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17th July 2003



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Specification

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APPARATUS FOR RELAY BASED MULTIPLE POINT PROCESS FREQUENCY RESPONSE ESTIMATION AND CONTROL TUNING

BACKGROUND OF THE INVENTION

Field of the Invention

The present invention relates to closed loop systems having a regulator for control of a process, and more particularly to an apparatus for estimating the process frequency response and using the estimated frequency response to design the regulator.

Brief Description of the Prior Art

Controllers for tuning the response of a system or a process have been the subject of attention in the prior art. A typical block diagram of such a controller and process is shown in Fig. 1 wherein the output y of a process 10 is fed back through a feedback network 12, the output y' of which is subtracted from a reference signal r by a comparator 14 which outputs an error signal e . A control circuit 16 is designed to respond to the error signal e by outputting a process control signal u for directing the process 10 to output a corrected y such that the error e is zero.

Achieving automatic tuning involves identification of the process 10, and the design of the controller 16. The control output u as a function of time is applied to the process 10 to assure the desired output y . It is generally convenient in analysis of such processes and control circuitry to mathematically transform the parameters from the time domain to a frequency domain. This process is done in electrical circuit analysis by means of the well known Fourier Transform. Manipulation of design parameters is much more conveniently achieved in a frequency domain. The frequency range of interest for such applications is usually from zero up to the process critical frequency ω_c i.e. the frequency at which the phase of a process frequency response crosses -180° . This critical frequency and the process frequency response at this point are useful for control analysis and design. For traditional frequency response

1 identification, the frequencies of the exciting signals u fed into the process/system 10 should be
 2 carefully selected based on the process bandwidth. The use of relay feedback can
 3 automatically excite an unknown stable process around this frequency ω_c . Such a method is
 4 described in U.S. Pat. No. 4,549,123 by Hagglund and Astrom. Furthermore, traditional
 5 frequency response identification usually involves an open loop test, while the use of relay
 6 feedback involves a closed-loop controller. A closed-loop test is preferred to an open-loop test
 7 in control applications, since it keeps the process close to the set point so that the process
 8 operates in a linear region where the frequency response is of interest.

9 Relay based process identification and control tuning have received a great deal of
 10 attention. The method of Astrom and Hagglund is useful in many process control applications,
 11 but it also faces two major problems. First, due to the use of a describing function
 12 approximation, the estimation of the critical point is not accurate enough for some kinds of
 13 processes. Second, the method of Astrom and Hagglund depends on the single critical point
 14 ω_c , and only crude controller settings can be obtained based on this single point. Attempts
 15 have been made to solve this problem, including several modified identification methods using
 16 relay-based feedback control systems in order to identify more than one point on the process
 17 frequency response. In order to accomplish this, additional linear components (or varying
 18 hysteresis width) have been connected into the system requiring additional relay tests to be
 19 performed. These methods are time consuming and the resultant estimations are still
 20 approximate in nature since they rely on repeated use of the standard method as described in
 21 Astrom and Hagglund.

22 It is therefore clear that a method is needed whereby more points on the process
 23 frequency response can be accurately identified from a single relay test.

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SUMMARY OF THE INVENTION

26 It is therefore an object of the present invention to provide an improved apparatus for
 27 process frequency response estimation and controller tuning.

28 It is a further object of the present invention to provide an apparatus for process
 29 frequency response estimation and controller tuning for accurately identifying multiple points

1 on the process frequency response with a single relay test.

2 It is a still further object of the present invention to provide an apparatus for process
3 frequency response estimation and controller tuning for accurately identifying multiple points
4 on the process frequency response with a single relay test, whereby the controller regulator is
5 tuned with respect to the identified points on the process frequency response, whereby the
6 controller system achieves an improved, uniform response.

7 Briefly, a preferred embodiment of the present invention includes an apparatus for
8 tuning a regulator in a process controller. The digitized output from a standard relay and a
9 parasitic relay are fed through a digital to analog converter providing signals at $.5\omega_c$, ω_c and
10 $1.5\omega_c$, to the process, where ω_c is the process critical frequency. The apparatus records the
11 digitized relay output u' and the digitized process output y' until stationary oscillations are
12 reached. The outputs u' and y' are then used to calculate the process frequency response at
13 $.5\omega_c$, ω_c and $1.5\omega_c$. This data is then used to design the regulator, the output of which is
14 converted to analog form and inputted to control the process.

15 An advantage of the auto-tuner of the present invention is that it can estimate multiple
16 points on a process frequency response simultaneously with one single relay test, resulting in
17 time savings.

18 A further advantage of the present invention is that it gives accurate results since no
19 approximations are made, and the computations involved are simple so that it can be easily
20 implemented on microprocessors.

21 A still further advantage of the present invention is that the method is insensitive to
22 noise and step-like load disturbances, and non-zero initial conditions.

23 Another advantage of the present invention is that the controller tuning methods
24 employed work well for the general class of linear processes with different dynamics,
25 achieving consistent and satisfactory responses.

26 IN THE DRAWING

27 Fig. 1 shows a prior art process control system;

28 Fig. 2 is a block diagram of the preferred embodiment of the present invention;

1 Fig. 3A shows the digitized modified relay output for the process frequency response
 2 estimation relay test;
 3 Fig. 3B shows the process output during the relay test;
 4 Fig. 4 illustrates a relay with hysteresis for noise reduction ;
 5 Fig. 5 is a Nyquist plot;
 6 Fig. 6 shows a regulator using an IMC systems configuration; and
 7 Fig. 7 shows the results of PID and IMC regulator designs.

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9

DESCRIPTION OF THE PREFERRED EMBODIMENT

10 Referring now to Fig. 2 of the drawing, a preferred embodiment of the system of the
 11 present invention 18 is shown to include an auto-tuner 20 which is responsive to a digitized
 12 reference signal r at input 22 and a digitized process output signal y' at input 24 to provide
 13 signals u' at output 26.

14 The digitized signals u' are converted to analog signals u by digital to analog converter
 15 28 and outputted at 30 to the process 32. The output y at 34 of the system 18 is shown
 16 schematically to include the sum of the output at 36 of process 32, and noises n and
 17 disturbances w , the addition schematically illustrated through use of summing networks 38
 18 and 40. The output y is converted by analog to digital converter 42 to provide the digitized
 19 input y' at 24.

20 The functions performed by the auto-tuner 20 are schematically illustrated inside the
 21 auto-tuner 20 block.

22 The objective of the auto-tuner 20 is to automatically identify i.e., estimate the
 23 frequency response /dynamics of the process 32, and then to automatically design/configure a
 24 regulator to control the output y of the process 32 according to the applied reference signal r .

25 A more detailed description of the functions performed by the auto-tuner 20 will now
 26 be given in reference to the circuitry illustrated in Fig. 2 inside the auto-tuner 20 block. In
 27 order to identify the frequency response of the process 32, the switch 44 is set to connect the
 28 output 46 of summing network 48 to the input 26 of the digital to analog converter 28. In this
 29 mode, the signal output at 46, provided by the relays 50 and 52, controls the process 32. The

1 digitized process input u' at 26 and digitized process output y' at 24 are then recorded by
 2 conventional circuitry (not shown) until stationary oscillations are reached i.e., until there are
 3 no variations in amplitude or period of oscillations in the output y . The output at 46 includes
 4 process excitation frequencies including $0.5\omega_c$, ω_c and $1.5\omega_c$, and in response, the output y of
 5 the process also contains these frequencies. When the stationary oscillation condition is
 6 reached, the recorded/saved u' and y' are then used to compute the process 32 frequency
 7 response data at the frequencies $.5\omega$, ω_c and $1.5\omega_c$ using equations 2 and 3, which will now be
 8 described in detail. The auto-tuner 20 circuitry then uses the calculated process frequency
 9 response data to design the regulator 54. When the regulator 54 is determined, the switch 44 is
 10 activated to connect the regulator output 56 to the input 26 of the digital to analog converter
 11 28. At this point the regulator 54 responds to the regulator input e at 56, e being the difference
 12 between the digital reference r and digitized process output y' by applying a signal u' , which
 13 converted by digital to analog converter 28, provides the signal u to the processor to drive the
 14 processor to output a y at 34 corresponding to the reference r .

15 The procedure described above and with the following text as performed by the auto-
 16 tuner 20 can be implemented in a computer, the inputs being the digital reference r and the
 17 digitized process output y' , and the output being the digital control signal u' . In addition to the
 18 specific blocks noted within block 20, block 20 also represents the digital apparatus responsive
 19 to r and y' for calculating the process frequency response, and for calculating the required
 20 regulator parameters and setting the regulator as required. Such operations performed
 21 according to the equations disclosed will be understood by those skilled in the art.

22 As discussed above, Fig. 2 is an arrangement of apparatus for automatically estimating
 23 a process frequency response and designing a process regulator. The details of the process
 24 frequency estimating apparatus will now be described in detail.

25 Assume that initially, process 32 runs in open loop. Then a standard relay 50 and the
 26 novel parasitic relay 52 of the present invention are applied to process 32. This results in input
 27 and output time responses $u(t)$ and $y(t)$ of process 32. $u(t)$ and $y(t)$ are sampled at a sampling
 28 period of T and are recorded until stationary limit cycles of $u(t)$ and $y(t)$ have been reached.

29 The key of the present invention is the modified relay which consists of the standard

1 relay 50 and parasitic relay 52, as shown in Fig. 1. The standard relay operates as usual with
 2 the amplitude of the sampled output $u_1(k)$ being h , where $u_1(k)$ is the k -th sample of $u_1(t)$. It is
 3 well known that the standard relay can excite process 32 mainly at frequency ω_c . In order to
 4 provide excitation of the process 32 at frequencies other than ω_c , for use in control while
 5 maintaining the process output oscillation under such an arrangement, a parasitic relay 52 with
 6 output amplitude αh and twice the period of $u_1(k)$ is introduced and superimposed on $u_1(k)$.
 7 This implies that the output $u_2(k)$ of the parasitic relay 52 flip-flops once every period of
 8 oscillation in $u_1(k)$. The parasitic relay 52 output $u_2(k)$ is described by the following equations,

$$\begin{cases} u_2(0) = \alpha h; \\ u_2(k) = \alpha h \cdot \text{sign}(u_2(k-1)), \text{ if } u_1(k-1) > 0 \text{ and } u_1(k) < 0; \\ u_2(k) = u_2(k-1), \text{ otherwise;} \end{cases}$$

9 (1)

10 where α is a constant coefficient. α should be large enough to have sufficient stimulation on
 11 the process 32 and it should also be small enough such that the parasitic relay 52 will not
 12 change the period of oscillation generated by the main relay 50 too much. The preferred range
 13 of values for α is from 0.1~0.3. The output at 46 of the modified relay apparatus is thus given
 14 by $u(k) = u_1(k) + u_2(k)$, and is sent as the input to process 32.

15 In this way, the process is stimulated by two different excitations whose periods are T_c
 16 and $2T_c$ respectively. The digital input u' at 28 resulting from the modified relay apparatus is
 17 shown in Fig. 3A. The output y at 34 from the process 32 will appear as shown in Fig. 3B, and
 18 will reach a stationary oscillation with the period being $2T_c$. Due to two excitations in u , y
 19 consists of frequency components at

20 $\frac{2\pi}{T_c}, \frac{\pi}{T_c}$
 21 and their odd harmonics

$$\frac{6\pi}{T_c}, \frac{10\pi}{T_c}, \dots \text{and } \frac{3}{T_c}, \frac{5\pi}{T_c}, \dots \text{respectively.}$$

22

23

24

1 For a linear process 32, the process frequency response can be obtained by

$$G(j\omega_l) = \frac{\int_0^{2T_c} y_s(t) e^{-j\omega_l t} dt}{\int_0^{2T_c} u_s(t) e^{-j\omega_l t} dt}, \quad i = 1, 2, \dots,$$

2 (2)

$$\omega_l = \frac{(2l-1)2\pi}{2^l T_c}, \quad l = 0, 1,$$

3 where

4 are the basis frequency and its odd harmonic frequencies in u_s and y_s . u_s and y_s have a period
 5 ($2T_c$) related to the stationary oscillations of $u(k)$ and $y(k)$ respectively. $G(j\omega_l)$ in (2) can be
 6 computed using the FFT algorithm as

$$G(j\omega_l) = \frac{FFT(y_s)}{FFT(u_s)}$$

7 (3)

8

9 The spectrum analysis method used by the auto-tuner 20 in estimating the process frequency
 10 response gives a more accurate result than methods relying on the describing function. The
 11 method of the present invention employs the FFT only once and the required computation
 12 burden is modest. It can identify multiple points on the process frequency response from a
 13 single test using the modified relay of the present invention. Moreover, the method can be
 14 easily extended to find additional points on the frequency response. One can flip-flop the
 15 parasitic relay every 3rd or 4th period of the main oscillations generated by the standard relay
 16 and get other frequency points. One can also use more than one parasitic relay in a relay test
 17 and find more points on the frequency response in one relay test. To estimate the static gain of
 18 a process, either a small set-point change at reference may be introduced or a bias may be
 19 added to the modified relay.

20 In a realistic environment, the major concerns for any identification method are
 21 disturbance and noise. It should be noted that the identification method is unaffected by the
 22 step-like load disturbance w referred to in Fig. 1, which is a common case in practice. This can

1 easily be shown from (2) as

$$\begin{aligned}
 & \frac{\int_0^{2T_c} \tilde{y}_s(t) e^{-j\omega_i t} dt}{\int_0^{2T_c} \tilde{u}_s(t) e^{-j\omega_i t} dt} = \frac{\int_0^{2T_c} [y_s(t) + w(t)] e^{-j\omega_i t} dt}{\int_0^{2T_c} (u_s(t) + \bar{u}) e^{-j\omega_i t} dt} = \frac{\int_0^{2T_c} y_s(t) e^{-j\omega_i t} dt + w \int_0^{2T_c} e^{-j\omega_i t} dt}{\int_0^{2T_c} u_s(t) e^{-j\omega_i t} dt + \bar{u} \int_0^{2T_c} e^{-j\omega_i t} dt} \\
 & = \frac{\int_0^{2T_c} y_s(t) e^{-j\omega_i t} dt}{\int_0^{2T_c} u_s(t) e^{-j\omega_i t} dt} = G(j\omega_i), \quad i = 1, 2, \dots
 \end{aligned}
 \tag{4}$$

2
3
4 where ω_i has the same definition as in (2). The modified relay estimating apparatus of the
5 auto-tuner 20 of the present invention is also insensitive to noise. A preferred method of
6 reducing the noise in the process output $y(t)$ is by introducing a hysteresis in the standard relay
7 50. This is illustrated in Fig. 4. The width of hysteresis should be bigger than the noise band
8 and is usually chosen to be 2 times larger than the noise band. Filtering is another possibility.
9 To reduce the effect of noise further, especially in the case of large noise-to-signal ratios, the
10 estimator uses the average of the last 2-4 periods of oscillation as the stationary oscillation
11 period, depending on the noise level. With these anti-noise measures, the method of the
12 present invention can reject noise very effectively, and provide accurate frequency response
13 estimation at frequencies of $0.5\omega_c$, ω_c and $1.5\omega_c$. It should be also noted that a non-zero initial
14 condition of the process at the start of a relay test has no effect on the estimation because only
15 stationary oscillations u_s and y_s after transient are used in the estimation, where u_s and y_s are
16 independent of the initial condition.

17 **Example.** In order to test the method in a realistic environment, extensive real-time
18 relay testing was performed using the *Dual Process Simulator KI 100* from *KentRidge*
19 *Instruments, Singapore*. For illustration, a process was configured on the Simulator as

$$G(s) = \frac{1}{5s + 1} e^{-5s}$$

20
21 In the test, the standard relay amplitude was chosen as 0.5 and the parasitic relay height
22 was set to 20% $\times$ 0.5. Without additional noise, the noise-to-signal ratio of the inherent noise in

1 the test environment was measured at $N_1=0.025\%$, where N_1 is defined as

2

$$N_1 = \frac{\text{mean power spectrum density of noise}}{\text{mean power spectrum density of signal}}$$

3

4 or as $N_2 = 4\%$, where the noise-to-signal ratio N_2 is

$$N_2 = \frac{\text{mean (abs (noise))}}{\text{mean (abs (signal))}}$$

5

6 The identification error ERR is 2.5% , where

$$ERR = \max_i \left\{ \left| \frac{\hat{G}(j\omega_i) - G(j\omega_i)}{G(j\omega_i)} \right| \times 100\%, i = 1, 2, 3 \right\}$$

7

8 in which $G(j\omega_i)$ and $\hat{G}(j\omega_i)$ are the actual and the estimated process frequency responses
 9 respectively. To see noise effects, extra noise was introduced with the noise source in the
 10 Simulator. Time sequences of $y(t)$ and $u(t)$ in a relay test under $N_1=10\%$ ($N_2=31\%$) are shown
 11 in Figs. 3A and 3B. The first part of the test in Figs. 3A and 3B ($t=0\sim12$) was the “listening
 12 period”, in which the noise bands of $y(t)$ and $u(t)$ at the steady state were measured. Under this
 13 noise, the hysteresis was chosen as 0.3. With averaging 4 periods of stationary oscillations, the
 14 estimated frequency response points under this noise level are shown in Fig. 4.

15 Upon completion of the process frequency response estimation, the auto-tuner uses the
 16 estimated data to automatically design the regulator 54. This process will now be described in
 17 detail.

18 The three estimated points of frequency response are used for tuning the regulator 54 in
 19 Fig. 2. The three points are first converted into a second order plus dead time model with the
 20 following structure

$$\hat{G}(s) = \frac{e^{-sL}}{as^2 + bs + c},$$

21

(5)

1 where a , b , c and L are unknowns to be determined. It follows that

$$\hat{G}(j\omega_i) = \frac{1}{(j\omega_i)^2 a + j\omega_i b + c} e^{-j\omega_i L}, \quad i = 1, 2, 3. \quad (6)$$

2
3 The magnitude of both sides of (6) is taken as

$$[\omega_i^4 \ \omega_i^2 \ 1]\theta = \frac{1}{|G(j\omega_i)|^2}, \quad i = 1, 2, 3. \quad (7)$$

4
5
6 where

$$\theta = [a^2 \ b^2 \ -2ac \ c^2]^T.$$

7 Then,

$$[a \ b \ c] = [\sqrt{\theta_1} \ \sqrt{\theta_2 + 2\sqrt{\theta_1\theta_3}} \ \sqrt{\theta_3}]$$

8
9 In addition, the phase relation in (6) gives

$$\omega_i L = -\arg[G(j\omega_i)] - \tan^{-1}\left(\frac{b\omega_i}{c - a\omega_i^2}\right), \quad i = 1, 2, 3. \quad (8)$$

10
11
12 It will be clear to those skilled in the art that L can be obtained with the least squares
13 method.

14 With the model in (5) for the process available, a controller can be designed. A PID
15 controller can be written in the form:

$$K(s) = K\left(\frac{As^2 + B + C}{s}\right), \quad (9)$$

1 where

$$A = \frac{K_D}{k}, \quad B = \frac{K_P}{k}, \quad C = \frac{K_I}{k}$$

2

3 The controller zeros are chosen to cancel the model poles, i.e. $A=a$, $B=b$, $C=c$, where a , b and c
 4 are the model parameters determined from (7) and (8). Then the resultant open-loop transfer
 5 function $G(s)K(s)$ is approximated by

$$\hat{G}(s)k(s) = \frac{ke^{-sL}}{s}. \quad (10)$$

6

7 Its closed-loop poles can be selected from the root locus of the loop by assigning a
 8 proper value for k .

9

10 For the transfer function in (10), real or complex roots may be obtained for the closed-
 11 loop. In the method of the present invention, the controller zeros have been chosen to cancel
 12 the model poles. Exact cancellation is never expected since the process can be of any order
 13 while the model is only of second order. For highly oscillatory processes, it is possible that the
 14 uncanceled dynamics will drive the system to heavy oscillation, and hence it is reasonable not
 15 to create additional oscillatory dynamics by having complex close-loop poles. Real closed-
 16 loop poles are chosen for the system instead. On the other hand, for non-oscillatory or lightly
 17 oscillatory processes, the uncanceled dynamics will not bring the system to severe oscillation,
 18 and hence it is advisable to introduce some overshoot by selecting complex closed-loop poles
 19 so as to speed up the response.

20 For ease of presentation, some terms will now be defined that will be used in the
 21 following description. The equivalent time constant τ_o of a process is inversely proportional to
 22 its speed of response. For monotonic processes, the speed of response is reflected by the
 23 locations of the dominant poles. For oscillatory ones, it is related to the real part of the
 24 complex poles which determines the system attenuation and hence serves as a measure of the
 25 process speed. According to equivalent time principles, we have

26

$$\frac{1}{\tau_o} = \begin{cases} \frac{c}{\sqrt{b^2 - 4ac}}, & b^2 - 4ac \geq 0, \\ \frac{b}{2a} & b^2 - 4ac < 0 \end{cases} \quad (11)$$

where a , b and c are model parameters that can be obtained from (7). Another variable of interest to the design method is the damping ratio ζ_o of the open-loop plant which is defined as

$$\zeta_o = \begin{cases} \frac{b}{2\sqrt{ac}}, & b^2 - 4ac < 0, \\ 1, & b^2 - 4ac \geq 0 \end{cases} \quad (12)$$

with a , b and c being the model parameters. The solution [to the regulator design?] will now be presented by separating it into three cases.

Case I. $\zeta_o > 0.7071$ or $\frac{L}{\tau_o} < 0.15$ or $\frac{L}{\tau_o} > 1$.

Complex closed-loop poles on the root locus are chosen based on the reasoning given earlier. In order for a pair of the desired poles

$$s + -\omega_n \zeta_n \pm j\omega_n \sqrt{1 - \zeta_n^2}$$

where ζ_n is the closed-loop damping ratio, to be on the root locus for the system in (10), the phase condition

$$-\omega_n \sqrt{1 - \zeta_n^2} L - (\pi - \cos^{-1} \zeta_n) = -\pi \quad (13)$$

1 has to be satisfied, giving

$$\omega_n = \frac{\cos^{-1} \zeta_n}{L\sqrt{1-\zeta_n^2}}. \quad (14)$$

3 The magnitude condition then assigns the value of k to

$$k = \omega_n e^{-\omega_n L \zeta_n}. \quad (15)$$

6 Simulation results show that the feedback system gives satisfactory responses if closed-loop poles of damping ratio $\zeta_n = 0.7071$ are chosen. Substituting $\zeta_n = 0.7071$ into eq. (14) and (15) results in

$$k = \frac{0.5064}{L} \quad (16)$$

Case II. $\zeta_o \leq 0.7071$ or $0.15 \leq \frac{L}{\tau_o} \leq 1$.

10 In this case, we choose the closed-loop poles as real double poles on the root-locus. Its
11 location is selected to have a similar response speed to that of the open-loop one if it is
12 achievable, i.e. it is before the breakaway point. Otherwise the breakaway point will be used.
13 In order for

$$s_{1,2} = -\frac{l}{\tau_o}$$

14 to be on the root locus,

$$k = \frac{l}{\tau_o} e^{-\frac{L}{\tau_o}}. \quad (17)$$

17 To find the breakaway point, it is noted from eq. (14) that

$$\lim_{\zeta_n \rightarrow 1} \omega_n = \lim_{\zeta \rightarrow 1} \left(\frac{\cos^{-1} \zeta_n}{L \sqrt{1 - \zeta_n^2}} \right) = \frac{1}{L}$$

(18)

Hence, the breakaway point is

$$s_{1,2} = -\omega_n \zeta_n \pm j \omega_n \sqrt{1 - \zeta_n^2} = -\frac{1}{L}$$

and the corresponding gain is

$$k_{break} = -s e^{sL} \Big|_{s = -\frac{1}{L}} = \frac{1}{eL}$$

(19)

Therefore, the value of k should be taken as

$$k = \min \left\{ \frac{1}{\tau_o} e^{-\frac{L}{\tau_o}}, \frac{1}{eL} \right\}$$

(20)

Case III. $\frac{L}{\tau_o} > 2$.

Processes with long dead time tend to produce very slow responses. In fact, the single-loop structure is inadequate in handling such processes. One possible solution is to apply the Internal Model Control (IMC) strategy to compensate for the long dead time. The controller structure is shown in Fig. 6. The model of the process is denoted as $\hat{G}(s)$ in the figure, and

1

$$\hat{G}(s) = \frac{e^{-sL}}{as^2 + bs + c}$$

2

3 where the parameters a, b, c and L are given in (7)-(8). The controller is now modified to

$$K(s) = \frac{as^2 + bs + c}{\left(\frac{1}{\alpha}s\tau_o + 1\right)^2}$$

4

(21)

5

6 where τ_o is the equivalent time constant defined in (11) and α is a design parameter to adjust
 7 the speed of response. The larger the value of α , the faster the response becomes. The typical
 8 range of α is $\alpha \in [0.5, 5]$. If the model is exact, the system will have an equivalent transfer
 9 function of

$$H_{yr}(s) = \frac{1}{\left(\frac{1}{\alpha}s\tau_o + 1\right)^2} e^{-sL}$$

10

(22)

11

12 The dead time is therefore brought out of the loop, and the speed can be chosen to be as
 13 fast as desired by selecting a large α value. Since this inevitably increases the size of the
 14 control effort, the limit is set by the saturation of the control signal and the accuracy of the
 15 model.

16 When the regulator 54 in Fig. 2 is designed, the switch 44 is connected to the lower
 17 node 56, and the auto-tuning is completed.

18

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21

1 **Example** Consider the following long dead time process given by the transfer function

$$G(s) = \frac{1}{(s^2 + 2s + 3)(s + 3)} e^{-2s}$$

2

3 The model of the process is found to be

$$\hat{G}(s) = \frac{1}{2.921s^2 + 6.536s + 9.031} e^{-2.289s}$$

4

5 The PID controller is designed to be

$$K(s) = 1.446 + \frac{1.998}{s} + 0.646s.$$

6

7 Since the process is dead time dominated, the IMC approach can be employed to speed up
8 the response. Choosing the value of $\alpha = 2$ in eq. (21), the controller is computed to be

$$K(s) = \frac{2.921s^2 + 6.536s + 9.031}{\left(\frac{1}{2} \times 0.894s + 1\right)^2}$$

9

10 The control performance of the IMC and PID controllers is also shown in Fig. 7.

11 The performance of the IMC controller is an improvement over the PID controller. The
12 speed is much increased, and the settling time and overshoot are significantly reduced.

13 Although a preferred embodiment of the present invention has been described above, it
14 will be appreciated that certain alterations and modifications thereof will be apparent to those
15 skilled in the art. It is therefore intended that the appended claims be interpreted as covering
16 all such alterations and modifications as fall within the true spirit and scope of the invention.

17 What is claimed is:

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CLAIMS

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1. An apparatus for estimating the frequency response of a linear stable process comprising:
 - a. signal excitation means for providing a process control signal including a process critical frequency and a plurality of signals of different frequencies; and
 - b. identification means based on the fast fourier transform for estimating a plurality of points on said frequency response of said process from input and output stationary oscillations of said process.
2. An apparatus as recited in claim 1 wherein said signal excitation means includes
 - a. a standard relay for providing said process critical frequency; and
 - b. a parasitic relay for providing said plurality of signals.
3. An apparatus as recited in claim 2 wherein said parasitic relay is set to flip-flop every k -th period of said standard relay, where k is selected from the group including .5 and positive integers.
4. An apparatus as recited in claim 2 wherein an output of said standard relay and said parasitic relay can be biased to output a signal indicative of an estimate of a static gain of said process.
5. An apparatus for tuning a regulator in a controller of a process comprising:
 - a. means for estimating the frequency response of said process including
 - i. signal excitation means for providing a process control signal including a process critical frequency, and a plurality of signals of different frequencies; and
 - ii. identification means based on the fast fourier transform for estimating a plurality of points on said frequency response of said process from input and output stationary oscillations of said process; and
 - b. regulator design means for calculating parameters of a regulator, said design means using said plurality of points on said frequency response of said process.
6. An apparatus as recited in claim 5 wherein said regulator is a PID regulator.
7. An apparatus as recited in claim 1 wherein said regulator is an IMD regulator.
8. An apparatus as recited in claim 5 wherein said regulator design means includes
 - a. means for canceling model poles with controller zeros; and

- 3 b. means for computation of a constant gain of said controller; wherein said means for
4 canceling and means for computation are used to generate closed-loop poles for a
5 controller feedback system.
- 1 9. An apparatus as recited in claim 5 wherein said signal excitation means include
2 a. a standard relay for providing said process critical frequency; and
3 b. a parasitic relay for providing said plurality of signals.
- 1 10. An apparatus as recited in claim 9 herein said parasitic relay is set to flip-flop every k -th
2 period of said standard relay, where k is selected from the group including 0.5 and positive
3 integers.
- 1 11. An apparatus is recited in claim 9 wherein an output of said standard relay and said
2 parasitic relay can be biased to output a signal indicative of an estimate of a static gain of
3 said process.
- 1 12. An apparatus as recited in claim 1 wherein said signal excitation means includes
2 a. first programmed computer means responsive to a digitized reference input signal r and
3 a digitized process output signal to provide a digital excitation means output signal;
4 b. digital to analog converter means for converting said digitized process output signal to
5 a process excitation signal; and
6 c. analog to digital converter means for converting a process output signal to said
7 digitized process output signal.
- 1 13. An apparatus as recited in claim 5 wherein said signal excitation means includes
2 a. first programmed computer means responsive to a digitized reference input signal and a
3 digitized process output signal to provide a digital excitation means output signal;
4 b. digital to analog converter means for converting said digitized process output signal to
5 a process excitation signal; and
6 c. analog to digital converter means for converting a process output signal to said
7 digitized processed output signal.
- 1 14. An apparatus as recited in claim 13 wherein said identification means includes second
2 programmed computer means.
- 1 15. An apparatus as recited in claim 14 wherein said regulator design means includes

- 2 a. third programmed computer means for performing said calculating, wherein said
3 regulator is responsive to said digitized reference input signal and said digitized process
4 output signal to provide a digitized regulator means output signal.
- 1 16. An apparatus as recited in claim 15 further comprising digital to analog converter means
2 for converting said digitized regulator means output signal to a process control signal.