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Anti Windup Position control of synchronous machine with current limitation

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Abstract. *This paper deals with a systematic approach for design of a robust controller in presence of input saturation. The proposed approach is based on a describing function analysis of classical anti wind-up schemes. This synthesis uses a new parametrization of the anti wind-up scheme elements, allowing us a efficient tuning of the anti wind-up loop. Practical result on position control of synchronous machine with current limitation exhibits the high potential of the method to avoid limit cycle risk, while preserving good performances. The experimentation involve a simulation of the plant with POSTMSYNC software and the implementation on the benchmark*

1. INTRODUCTION

On every plant there are some variables which can saturate, especially the input signal. For example concerning electrical actuator, physical limitation of electronic power components composing the power amplification, imposes limitation on voltage and current amplitudes. This kind of constraints can be represented by input saturation as shown in fig. 1.

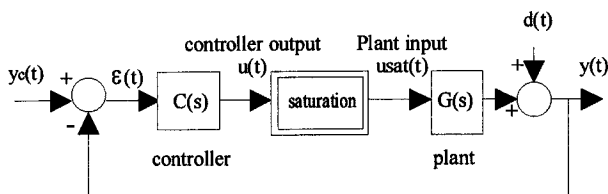


Fig. 1. controlled process with input saturation

The presence of input saturation may develop problem on the output of the plant. In fact, when the output signal $u(t)$ of the controller reaches the saturation level, the signal $u_{sat}(t)$, which actually feeds the plant, becomes saturated, producing large error at the input of the controller. Then the controller reacts uselessly by building large output signal $u(t)$. This excess control signal may produce large overshoots in the process output when the controlled plant emerges from saturation, and even leads to limit cycles or instability.

Anti wind-up control (AWC) principle consists in using the signal $u_{sat}(t)$ in order to build the controller output. However, the signal $u_{sat}(t)$ is generally not measurable. So all the classical approaches introduce a saturation model in the controller to rebuild the $u_{sat}(t)$ signal. As the new saturation included in the controller avoids the effect of the process saturation, the process saturation will be omitted on following diagrams.

The solution proposed in [1] consists in inserting a model of the saturation non linearity at the input of the process model, for an internal model control scheme (IMC). Another solution, proposed in [2], consists to build the anti windup system so that the open loop around the saturation model behaves roughly like an integrator with gain k . He suggests to adjust k so that this loop has large gain and bandwidth compared to those of the open loop $G(j\omega).C(j\omega)$. In [3] those two systems are compared with another system called "Conventional Anti-windup System", or CAW, and it's own approach, which differs of CAW only in the multivariable case. The principle for tuning CAW consist to build the difference between the controller output $u(t)$ and the system input $u_{sat}(t)$, and then to feedback this difference through a gain X , into some coprime factorisation $C(s)=V^{-1}(s).U(s)$ of the linear controller. If the associated loop transfer function $L_X = V^{-1}(s).X(s)$ has gain and bandwidth much higher than $G(j\omega).C(j\omega)$, this prevents the windup. The conclusion after a simulation on a SISO example is that Horowitz [2] and IMC [1] methods frequently lead to much poorer responses than CAW, but also that the CAW method, can lead to limit cycles or instability. A conclusion is that some "intermediate" design has to be found, between Horowitz and CAW methods. An approach to determine what must be this "intermediate" design, can be found in [4]. This approach, based on the describing function method, will be explained in detail in section 2.5.

In the first part of the paper, an interpretation of the scheme fig. 1 as an observed state feedback control scheme, leads to break it in two parts, so as to feed the observer with $u_{sat}(t)$ instead of $u(t)$. The signal $u(t)$ then becomes the same as the one generated

by true state feedback control law, in presence of saturation. In the second part, a phase distance criterion is used to decide what part of closed loop dynamics must be interpreted as "observer dynamics", to ensure good performances under saturation. The conclusion is that it may arise that state feedback leads to bad performances in presence of saturation, and a design procedure based on the use of this phase distance criterion is then proposed to adjust the dynamics of state feedback.

The sequel of this paper is organised as follows. In section 2, the approach adopted for tuning an AWC in case of SISO process is presented. A new parametrization of the elements of anti-windup schemes is used to bring to the fore the effects of an anti windup structure. So a method to design these elements to avoid windup or limit cycles is proposed, based on the use of the phase distance criterion proposed in [4], and the preceding parametrization. In section 3, this new control law is tested and compare on electrical drive constituted by a current controlled synchronous machine. Some simulation results are explains.

2. ANTI-WINDUP CONTROL

A. Objectives

An AWC has to deal with two different objectives.

Firstly it has to ensure that no large overshoots, limit cycles, or instability, will appear, when the input signal $u_{sat}(t)$ actually applied to the system reaches the saturation level.

Secondly, in linear case, the saturation behaves like a unity gain, and the AWC must have the same transfer function $C(s)$ between $e(s)$ and $u(s)$ as the linear controller of fig. 1 which has been designed without any saturation.

B. Anti wind-up schemes

The presented AWC structures are all based on the same key idea, which consists to use the signal $u_{sat}(t)$ as an auxiliary input, to build the controller $C(s)$.

This one is then realised with two different transfer blocks, $C_u(s)$ and $C_y(s)$. The essential property is to produce an open loop transfer function $\beta_{NL}(s)$ around the saturation element, which can differ from the open loop transfer function $\beta(s) = C(s)G(s)$ obtained in fig. 1. An appropriate tuning of $\beta_{NL}(s)$ permits to get better performances in case of saturation. The question discussed here is not about anti-windup structure, which is depicted in fig. 2, but rather in the tuning of $C_u(s)$ and $C_y(s)$ so as to obtain good performances.

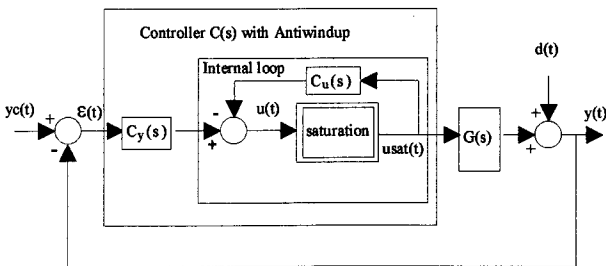


Fig. 2. Anti wind-up system

C. Polynomial parameter of C_u and C_y

In linear mode the saturation behaves like a unity gain transfer function, which gives the following equation:

$$u_{sat}(s) = \frac{C_y(s)}{1 + C_u(s)} \cdot e(s) \quad (1)$$

Furthermore, as it can be seen from fig. 1, $C(s)$ is of the form

$$u_{sat}(s) = C(s) \cdot e(s) \quad (2)$$

When no input saturation occurs, equalling the two transfer functions leads to:

$$\frac{C_y(s)}{1 + C_u(s)} = C(s) \quad (3)$$

So that the structure in fig. 2 behaves like the structure in fig. 1, in linear functioning.

In the sequel, and without any lost of generality, assume that C_y and C_u have the same denominator $\Delta(s)$. Then the two parts of the controller can be written as :

$$C_y(s) = \frac{N_y(s)}{\Delta(s)}, \quad C_u(s) = \frac{N_u(s)}{\Delta(s)}$$

The linear controller can be written as :

$$C(s) = \frac{N_C(s)}{D_C(s)} \quad (4)$$

With these notations the equality (3) becomes :

$$\frac{N_y(s)}{N_u(s) + \Delta(s)} = \frac{N_C(s) \cdot D(s)}{D_C(s) \cdot D(s)} \quad (5)$$

where $D(s)$ is an arbitrary polynomial of degree p , all zeros of which are left half plane. The cancellation between right half plane poles and zeros being excluded.

D. Open loop transfer function $\beta_{NL}(s)$

The open loop transfer function $\beta_{NL}(s)$ for the scheme fig. 2 is the transfer function between $u_{sat}(t)$ and $-u(t)$ obtained when the saturation is removed from the scheme, which gives the following expression :

$$\begin{aligned} \beta_{NL}(s) &= C_u(s) + C_y(s) \cdot G(s) \\ \beta_{NL}(s) &= \frac{N_u(s)}{\Delta(s)} + \frac{N_y(s)}{\Delta(s)} \cdot \frac{N_G(s)}{D_G(s)} \end{aligned} \quad (6)$$

with the following plant transfer function :

$$G(s) = \frac{N_G(s)}{D_G(s)}$$

From the equality (5)

$$N_y(s) = N_C(s) \cdot D(s) \quad (7)$$

$$N_u(s) = D_C(s) \cdot D(s) - \Delta(s) \quad (8)$$

So the equation (6) can be written :

$$\beta_{NL}(s) = -1 + \frac{D_C(s) \cdot D(s)}{\Delta(s)} \cdot [1 + C(s) \cdot G(s)] \quad (9)$$

E. Shape of Anti-windup controller

Using the describing function method [5], we will now present the way we propose to shape $\beta_{NL}(j\omega)$ in the frequency domain. The ideas about what to do to correctly shape $\beta_{NL}(j\omega)$ are taken in [4]. Firstly, provided that $\beta_{NL}(j\omega)$ has sufficient low-pass properties, non intersection between $\beta_{NL}(j\omega)$ and the negative inverse amplitude-dependent describing function $N(u_1, u_0) = \frac{-1}{s(u_1, u_0)}$ will ensure that no limit cycles or instability will occur in

non-linear functioning. Suppose, $u(t) = u_0 + u_1 \sin \omega t$ then for a constant value u_0 , $s(u_1, u_0)$ is a positive real decreasing function of u_1 . The max value of which is unity, in linear functioning. However, if the distance between $\beta_{NL}(j\omega)$ and $N(u_1, u_0)$ becomes too short at some given frequency, bad damped oscillations may appear in presence of saturation. The key idea to tune the AWC consists in maintaining a sufficient distance between $\beta_{NL}(j\omega)$ and $N(u_1, u_0)$ at all frequencies. In [4], a criterion is proposed which roughly corresponds to choose $\beta_{NL}(j\omega)$ with an argument greater than -140° , when its modulus is greater than unity. Remember that for a saturation non linearity, $N(u_1, u_0)$ is a half straight line in the Nichols chart, with an argument equal to -180° , and a modulus greater than unity. So this criterion can be interpreted as a constraint of minimal phase distance between $\beta_{NL}(j\omega)$ and $N(u_1, u_0)$. However solution proposed in [4] is based on observed state feedback analysis. This approach causes the choice of polynomials to be restricted to $D(s) = 1$ and $\Delta(s)$ as a divisor of closed loop denominator with the same degree as $D_c(s)$. We will see on the example that this choice may sometimes be too restrictive to permit a good tuning of the anti windup system.

F. New approach of the problem

First note that for the internal anti wind-up loop formed by $C_u(s)$ and the saturation model to be realisable, the transfer function $C_u(s)$ must be strictly proper, so the $N_u(s)$ degree must be lower than the $\Delta(s)$ degree. In this case $C_y(s) = C(s)(1 + C_u(s))$ is also realisable. Assuming this condition, the equation (8) gives a solution, if and only if the polynomials $\Delta(s)$ and $D_c(s).D(s)$ have the same degree and highest-degree terms. This implies that:

$$\lim_{\omega \rightarrow \infty} \frac{D_C(j\omega).D(j\omega)}{\Delta(j\omega)} = 1 \quad (10)$$

Assume now that a linear controller $C(s)$ has been previously computed to ensure good performances in linear functioning. Then nearby the crossover frequency ω_u , these performance objectives proscribe intersection or too short phase distance between $N(u_1, u_0)$ and $\beta(j\omega)$. So we can choose $\beta_{NL}(j\omega)$ close to $\beta(j\omega)$, in this frequency range. This consideration suggest us to choose the two polynomials $D(s)$ and $\Delta(s)$ such as:

$$\frac{D_C(j\omega).D(j\omega)}{\Delta(j\omega)} \approx 1 \quad (11)$$

near the crossover frequency.

In the high frequency range, where $|\beta(j\omega)| \ll 1$, the AWC has just to ensure that $\beta_{NL}(j\omega)$ will have small gain, essentially in order to ensure robust stability of the whole AWC scheme. This

small gain condition will be satisfied if the equation (11) is true in the high frequency range.

In the low frequency range, where $|\beta(j\omega)| \gg 1$, that we have to maintain the argument of $\beta_{NL}(j\omega)$ far enough from the -180° of $N(u_1, u_0)$, to prevent windup or instability. Assume now that :

$$|\beta(j\omega)| \gg 1, \text{ and } \left| \frac{D_C(j\omega).D(j\omega)}{\Delta(j\omega)} \cdot \beta(j\omega) \right| \gg 1 \quad (12)$$

then we obtain from (9) the following expression of $\beta_{NL}(j\omega)$:

$$\beta_{NL}(j\omega) \approx \frac{D_C(j\omega).D(j\omega)}{\Delta(j\omega)} \cdot \beta(j\omega) \quad (13)$$

Using this expression makes it very easy to choose $D(s)$ and $\Delta(s)$ to respect the phase distance criterion, using a classical frequency domain open loop shaping approach.

G. Polynomial expressions of $C_u(s)$ and $C_y(s)$

The expression (11) implies that poles of $C(s)$ corresponding to high frequency dynamics must be simplified by some zeros of $\Delta(s)$.

The Nichols plot of $\beta(j\omega)$, and the expression of $D_C(s)$, are used to build the polynomials $\Delta(s)$, and $D(s)$. If the tuning of $\beta_{NL}(j\omega)$ is not satisfied with $D(s) = 1$, then it is possible to increase the degree of $D(s)$ accordingly to the one of $\Delta(s)$, until a sufficient phase distance between $\beta_{NL}(j\omega)$ and $C(u_1)$ is found.

3. EXAMPLE

The plant is a electrical drives constituted by a synchronous machine with field winding fed by three phase inverter.

A. Studied device presentation

The current control into a synchronous machine is compulsory to control the torque. With the set value of the R.M.S. phase current I_{ref} and the load angle ψ , one can set the currents by different ways. Notably, the inverter control can be realised either using a hysteretic control. Its consist in use a three independent on-off regulators (fig. 3).

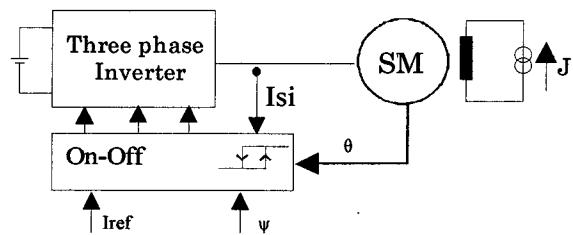


Fig. 3

This structure of control is particularly efficient especially in dynamic aspect. The inverter can be represented by a little time constant t_a . By another way, the torque of the synchronous machine depends heavily on the load angle. The angle can then be used to control the stator-rotor inductive coupling. It can either optimise the motor operation by giving a maximum torque for a specific current, or optimise the inverter action by imposing a unit power

factor. With a smooth pole machine the torque is maximum when the load angle kept equal to zero.

$$T_m = p \cdot M_{sr} \cdot j \cdot \sqrt{2} \cdot I_{eff} \cdot \cos(\psi) = K_a \cdot I_{eff} \quad (14)$$

j : field current
 M_{sr} : mutual inductance stator-rotor
 p : number of poles pair
 I_{eff} : RMS current
 ψ : load angle.

This strategy of control generate a torque quasi-proportional to reference current I_{ref} . Like this the position control can be realized by implementation only one regulator, see fig. 4.

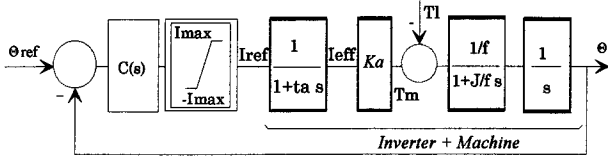


Fig. 4

The current input is constrained by saturation amplitude in order to protect machine. The behaviour of this drive depend a regulator design.

B. Controller design

We propose to control the rotor position with a numerical anti-windup controller (fig.5-a) and compare with a classical structure (fig.5-b) use en electrical drive, in[6].

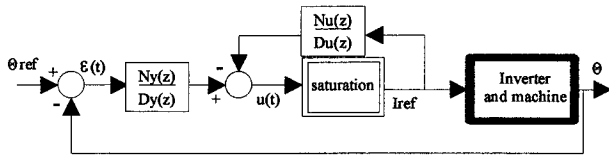


Fig. 5-a

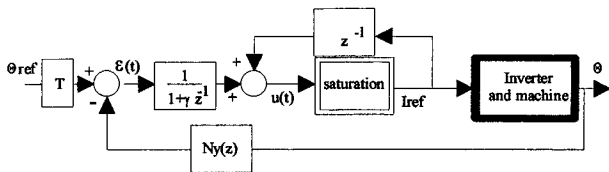


Fig. 5-b

with $T=N_\gamma(1)$ and $D_\gamma(z)=(1-z^{-1})(1+\gamma z^{-1})$

The sampled control is designed as a pseudo-continuous problem by using the bilinear transform $z = \frac{1+w}{1-w}$. The synthesis uses the same approach presented on the second paragraph but in the w plane. The plant model $G(z)$ and the lead-lag type controller $C(z)$ are given by :

$$G(z) = TZ \left\{ \frac{1-e^{T_s} e^{s}}{s} * \frac{K_a / f}{(1+t_a s)(1+\tau s)} \right\} \quad (15)$$

$$C(z) = \frac{988,8 - 1910z^{-1} + 922,1z^{-2}}{1 - 1,4274z^{-1} + 0,4274z^{-2}} \quad (16)$$

with $t_a = 10^{-4} s$, $t_e = 10^{-3} s$, $K_a = 3,818$, $\tau = \frac{J}{f}$, $f = 62,7 \cdot 10^{-3}$

Nm.s/rd ; $J = 32 \cdot 10^{-3} kg.m^2$;

and saturation levels at +/- 4.8 A. The Nichols plot of the open loop $\beta(j\omega) = C(j\omega)G(j\omega)$ is depicted fig. 6.

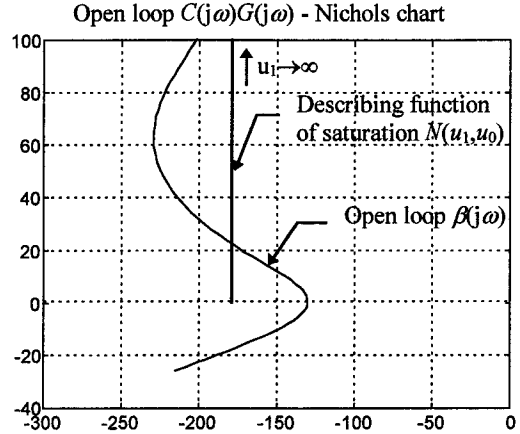


fig. 6 : Nichols plot $\beta(j\omega)$

It can be seen on fig. 6 the presence of one intersection between $\beta(j\omega)$ and $N(u_0)$, for pseudo-frequencies $\nu = 0.015$. This point corresponds to an unstable point, see in [5] for details.

The design of the two polynomials $D(z^{-1})$ and $\Delta(z^{-1})$, with the use of the phase distance criterion will now be developed.

On this example the polynomial $D(z^{-1})$ is not used. For the AWC to be realisable, $\Delta(z^{-1})$ must be a monic polynomial of degree two, such as $D(1)=\Delta(1)$ in order to ensure that the controller $C_u(z^{-1})$ to be realizable.

For example choose

$$\Delta(z^{-1}) = \alpha(z^{-1} + z_{\Delta_1})(z^{-1} + z_{\Delta_2}) \quad (17)$$

So that

$$\frac{D_c(z^{-1})}{\Delta(z^{-1})} = \frac{1 - 1,427443z^{-1} + 0,427443z^{-2}}{\alpha(z^{-1} + z_{\Delta_1})(z^{-1} + z_{\Delta_2})} \quad (18)$$

Then compute parameters z_{Δ_1} and z_{Δ_2} so as to ensure phase distance of 40 degrees between $\beta(j\omega)$ and $N(u_0)$, (classical frequential lead controller design in w -plane).

The numerical expression of the polynomials $N_y(z^{-1})$, $N_u(z^{-1})$, $D_y(z^{-1})$, and $D_u(z^{-1})$, by use of equations (7) and (8), are

$$N_u(z^{-1}) \approx -0,05825z^{-1}$$

$$D_u(z^{-1}) \approx 1 - 0.941747z^{-1}$$

$$N_y(z^{-1}) \approx 988.8 - 1910z^{-1} + 922.1z^{-2}$$

$$D_y(z^{-1}) \approx 1 - 1.36919z^{-1} + 0.402543z^{-2}$$

The open loop $\beta_{NL}(j\omega)$ obtained for these values is depicted on fig. 7. It can be seen on this fig. that the phase distance criterion is roughly respected. So, reset windup, limit cycles or instability possibility have been avoided.

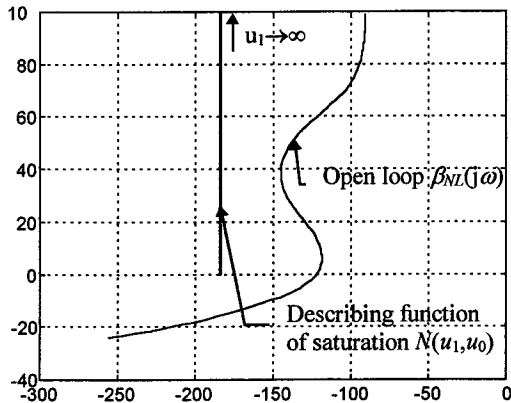


fig. 7 : Nichols plot of $\beta_{NL}(j\omega)$

C. Simulation results :

The investigation is lead with a POSTM5 software, see [7]. This software is specially design for a test and compare a control law. The software is divided into two modules. The first one concerns the process while the second concerns the control . Because the mathematical models are already implemented, the first part is totally flexible. That is, the user can select the model he wishes to run among a library of pre-defined models. The parameters associated with the selected mathematical model may then be specified. This part of the software also includes the time management, the control of each module state and the writing of the results in a file. The control module of the software corresponds to a file specified in the C language and following a pre-defined format. This file allows the user to specify its own control laws which would then communicate with the other modules via its input/output.

The reached performances will be compared taking into account load torque variation and step reference position.

C.a: Small displacement:

In order to compare the performances of these laws we propose a no-load start up with a reference fixed at 10 degrees. Fig. 8 shows the greater behaviour generate by the AWC law.

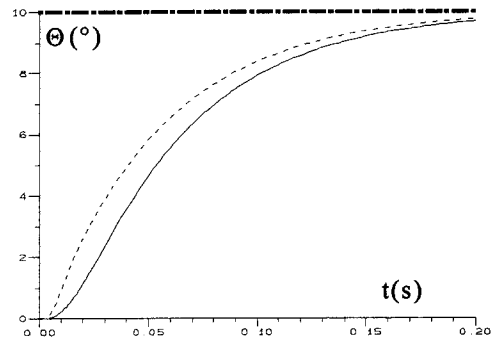


fig. 8.

In spite the small reference the input signal reach the saturation level and permit to obtain a good time response, fig. 9.

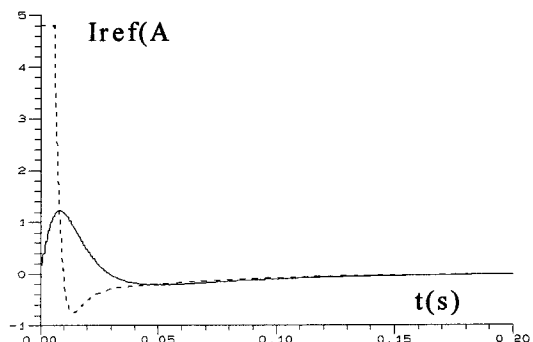


fig. 9.

C.b Large displacement:

For large step reference the AWC assure a better transient (fig. 10) and the time response is practically preserved.

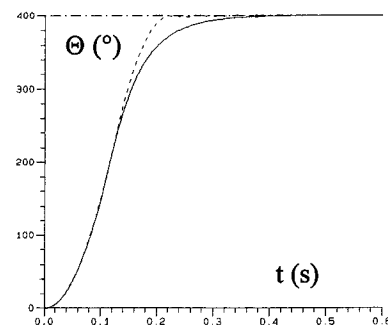


Fig. 10

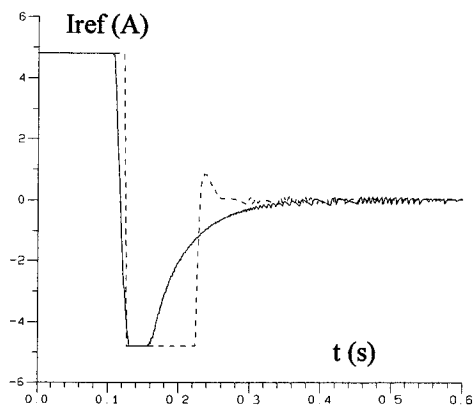


Fig. 11

C.c: Disturbance rejection

The fig. 12 shows the behavior of this drive in presence of torque disturbance. At 0.4 s the load torque go to 10 N.m. The both laws present the same response because the input signal Iref stay in linear area.

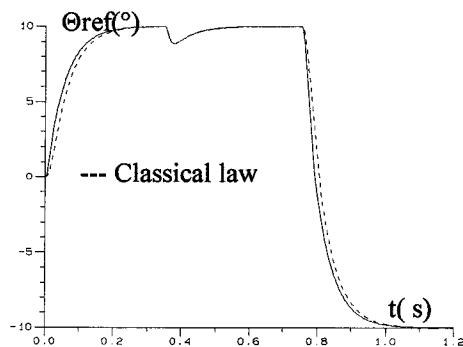


fig. 12.

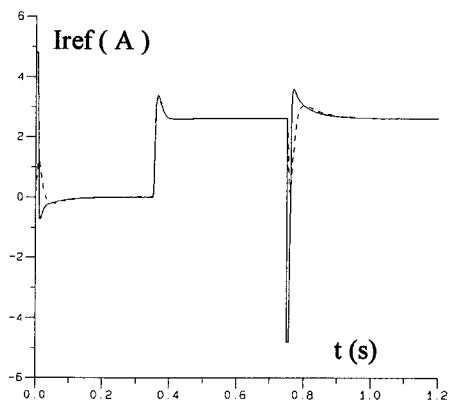


fig. 13.

4. CONCLUSIONS

In this paper, a method to design a anti windup system in presence of input saturation has been presented. This method, based on a special parametrization of classical anti-windup systems elements and a phase distance criterion allows it to define a

efficient tuning of the Anti Windup Controller, with a classical frequential loop shaping approach. This method exhibits the same good performances than conventional anti-windup for "little levels" of saturation, but ensure the system to remain stable, and avoid limit cycles. Note This method is of particular interest when you do not know exactly the model of your system. Assuming all possible models have the same number of positive real part poles, it is quite easy to choose the parameters of the anti-windup scheme so that the set of possible frequential responses $\beta_{NL}(j\omega)$ respects the phase distance criterion.

REFERENCES

- [1] Morari and Zafriou (1987)
- [2] Horowitz I. (1983) *A synthesis theory for a class of saturating systems* International Journal of Control VOL. 38 N° 1, 169-187
- [3] Doyle J.C., R.S. Smith, D.F. Enns (1987) Control of plants with input saturation nonlinearities. Proc. 1987 American Control Conference, 1034-1039
- [4] Wurmthaler Ch., P. Hippe (1991). *Systematic compensator design in the presence of input saturation* European Control Conference, 1268-1273, Grenoble, France.
- [5] Gelb A., W.E. Vander Velde, (Ed) (1968) *Multi-input describing functions and nonlinear system design*, McGraw-Hill, New York.
- [6] Chouzal F., M. Fadel, B. Dagues *Comparative study of different control laws of a synchronous position drive by means of the simulation program POSTM5 SPEEDAM*, June 1994, Taormina, Italy.
- [7] Dagues B., M. Fadel, X. Roboam, P. Azéma :*Post Simulation Software : Study and teaching of digital control of static converter / electrical machine assemblies*.EPE'95 Seville Spain.