

Interaction between the MD and CD control processes in papermaking and plastics machines

Optimal tuning criteria

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The transition from the low-frequency range to the high-frequency range is continuous, so a method of control based on frequency analysis must accurately decouple the MD and CD frequency signals.

Control in the machine direction (MD) and control in the cross-machine direction (CD) are not independent. If their interaction is neglected in the analysis of control, this oversight leads to a deterioration in the quality of control. The interactive nature of CD and MD control is described well by others (1).

Generally speaking, the MD control loops are designed to alleviate low-frequency disturbances, while the CD control loops have to compensate for the high-frequency ones. Because the transition from the low-frequency range to the high-frequency range is continuous, a method of control based on frequency analysis must accurately identify the boundaries beyond which MD high-frequency signals pass into the CD control and vice versa. The ability to distinguish these boundaries would make it possible to decouple the inherent interaction and separate the parameters needed for both control loops.

Such an identification procedure leads to various other criteria that can be used either to determine the performance limits of an MD control process in a certain environment or to quantify the performance limits of the unit (the actuator or machine) to be controlled.

Processes with time lag and transport delay

Consider a process with a time delay and a first-order time constant (Fig. 1). This process is described by the following transfer function

$$G_s(j\Omega) = V_s \frac{e^{-T_t j\Omega}}{1 + T_s j\Omega} \quad (1)$$

where

- V_s = gain factor of the process
- T_t = time delay of the process
- T_s = time constant of the process.

This simplified mathematical model could represent the control of basis weight, for example. A Smith Predictor is usually implemented to control processes in which the time delay is dominant. The Smith Predictor is described by the transfer function:

$$G_k(j\Omega) = V_k \frac{1 + T_s j\Omega}{1 + T_g j\Omega - e^{-T_t j\Omega}} \quad (2)$$

where V_k and T_g are tuning parameters. (Incidentally, this is the method frequently used for basis weight and caliper control. This is not always the best solution for processes with noise disturbances, as will be shown later.)

When the controller unit is switched on, the closed loop transfer function is described through:

$$G_g(j\Omega) = \frac{e^{-T_t j\Omega}}{1 + T_g j\Omega} \quad (3)$$

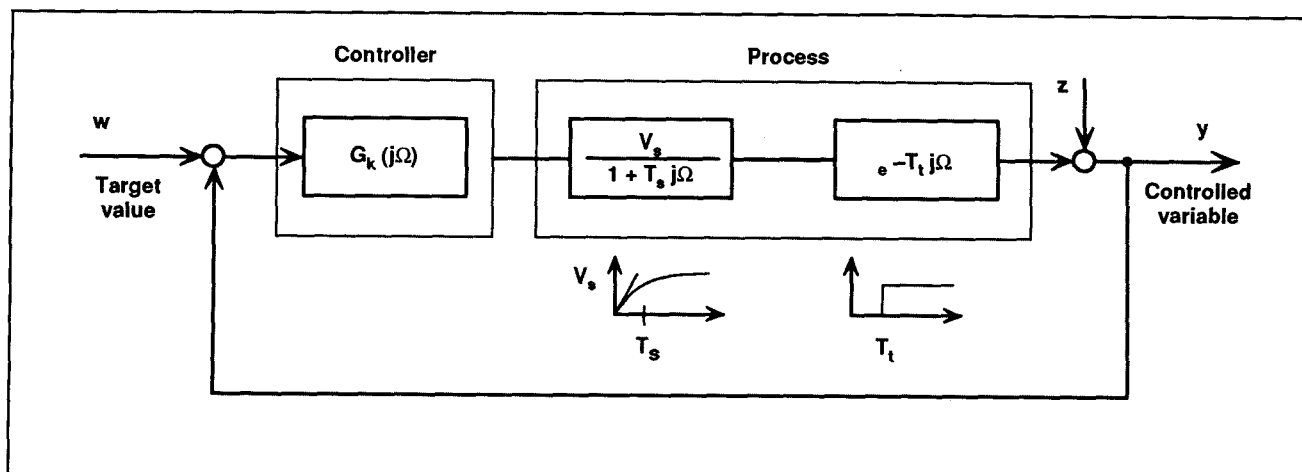
with

$$V_k = 1/V_s$$

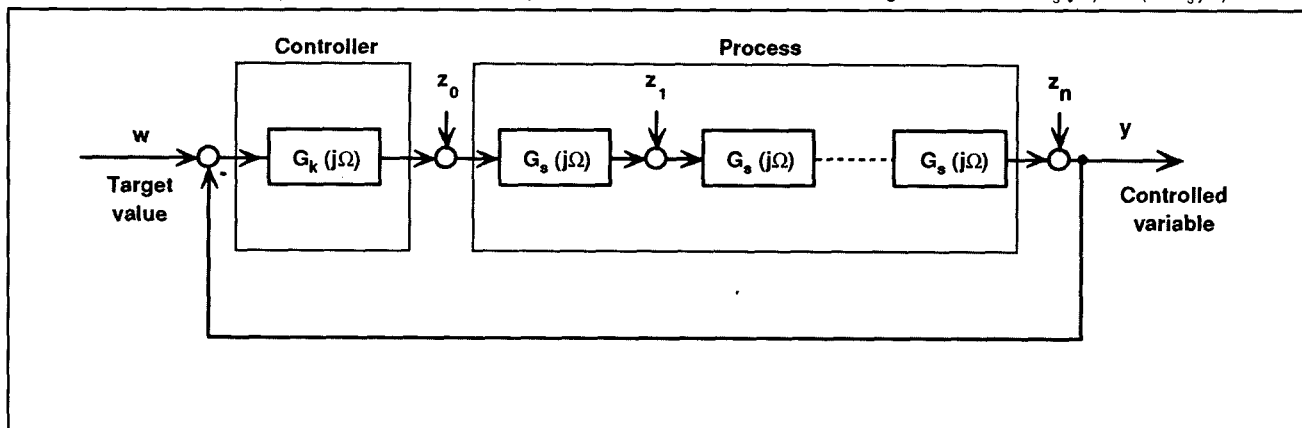
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Because the process time delay cannot be changed, the behavior of the control loop can be adjusted with the

1. Controller unit (GK) for a process with time delay and time constant. The variable z represents an external disturbance signal.



2. Control loop with different places of disturbance. The process consists of n first-order time lag blocks, where $G_s(jΩ) = 1/(1 + T_s jΩ)$



parameter T_g . But how is T_g to be selected? Under which quality criteria?

Besides the demands for stability, a design criterion for optimal control will be the speed of the loop behavior. This criterion is important, but it is more important to adapt the design to the expected interaction between MD and CD control. Again, a decision must be made.

Just for that decision, it is essential to have a systematic approach helping to quantify the interactions by identifying the critical noise frequencies. Once the component frequencies of the machine noise are identified and analyzed, then one will be able to specify the corresponding controller type for synthesis, and the expected control quality can be quantitatively assessed and judged.¹

Here, we derive a procedure to determine a criterion that will allow the noise frequencies to be identified on-line. This procedure will allow us to decouple the influences in the MD and CD directions.

Closed-loop behavior of processes with time delay and time constant, driven by disturbance

The closed loop subjected to disturbance behavior describes the dynamics of the controlled variable (such as basis weight) as it responds to disturbances. This disturbance transfer function of the process (Eq. 3) is analyzed.

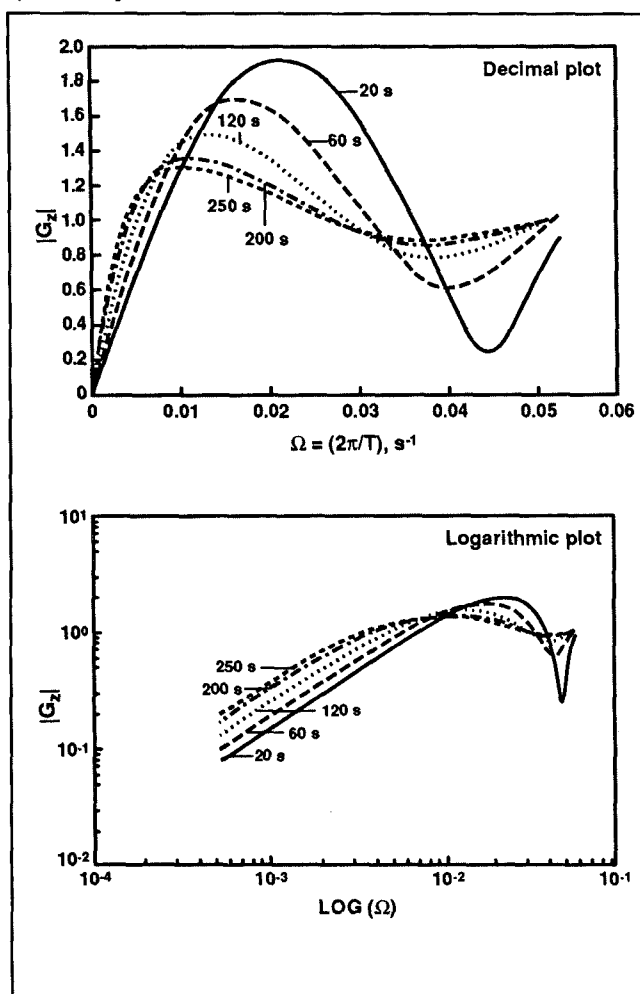
It is assumed that the disturbance interacts at the output of the controlled process. This assumption represents the worst case situation. If the noise takes place within, or even at the input of the process (such as headbox instabilities), then the remaining part of the process is relevant for its propagation in the loop. If the remaining process part has a first-order time lag behavior, the disturbance will reach the controller with a dampened intensity.

To clarify this, the process is considered to be built as a series of n first-order time lag blocks (such as for a dryer section for a moisture control loop). The disturbance transfer function can be described through Eq. 4, with Fig. 2:

$$G_z(jΩ) = \frac{[G_s(jΩ)]^{n-k}}{1 + G_k(jΩ) [G_s(jΩ)]^k [G_s(jΩ)]^{n-k}} \quad (4)$$

¹When the frequency components have been analyzed, it is possible to remove the disturbance parts that are beyond the controllable range. To achieve this, these signal parts must be simulated via an autoregressive model and fed back into the process (feed forward control).

3. Frequency responses of control loops with time delay and time constant using a Smith controller. T_g is varying from 20 s to 250 s. $T_i = 120$ s, $T_s = 35$ s.



where k gives the point where the noise comes into the system.

Now, considering again processes with time constant and time delay as they are described in Fig. 1, using a Smith Predictor, the highest controllable frequency is given by the limit

$$\Omega_{\max} \leq \frac{1}{T_i + T_g} \quad (5)$$

where

T_i = time delay of the controller

T_g = time constant of the controller.

For practical reasons which will be seen later, it is not possible to go beyond this limit. The disturbance transfer function of the closed loop is

$$G_z(j\Omega) = \frac{Y}{Z} = \frac{1}{1 + G_k(j\Omega) G_s(j\Omega)} \quad (6)$$

Its response to a sinusoidal input signal, in the continuous representation, is obtained from the relationships

$$G_z(j\Omega) = 1 - \frac{e^{-T_i j\Omega}}{1 + T_g j\Omega} \quad (7)$$

and

$$|G_z(j\Omega)| = \sqrt{\frac{2 + T_g^2 \Omega^2 + 2(T_g \Omega \sin \Omega T_i - \cos \Omega T_i)}{1 + T_g^2 \Omega^2}} \quad (8)$$

With a discrete model, the response to a sinusoidal input signal is obtained from the relationships

$$G_z(z) = 1 - \frac{(1 - a_R) z^{-(n+1)}}{1 - a_R \cdot z^{-T}} \quad (9)$$

and

$$|G_z(z)| = \sqrt{\frac{2[1 - a_R + a_R^2 - a_R \cos(\Omega \Delta T)] - (1 - a_R) \{\cos[(n+1)\Omega \Delta T] - a_R \cos(n\Omega \Delta T)\}}{1 + a_R^2 - 2a_R \cos(\Omega \Delta T)}} \quad (10)$$

where

ΔT = the sample time.

The appropriate value for the tuning parameter T_g can be derived from the Eqs. 8 or 9. Its value must be chosen so that Eq. 11 is valid:

$$|G_z(j\Omega)| \leq 1 \quad (11)$$

Equations 8 and 11 give the relationship in Eq. 5, and thus the limit for T_g is determined.

Figure 3 shows the disturbance transfer function for an MD control described by Eq. 3. The sample time was chosen to be 10 s (or $\Delta T = 10$ s).

The consequences from the development of the response to the disturbance are obvious. The range of frequencies that can be controlled can be varied with the time constant T_g in the closed loop controller.

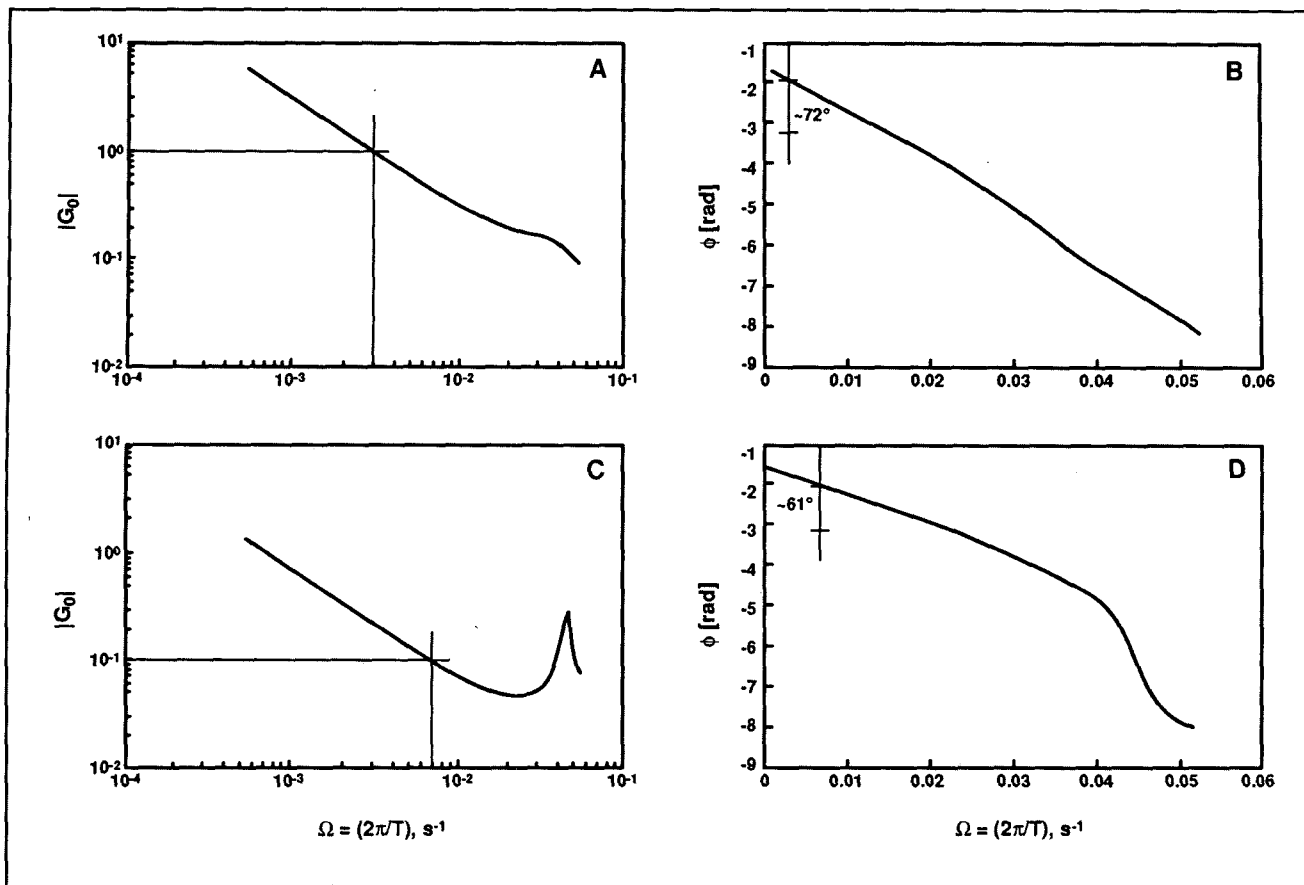
If T_g is chosen to be close to the time constant T_s , (or $T_g \approx T_s$), then the disturbance frequencies, they can be compensated, have a ceiling of $\Omega = 1 / (T_i + T_s)$. That means that noise frequencies in the range of $\Omega \geq \Omega_{\max}$ will be amplified. It is not possible, however, to reduce T_g as one would like in the expectation that higher frequencies can be influenced, because the intersection frequency will be shifted to the right and thus the phase margin will be reduced to the region around -180° , where stability becomes a problem. The stability margins are already narrow because of the time delay, which produces a fast roll-off above the frequency $\Omega = 1 / T_i = 0.01$, and should not be made more narrow.

These aspects can be read from the gain and phase plots shown in Fig. 4. They are the characteristics of the open loop controller

$$G_o(j\Omega) = V_s \frac{e^{-T_i j\Omega}}{1 + T_g(j\Omega)} V_R \frac{1 + T_s(j\Omega)}{1 + T_g(j\Omega) - e^{-T_i j\Omega}} \quad (12)$$

The characteristic equation for the behavior of the closed loop controller can be determined from the open loop characteristic curves (2).

4. Gain and phase plots for an open loop with Smith Predictor. A: gain with $T_g = 200$ s. B: phase with $T_g = 200$ s. C: gain with $T_g = 20$ s. D: phase with $T_g = 20$ s.



An increase in T_g leads to a shift of the intersection frequency, Ω_d , to lower frequencies and thus to a slower controller behavior. It also leads to an increase in the phase margin, Φ_R , and hence to adequate stability.

This fact can also be seen in Fig. 4, Plots B and D. A change in T_g from 20 s to 200 s leads to an increase in the phase margin and hence an increase in stability of about 11° or 12° (from 61° to 72°). The price for more stability is a decrease of the intersection frequency in lower frequency area and thus a reduction in the speed of the controller response.

As a rule of thumb, an adequate phase margin is considered to be around 70° .

With this explanation, it is now possible to determine the following rules for the MD control. First, T_g has to be chosen according to Eq. 5, so that Eq. 11 is valid. Second, concerning the values of T_s and V_s , T_g has to be chosen according to the following considerations:

The Smith Predictor has the Transfer Function (Eq. 2):

$$\frac{u}{e} = G_k = \frac{(1/V_s) \cdot (1 + T_s(j\Omega))}{1 + T_g(j\Omega) - e^{-T_t(j\Omega)}} \quad (13)$$

where

u = actuator signal

e = error signal.

The step response, for $t \rightarrow +0$ is:

$$u_t \rightarrow +0 = (1/V_s) (T_s/T_g) \quad (14)$$

T_g should be chosen so that the value $u_t \rightarrow +0$ is "reasonable" in respect to the error magnitude. For example, for the basis weight control, a tuning such as $T_g = T_s$ results in equal amounts of changes in the stock flow as in the weight to be controlled. Instead, a choice of T_g according to $T_g = 0.5 T_s$ causes a two times higher change in the stock flow for the same result.

Now, let us go back to the tuning Instruction A.

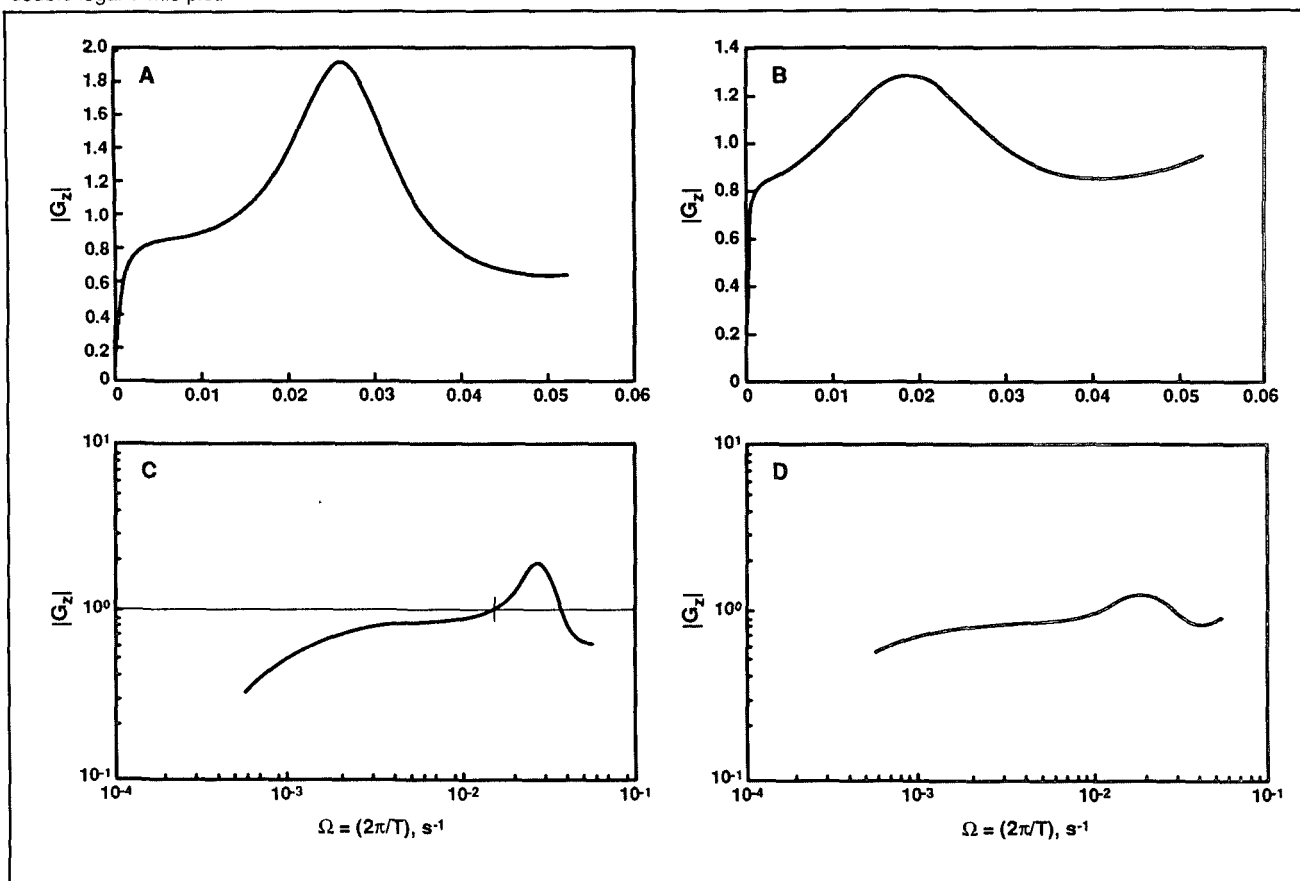
Unfortunately, above this frequency an undesired effect takes place. Figure 3 shows that high noise frequencies in the MD control loop are almost doubled. This means that a trade off should be made between the controllability in the disturbance region $\Omega \leq \Omega_{\max}$ and the amplification of frequencies $\Omega \geq \Omega_{\max}$.

A choice of T_g in the area of $T_g = 3(T_t + T_s)$ will produce instead an amplification of around 20% and is preferable, especially for paper machines with high frequency disturbances (headbox instabilities, pump pulsations, etc.).

Such a situation becomes more critical if a CD control process is implemented alongside the MD control (such as for basis weight and moisture). The CD control loop receives these amplifications in the high frequency range in full power.

This jump-in of the MD disturbances results in an interference between both controls with negative effects. The question arises whether the normally used exponential filters are capable of cutting off these interactions. After

5. Disturbance behavior of PID- and PI-controlled processes. A and C: PID and PI in simple dimensioned plot. B and D: PID and PI in double logarithmic plot.



all, they are designed to achieve this goal.

These filters usually have a structure like

$$\hat{Y}_{i,T} = \hat{Y}_{i,T-1} + k(Y_{i,T} - \hat{Y}_{i,T-1}) \quad (15)$$

with

$\hat{Y}_{i,T}$ = the filtered value of element i in the T -th scan.

$Y_{i,T}$ = the observed value

and

$$k = \frac{T}{T_F + T} \quad (16)$$

with

ΔT = sample time

T_F = filter constant.

With a time scan of around 30 s and a value of k tuned to be about 0.3 (usual tuning), Eq. 16 yields $T_F = 70$ s = 1.16 min.

That means that the disturbance frequencies in the machine direction are influenced by the cross filter first with the frequency

$$\Omega \geq \Omega_{kr} = 1/70 = 0.014 \quad (17)$$

This is, unfortunately, no longer relevant. The disturban-

ces that are amplified through the MD controller cannot be attenuated with this filter.

A more suitable filter constant may be obtained, choosing T_g equal to 20 s, with $k = 0.15$ —that is, a filter time constant of $T_F = 167$ s = 2.78 min. In turn, this means that the disturbance effect would appear in the measured profile after a settling time of $t_{95} = 8.34$ min, which would be far too long a reaction time for control in the machine direction.

Consider now the situation if a PID controller is to be used as compensator. Its characteristics are well known. It gives good static precision and offers the possibility of adjusting either the phase margin or the intersection frequency.

Its transfer function is given by

$$G_k(j\Omega) = V_R \frac{[1 + T_{R1}(j\Omega)][1 + T_{R2}(j\Omega)]}{j\Omega} \\ = V_{R0} [1 + (1/T_N j\Omega) + T_v j\Omega] \quad (18)$$

If it is used as a controller for a process described through Eq. 1, then the Ziegler-Nichols tuning method of controller can be used.

According to that method, the values have to be chosen as

$$V_{R0} = 1.2 \frac{T_s}{V_s T_i}$$

$$T_n = 2 T_i$$

$$T_v = 0.5 T_i$$

These are tuning parameters that constitute a first attempt (default values) to tune the controller and are not necessarily optimal.

According with these parameters, the disturbance behavior over the frequency range as shown in Fig. 5, Frames A and B, is obtained. Here, the highest controllable frequency is with $\Omega_{\max} = 0.014$, practically double the frequency allowed with the Smith Predictor.

In addition, the PID controller, with $T_g = 20$ s and $\Omega_d = 0.008$ offers an even higher damping of the disturbances than the Smith Predictor in the region of the intersection frequency. Only in the small frequency band $0.003 \leq \Omega \leq 0.006$, does the Smith Predictor have the better attenuation.

But even a simple PI controller tuned with the Ziegler-Nichols method gives a clear improvement in the disturbance behavior, with better controllability of the disturbances and stronger attenuation than the Smith Predictor. This comparison is illustrated in Fig. 5, Frames C and D.

Figure 6 shows the disturbance relationship of an optimized control by tuning a PI controller. The Ziegler-Nichols tuning parameters that served as the initial values have been further optimized.

For a PI controller of the form

$$G_k(j\Omega) = V_R [1 + 1/[T_n(j\Omega)]] \quad (19)$$

Ziegler-Nichols recommends the values:

$$V_R = 0.9 T_s / (V_s T_i) = 0.3$$

$$T_n = 3.3 T_i = 295 \text{ s}$$

for a plant with

$$T_i = 90 \text{ s}$$

$$T_s = 30 \text{ s}$$

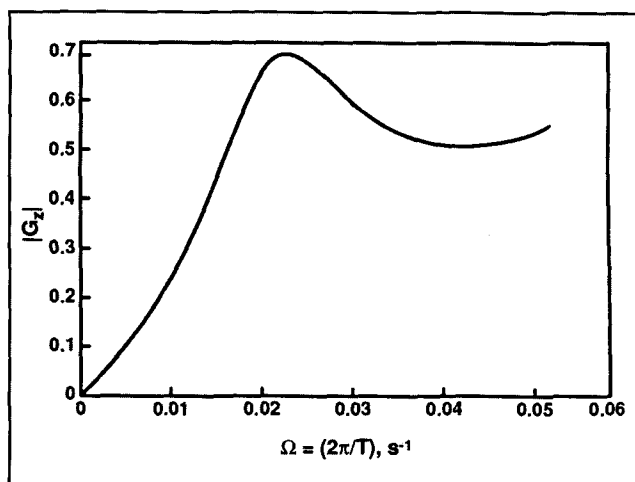
$$V_s = 1.$$

Here, the gain factor is tripled (faster response), and T_n is reduced (minimizing the integral absolute error value). The better performance is clear, if Figs. 5 and 6 are compared.

In conclusion, then, the Smith controller can be used for systems with minor disturbance. However, if the disturbances become noticeable, the control should be performed from a PID controller. With respect to the low-level, high-frequency noises (noise signals with low magnitude but high frequency) which are amplified passing the differential controller gain, a PI controller is preferable.²

²In a feedback loop, the PI (proportional integral) controller sums the error, or the difference between the control signal and the actual measured flow. The PID (proportional integral derivative) controller sums the rate of change of error, or the error multiplied by the proportional gain.

6. Optimized PI controller using the Ziegler-Nichols criteria as initial conditions



Description of a straightforward procedure to identify and decouple the CD and MD frequencies

The foregoing analysis of the frequency behavior allows us to determine a quantitative criterion on which to build a procedure that will lead to the identification of the coupling frequencies.

Step 1. A Fourier series analysis is performed on the MD measured signal (β) to obtain the frequency components of the signal. On the basis of this analysis, the required $\Omega_{\max}$ can be estimated.

Step 2. Given $\Omega_{\max}$, the control parameters for a Smith controller can be determined, so that the relationship

$$T_g \geq (1 - \Omega_{\max} T_i) / \Omega_{\max} \quad (20)$$

is always valid, and therefore

$$|G_z(j\Omega)| \leq 1 \quad (21)$$

for $\Omega \in [0, \dots, \Omega_{\max}]$.

Step 3. The control loop behavior once adjusted according to these steps should be plotted and analyzed with respect to the disturbance frequencies. To do that, the relationship given with Eq. 6 should be evaluated. It describes the frequency spectrum in respect to the disturbance through

$$B(\Omega) = |G_z(j\Omega)| = |1 / 1 + G_k(j\Omega) G_s(j\Omega)| \quad (22)$$

Steps 1-3 outline a procedure to identify the coupling frequency boundaries for the MD and the CD loop and allow us to identify the disturbance frequencies that come from the MD control.

Performance limits for the controller and the system to be controlled

We already have a criterion for CD profile control, which allows us to predetermine the performance abilities of a controller for a given system (β). The same procedure can be used, in exactly the same manner, to predetermine the

performance for an MD controller.

The MD signal must be Fourier analyzed. Using these methods, commissioning engineers would have a procedure to meet the required levels of performance and quality of control in both CD and MD directions.

The procedures could just as easily be used (3) for specifying systems for the customer and can be taken as a basis to specify the actuator dynamic behavior required in the mill. To do this, the designing engineer has to apply the same relationships in the reverse direction.

Summary

A procedure can be used to quantitatively identify the interaction between the MD and CD control processes. Using frequency analysis techniques, the control parameters required to decouple the interaction are determined. The same analysis can be applied to judge the quality of the achieved control as well as the expected actuator performance.

It should not be assumed that all disturbance frequencies can be so influenced. The Smith and PID controllers have their limits.

It is theoretically possible, using an on-line parameter identification method to design a plant-dedicated controller. Such a controller will only be characterized by limits superimposed from the installed actuators and the identification model used. Under these aspects, it is doubtful whether such an engineering investment will repay all the efforts. □

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