega_{vit})| = 1$.

It leads to the following analytic expressions :

$$\frac{1}{T_i} = \omega_{vit} \frac{\sin\left(\omega_{vit} \frac{T_e}{2}\right) \tan\left(\varphi_{mv} + \omega_{vit}(T_c + T_e)\right) + \cos\left(\omega_{vit} \frac{T_e}{2}\right)}{\tan\left(\varphi_{mv} + \omega_{vit}(T_c + T_e)\right)} \quad (3)$$

$$K_p = \frac{T_i \omega_{vit}^2}{\sqrt{\left(T_i \omega_{vit} \cos\left(\omega_{vit} \frac{T_e}{2}\right)\right)^2 + \left(1 - T_i \omega_{vit} \sin\left(\omega_{vit} \frac{T_e}{2}\right)\right)^2}} \quad (4)$$

2.1.5 Position loop controller tuning

The continuous time model of the position closed loop is :

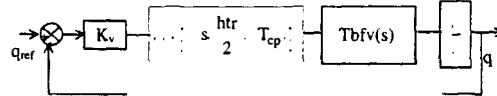


Figure 4: Position closed loop

htr is the position loop sampling time, ($htr \geq htr_{vel}$),

T_{cp} is the position loop computation delay,

$T_{bfv}(s)$ is the velocity closed loop transfer function.

K_v is calculated to ensure a phase margin φ_{mp} . Because there is only one degree of freedom in the P controller, the frequency ω_{pos} corresponding to φ_{mp} cannot be chosen independently of φ_{mp} . Indeed, it is given as the solution of the following equation :

$$\text{Arg}(T_{bfv}(j\omega_{pos})) = -\frac{\pi}{2} + \varphi_{mp} + \omega_{pos} \left(\frac{htr}{2} + T_{cp}\right) \quad (5)$$

No analytical solution exists, but a numerical one can be calculated with an optimization procedure 'fmins' [9]. Finally, K_v is computed such as :

$$\left| \frac{K_v T_{bfv}(j\omega_{pos})}{\omega_{pos}} \right| = 1 \Leftrightarrow K_v = \frac{\omega_{pos}}{|T_{bfv}(j\omega_{pos})|} \quad (6)$$

At the end of this procedure, performance indexes are checked with classical frequency techniques (Bode, Black-Nichols charts) and time analysis (step responses), with the complete model implemented in the simulator [7] developed in the environment Matlab/Simulink. If the performances are too far from the desired ones, it is possible to improve the previous controller tuning with an optimization procedure as detailed below.

2.2 Accurate tuning of the controller

Accurate tuning is performed using the full model implemented in the simulator which takes into account the lumped model of the mechanics, digital control and location of measurements.

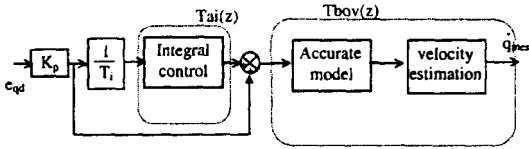


Figure 5: PI speed loop with the full model

Discrete transfer functions $T_{bov}(z)$ and $T_{ai}(z)$ (Figure 5) are calculated with the 'dlinmod' Matlab function [9]. Because of high order transfer functions, it is not possible to get analytical expressions for the gains. They are computed with optimization techniques, using 'fmin' Matlab function [9] starting with the initial values calculated with Eq. (3), (4) and (6) in order to satisfy the phase margins and bandwidth.

2.2.1 Speed loop controller

T_i is calculated by optimization such as :

$$\text{Arg}(T_{bov}(\omega_{vit})) = -\pi + \varphi_{mv} - \text{Arg}(T_{ai}(\omega_{vit})) \quad (7)$$

K_p is calculated such as :

$$K_p = \frac{1}{|T_{ai} T_{bov}|_{\omega_{vit}}} \quad (8)$$

2.2.2 Position controller

ω_{pos} is calculated such as :

$$\text{Arg}(T_{bop}(e^{-s})) = -\pi + \varphi_{mv} \quad (9)$$

K_v is calculated such as :

$$K_v = \frac{\omega_{pos}}{|T_{bfv}(e^{-s})|} \quad (10)$$

2.2.3 Gain margin tuning

In case of the complete model, the delays introduced by the CNC and the presence of multiple oscillatory modes make the control more delicate. In fact, sometimes, despite of a good phase margin, a bad stability of the complete model in closed loop results of the very small gain margin.

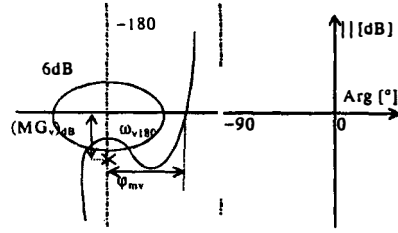


Figure 6: Speed Loop transfer obtained with an optimization based on phase margin criterion

The example Figure 6, illustrates the problem. In spite of a good phase margin, the gain margin is too small (typically 6dB gain margin is needed). The tuning procedure consists then in computing again by optimization the gain K_p if the problem is on the speed loop or the gain K_v if the problem occurs on the position loop with respect to gain margin criterion Eq. (11), (12).

i) Speed controller

$$K_p = MG_v / |T_{ai}(e^{T_s \omega_{v180}}) T_{bov}(e^{T_s \omega_{v180}})| \quad (11)$$

ω_{v180} is the critical frequency of the velocity loop such as $\text{Arg}(T_{bov}(j\omega_{v180})) = -180^\circ$, MG_v is the desired gain margin on the velocity loop.

ii) Position controller :

$$K_v = MG_p / |T_{bfv}(e^{j\pi \omega_{p180}})| \quad (12)$$

ω_{p180} is the critical frequency of the position loop ($\text{Arg}(T_{bop}(j\omega_{p180})) = -180^\circ$),

MG_p is the desired gain margin on the position loop.

3. Other control tuning strategies

Recently two procedures [2,5] have been proposed to tune controllers automatically. In this section, we will briefly describe their principles.

3.1 Robust pole placement

In [5], the robust poles placement is presented for a RST structure whose the classical scheme is given on Figure 7 :

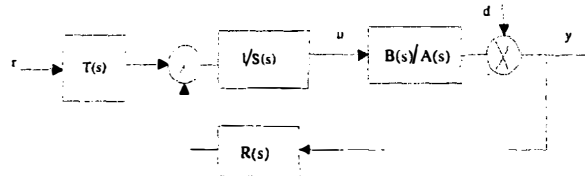


Figure 7 : RST structure

Let's define the output/disturbance sensitivity function :

$$S_{y/d}(s) = \frac{A(s)S(s)}{A(s)S(s) + B(s)R(s)} = \frac{A(s)S(s)}{P(s)} \quad (13)$$

$B(s)/A(s)$ is the process transfer function,

$$P(s) = A(s)S(s) + B(s)R(s) = C(s)F(s) \quad (14)$$

$C(s)$ and $F(s)$ are respectively interpreted as the control polynomial and the filtering polynomial.

The principle of the pole placement consists in choosing the stability polynomial $P(s)$ in order to guarantee constraint specifications on the sensitivity function. Three conditions are then required to guarantee a robust controller [5] :

1. $C(s)$ should be close to $B(s)$
2. $F(s)$ should be close to $sA(s)$ such that

$$|F(j\omega)| \geq |j\omega A(j\omega)|$$

The compromise between performances and robustness is achieved through two synthesis parameters T_c and T_0 which will be used in the poles definition of $C(s)$ and $F(s)$ in order to fulfill the three previous conditions. These parameters may be interpreted as two dominating time constants of the process. They assign the poles of the closed loop. Usually, T_c will be dedicated to the tracking performances and T_0 to the disturbance rejection. Details of the procedure may be found in [5].

3.1.1 Speed loop controller

Applying the robust pole placement technique with suitable degrees for tuning the speed loop PI controller needs the use of the first order rigid model (Figure 8).

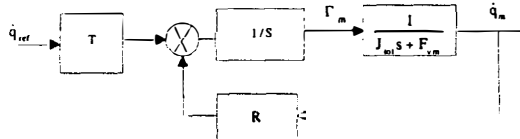


Figure 8 : Speed loop RST controller

Therefore the transfer function between the motor velocity and the torque is given by :

$$\dot{q}_m / \Gamma_m (s) = \frac{B(s)}{A(s)} = 1 / J_{tot} \left/ \left(s + \frac{F_{vm}}{J_{tot}} \right) \right. \quad (15)$$

The RST robust controller specifications give :

$$S(s) = s, \text{ degree of } R(s) = 1, C(s) = 1/J_{tot}, F(s) = \left(s + 1/T_{0s} \right)^2.$$

where T_{0s} is the desired "filtering horizon". A possible set for T_{0s} is : $1/10(J_{tot}/F_{vm}) \leq T_{0s} \leq 1/2(J_{tot}/F_{vm})$.

Eq. 14 leads to a simplified expression :

$$R(s) = \frac{F(s) - sA(s)}{1/J_{tot}} = J_{tot} \left(\frac{2}{T_{0s}} - \frac{F_{vm}}{J_{tot}} \right) s + \frac{J_{tot}}{T_{0s}^2} \quad (16)$$

Finally, the controller is written in the PI form as :

$$\frac{R(s)}{S(s)} = K_p \left(1 + \frac{1}{T_i s} \right) \quad (17)$$

$$K_p = \frac{2J_{tot}}{T_{0s}} - F_{vm} \text{ and } \frac{K_p}{T_i} = \frac{J_{tot}}{T_{0s}^2} \quad (18)$$

Several strategies may be investigated to design the pre-filter T . Hereafter, T is chosen to ensure a unitary static gain in closed loop i.e. $T(s) = R(0)$.

3.1.2 Position loop controller

In order to perform the tuning of the P position loop controller, the velocity closed loop is approximated by a first order transfer function : $1/(1+T_v s)$, where T_v is calculated as the (-45°) cut off frequency :

$$\text{Arg}(T_b f_v(j(1/T_v))) = -45^\circ$$

The P position controller is then obtained with :

$$S_p(s) = 1, R_p = K_v, C(s) = 1/T_v, F(s) = (s + 1/T_{op}) \quad (19)$$

where T_{op} is the tuning parameter of the position controller fixed by the desired time response. A possible set for T_{op} is approximately given by : $T_v/100 \leq T_{op} \leq T_v/10$. Finally, the controller can be written in the P form :

$$\frac{R_p(s)}{S_p(s)} = K_v \quad (20)$$

$$K_v = \frac{T_v}{T_{op}} - 1 \quad (21)$$

3.2 Non-Convex Optimization [2]

Let's define the process $G(s)$ and its PI controller. The loop transfer L is written as :

$$L(s) = (k + k_i/s)G(s) \quad (22)$$

The optimization problem is to find controller parameters that maximize k_i to the constraints that the closed-loop system is stable and that the Nyquist curve of the loop transfer function is outside a circle with center $s = -C$ and radius R . The design parameter M_s , which tunes the tradeoff between performances and robustness, represents the maximum of the sensitivity function $S_{yd}(s)$. It is defined :

$$M_s = \text{Max} |S_{yd}(s)| = \text{Max} \left| \frac{1}{1 + L(s)} \right| \quad (23)$$

The sensitivity constraint is then written as follows :

$$f(k, k_i, \omega) = \left| C + \left(k - i \frac{k_i}{\omega} \right) G(i\omega) \right|^2 \geq R^2 \quad (24)$$

The optimization problem is to maximize k_i subject to the sensitivity constraint (Eq. 24). An interesting advantage of this algorithm is that the optimization will compute the best solution (k_i, k) with respect to ω , given M_s , that is the best bandwidth. In spite of the very simple concept of the method, two important aspects of the procedure have to be underlined. First the problem is not trivial because the constraint defines a set in parameter space which is not convex. Moreover the initialization of the optimization algorithm play a crucial role. But for special classes of system such as systems with monotonic transfer function, a method for providing good initial conditions is given. An other point is the need to explicitly calculate the function $f(k, k_i, \omega)$ and its derivatives, which is not necessary simple.

As previously, this method is applied to a cascade structure composed of PI and P controllers. The last case leads to a simplified optimization on two parameters ω and k . The model used to perform the controller tuning is the rigid model with a delay to fulfill the required condition of monotonic decreasing phase necessary for providing good initial conditions. This delay is not restrictive at all in the context of the digital design.

4. Simulation Results

The different design procedures have been tested on a machine tool axis which is modeled by a simplified 2 masses and one spring model.

The digital elements such as delay, quantization and the dynamics and saturations of electric actuators are taken into account also.

The mechanical system is given by:

$$\Gamma_m = J_m \frac{dq_m}{dt} + F_{vm} q_m + K(Nq_m - q_l) \quad (25)$$

$$0 = J_l \frac{dq_l}{dt} + F_{vl} q_l + K(q_l - Nq_m)$$

K is the stiffness of the flexible drive joint, q_m and q_l are respectively the displacement of the motor and the load,

\dot{q}_m and \dot{q}_l are respectively the velocity of the motor and the load,

F_{vm} and F_{vl} are respectively the viscous coefficient of the motor and the load,

The simulations are carried out with the following parameters :

$$J_m = 0.0031 \text{ kg.m}^2, \quad J_l = 153.5366 \text{ kg}, \quad N = 0.0032 \text{ m}^{-1},$$

$$K = 2.7154 \text{ e}7 \text{ N/m}, \quad f_{vm} = 0.2250 \text{ Nm/rad/s}, \quad f_{vl} = 0.6800 \text{ N/(m/s)},$$

$$\text{htr} = 2 \text{ ms}, \quad \text{htr}_{vit} = 0.125 \text{ ms}, \quad T_c = 3 * \text{htr}_{vit}, \quad T_{cp} = \text{htr}.$$

A short analysis shows that the system has two oscillatory poles with a damping factor of $\xi = 0.02349$ and a natural pulsation of $\omega_o = 512.916 \text{ rad.s}^{-1}$. This pair of conjugate complex poles is classically due to the presence of the stiffness. The static behavior of the open loop is dominated by one slow mode (7.643 rad/s). Typically, this mode is introduced by viscous friction factors of the system.

Figure 9 (resp. Figure 10) represents the Nichols diagram of the velocity loop transfer function (resp. position) for each method. Figure 11 (resp. Figure 12) exhibits the step response of the velocity (resp. position) loop and a load disturbance rejection. The step magnitude is of 1 rd/s for the speed loop, 0.0011m for the position loop and for the disturbance. Figure 13 (resp. Figure 14) corresponds to the control signal of the velocity (resp. position) loop. The full line corresponds to the proposed design method. The dashed and plot line represents the robust pole placement and finally the dashed line corresponds to the method of Aström. In case of the proposed method, we choose $\phi_{mv} = 30^\circ$ at the frequency $\omega_{vit} = 256.46 \text{ rad.s}^{-1}$ depending on the first flexible

mode and $\phi_{mp} = 80^\circ$. The controller of the position loop is calculated in order to get a phase margin. The minimal gain margin of 6dB is imposed for each regulation loop.

In case of the robust pole placement, the controller tuning is applied with $T_{0s} = 1/4 (J_{tot}/F_{vm}) = 0.005$ and $T_{0p} = T_v/40$. Investigations on the choice of T_{0s} and T_{0p} must be perform

to guarantee the best regulation performances.

In case of the non convex optimization, the best results have been computed with $M_{sv} = 1.05$ and $M_{sp} = 1.3$. Greater values lead to unstable behaviors in closed loop when applied on the complete model. An alternative solution would be to perform the gains with this complete model but the need of analytic calculations make its use very prohibitive.

Table 1 exhibits the obtained speed loop stability margins :

	ϕ_{mv} (°)	ω_{vit} (rad/s)	ΔMG (dB)	ω_{cg} (rad/s)
Prop. Method	30	256.46	28.752	4537.5
De Larminat	69	263	21.3	4710
Aström	52	330	20.6	4710

Table 1 : Speed loop stability margins

Table 2 presents the resulting stability margins of the position loop :

	ϕ_{mp} (°)	ω_{pos} (rad/s)	ΔMG (dB)	ω_{cp} (rad/s)
Prop. Method	82.8	42.248	6.0	239.54
De Larminat	81.6	40.15	13.44	282.67
Aström	83.32	41.309	9.07023	295.19

Table 2 : Position loop results

Table 3 summarizes the gain tuning :

	K_p (s ⁻¹)	T_i (s)	K_v (s ⁻¹)
Prop. Method	141.685	0.00198	40.90
De Larminat	351.84	0.0088	39
Aström	378.83	0.0041	40.335

Table 3 : Gain tuning

Results obtained with all methods are similar in term of parameter values of controller and behavior in close loop (Figure 11, 12, 13, 14). Some differences appear in the velocity loop but the static gain of all the controller (K_p/T_i) is close. In spite of these results, the complexity of each technique must be also underline for a full comparison. In fact, the Aström's method is not easy to apply because some computations must be performed at first and furthermore optimization procedure is delicate to initialize.

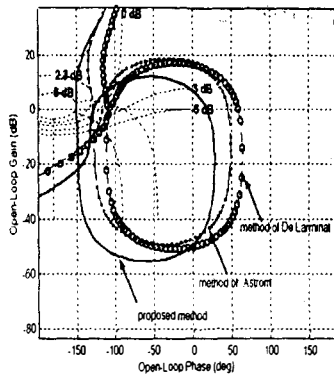


Figure 9: Speed loop Nichols diagram

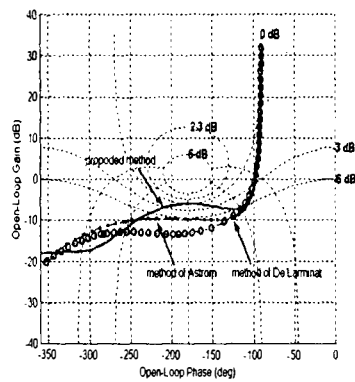


Figure 10: Position loop Nichols diagram

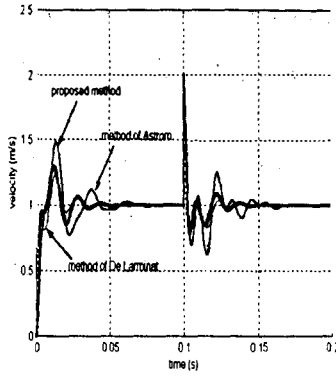


Figure 11: Step response of speed loop followed by a velocity disturbance

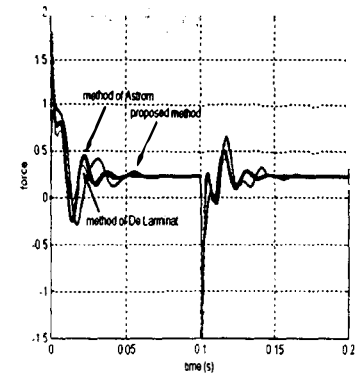


Figure 13: Control signal of speed loop

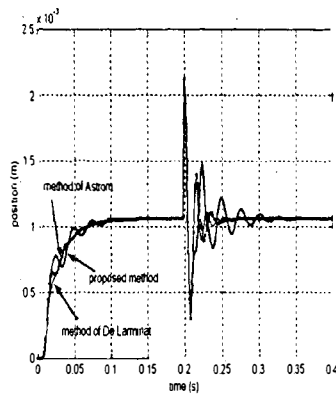


Figure 12: Step response of position loop followed by a load disturbance

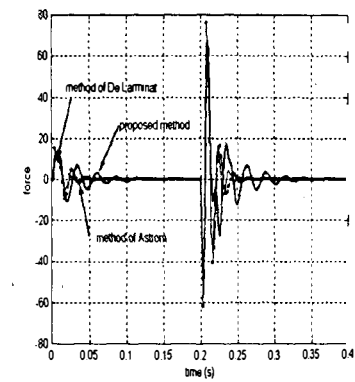


Figure 14: Control signal of position loop

5. Conclusion

This paper presents an automatic design method for cascade structure CNC controllers. The procedure takes into account the elasticity of bodies, the effect of numerical control as delays. The robustness is ensured by requiring minimal stability margins and the performances are followed from this margins. Finally, we show that results obtained with this method give an acceptable controller compare with other strategies.

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