Advanced Control Systems

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Module No. # 04
Design of Controllers
Lecture No. # 02

Design of Controllers for the Advanced Smith Predictor

Welcome to the lecture titled design of controllers for the advanced smith predictor. In this lecture, we shall discuss about the design technique for the controllers of the advanced smith predictor proposed by Majhi et al. In our last lecture, we have seen the structure for the Majhi et al's modified smith predictor and, controller; there were three controllers in the structure.

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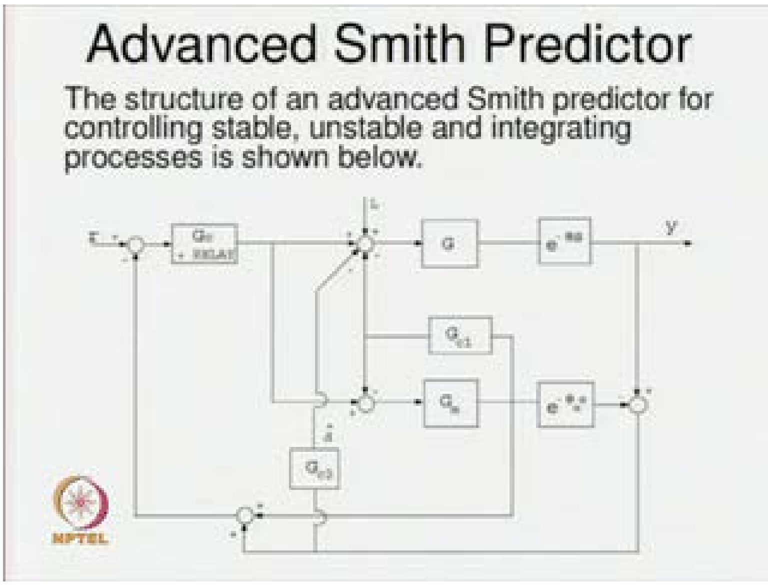

Now the structure has been reproduce, once more where the servo controller or servo tracking controller G c is also connected in parallel with a relay. The main objective of the relay is to identify the dynamics of the real time process. So, once the dynamics of the real time process is obtained, then that is now shown by the transfer function model G m e to the power minus theta m s.

The advanced smith predictor has got the three controllers: G c, G c 1, and G c 2; initially a relay is connected in parallel with the controller G c for inducing limit cycle signal, and using the limit cycle data, and the state space based exact analytical expression, we have to obtain and earlier for estimations of parameters of a (( )) parameters of a transfer function model, for the dynamics of a real time system. The G m e to the power minus theta m s or the time delay theta m, and parameters of delay free part of the transfer function model G m are obtained.

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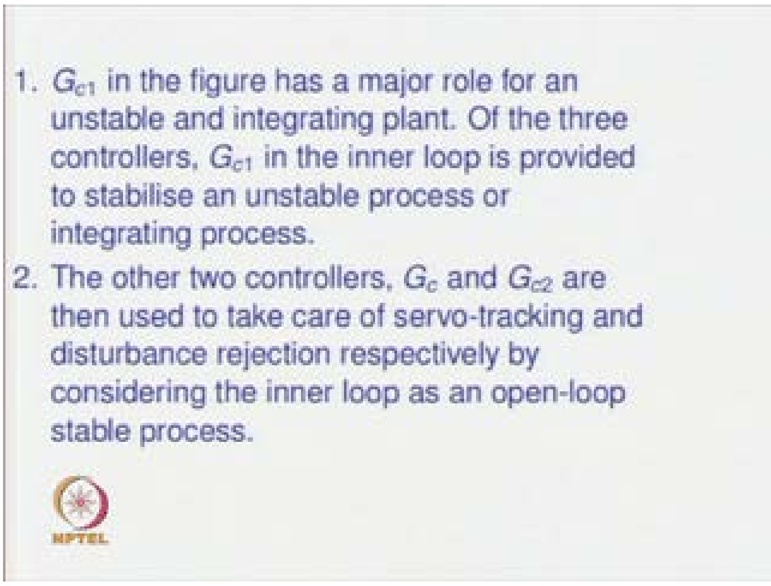

Next, G c 1 in the figure has a major role for an unstable, and integrating plant. The G c 1 stabilises, open loop unstable integrating plants of the three controllers G c 1 in the inner loop is provided to stabilise open loop unstable process or integrating process. Also it can relocate the positions of poles of stable processes. The other two controllers as I have said earlier G c, and G c 2 are used to take care of servo tracking, and disturbance rejection by considering the inner loop as an open loop stable process. What we mean by that with the help of G c 1, we are getting some modified process, and we assume that modified process to be a stable one; assuming that we are as if dealing with a stable process G c and G c 2 are designed to improve upon servo tracking, and disturbance performances

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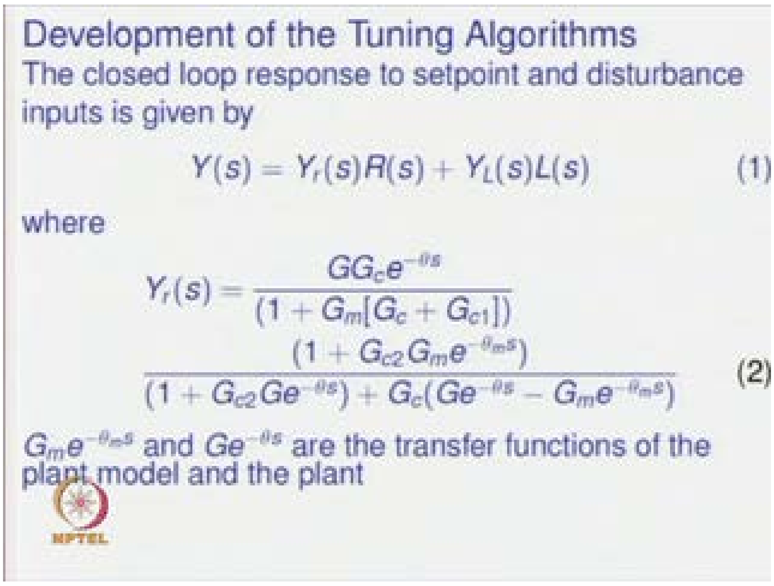

Now, how the tuning algorithms or the formulae for or them technique method for designing the three controllers are done will be explained now. The closed loop response to set point, and disturbance inputs is given by Y(S) is upon equal to Y r(S) R(S) plus Y L(s) L(s), where Y r(S) is the transfer function which is given by GG c e to the power minus theta s upon 1 plus G m times G c plus G c 1 times 1 plus G c 2 G m e to the power minus theta m s upon 1 plus G c 2 g e to the power minus theta s plus G c times G e to the power minus (( )) theta s minus G m e to the power minus theta m s.

The detail derivation is again avoided here, already we have discussed that with the use of signal flow graph; it is not difficult to obtain this expression Y r(S). Now, G m e to the power minus theta m s, and G e to the power minus theta s are the transfer functions of the plant model, and the plant. And please keep in mind G m is the delay free part of the transfer function model.

As is evident from this denominator under model matching condition, as you see in equation number; one part of the denominator will get eliminated or removed, thus giving us a relatively simpler expression for Y r(S).

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$$Y_L(s) = \frac{Ge^{-\theta s}}{(1 + G_m[G_c + G_{c1}])} \frac{(1 + G_m[G_c + G_{c1} - G_c e^{-\theta_m s}])}{(1 + G_{c2} G e^{-\theta s}) + G_c(Ge^{-\theta s} - G_m e^{-\theta_m s})} \qquad (3)$$

Based on the assumption that $G_m = G$ and $\theta_m = \theta$, Equations (2) and (3) reduce to

$$Y_r(s) = \frac{GG_c e^{-\theta s}}{1 + G(G_c + G_{c1})} \qquad (4)$$

$$Y(s) = \frac{Ge^{-\theta s}}{1 + G(G_c + G_{c1})} \frac{1 + G(G_c + G_{c1}) - GG_c e^{-\theta s}}{1 + GG_{c2} e^{-\theta s}} \qquad (5)$$

Similarly, Y L(S) the response or the transfer function between the output to the load disturbance can be given as Y L(S) is equal to this much, as shown in equation number three. Now, based on the assumption that G m is equal to G, and theta m is equal to theta; equations two and three. What is equation two? You see the equation two here. So, equation two, and three reduce to Y r(S) is equal to GG c e to the power minus theta s upon 1 plus G times G c plus G c 1, and Y L(S) is equal to G e to the power minus theta s upon one plus G times G c plus G c 1 into 1 plus GG c plus G c 1 minus GG c e to the power minus theta s upon one plus GG c 2 e to the power minus theta s.

Please look at equation number 4 carefully, the denominator is now free from any time delay term. And that particularly facilitates in designing controllers for the advanced smith predictor control structure. Also you please see that equation number 4, and five shows us that we have got two different denominators; the denominators of equation 4 is different from that of the equation number five. The equation 4 does not have a term like 1 plus GG c 2 e to the power minus theta s in the denominator. Therefore, the two responses; the set set point response, and the disturbance response are decoupled from each other. What we mean by decoupling of the (( )) responses; one response does not affect the other or indirectly speaking, the controller you have designed for set point or servo tracking will not get effected by that from the controller, you have designed for disturbance rejection.

Also, when G c 2 equal to zero. That means, when we do not have the disturbance response controller; particularly for unstable process Y L(S) will give unbounded output. That you can make out provided you x substitute GG c, G c 1 by some transfer functions.

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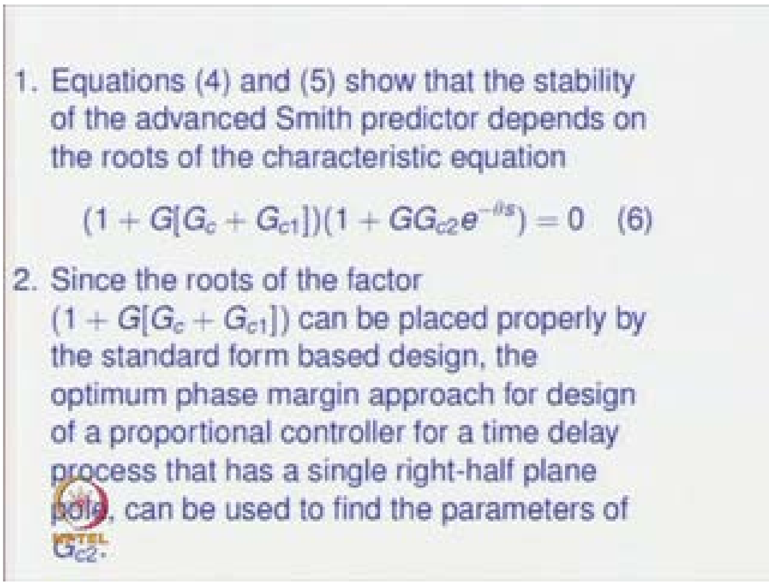

Equations equations 4 and five again show that, the stability of the advanced smith predictor depends on the roots of the characteristic equation, 1 plus G times G c plus G c 1 times 1 plus G G c 2 e to the power minus theta s is equal to zero. Particularly, if you look carefully equation number five, it has the denominator having the characteristic polynomial given over here or the characteristic equation given in equation number six.

Earlier we designed controller G c, and G c 1 based on equation number 4; since the roots of the factor 1 plus GG c plus G c 1 can be placed properly by by the standard form based design. Which we are going to discuss, after some time the optimum phase margin approach for design of a proportional controller, for a time delay process that has a single right half plane pole can be used to find the parameter of G c 2. So, basically equation number six has got three controllers, and those are G c, G c 1, and G c 2; design of G c, and G c 1 are done with the help of standard form. Whereas, G c 2 is obtained using some optimum phase margin criteria.

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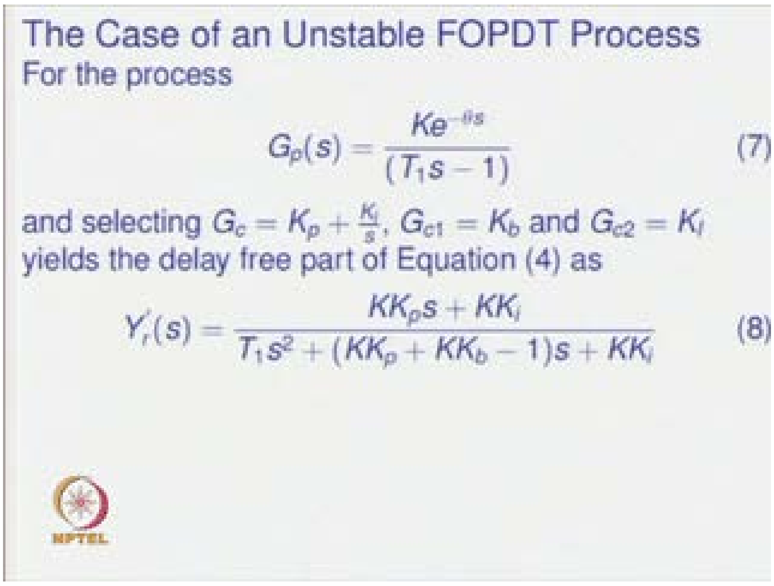

Consider the case of an unstable first order plus dead time process. The process can be given by G p(S) is equal to K e to the power minus theta S upon T 1 S minus 1. Now, I will design the controllers: three controllers for this unstable first order plus dead time process.

Let us select the form of the controller G c, as a p i controller having G c equal to K p plus K i upon S. Now, G c 1 the stabilizing controller is a proportional controller given by G c 1 is equal to K b, and G c 2 is equal to K l; please keep in mind this is not K i that is G c 2 is equal to K l. Now, when you substitute G c, G c, 1 G c 2, and G c, and G p, then the delay free part of equation 4, the delay free part of equation 4; please look at the delay free part of equation 4 means e to the power minus theta S will not be there. We will have GG c upon 1 plus G times G c plus G c 1 will yield an expression of the form Y r dash S is equal to K K p S plus K K i divided by T 1 S square plus K K p plus K K b minus 1 times S plus K K i. Now, how to obtain this in the standard form, if you divide the numerator of equation 8, and the denominator of equation 8 by the term K K i.

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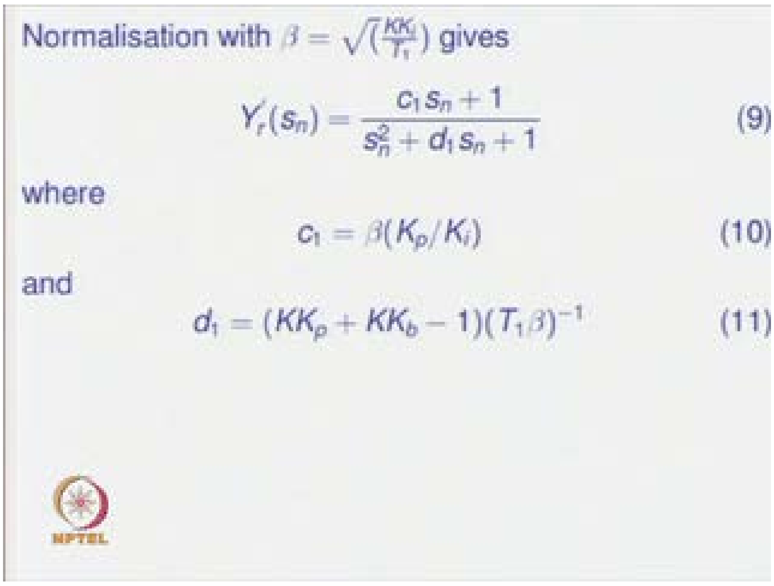

Then you will get another expression which can ultimately be expressed in the normalized form as Y r dash S n is equal to C 1 S n plus one divided by S n square plus d 1 S n plus 1. Where C 1 is equal to beta times K p divided by K I, and d 1 is equal to K K p plus K K b minus 1 times T 1 beta inverse. So, this is how, we have obtained a standard form, we have obtained a standard transfer function Y r dash S n with the help of the normalization with beta equal to square root of K K i K i upon T 1.

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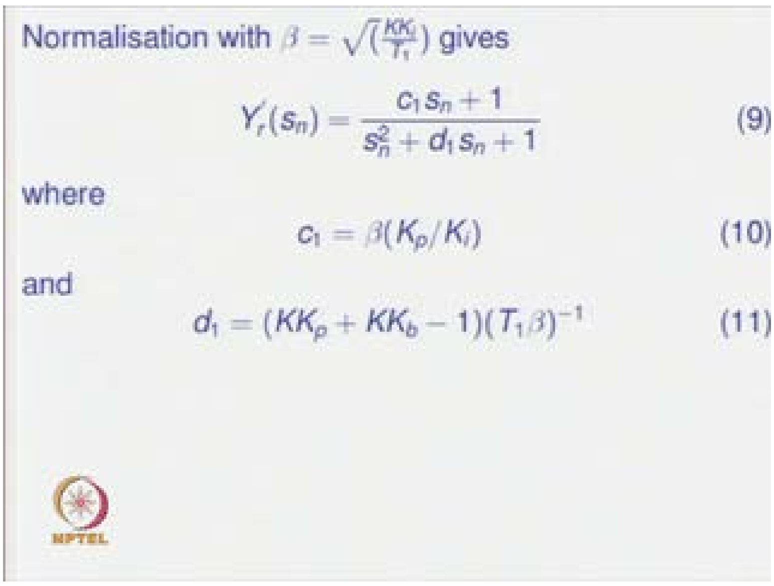

So, basically the dynamics of equation number 8, and equation number nine are not different from each other. Of course, there is a scaling factor, because you are getting a normalized frequency now in place of the normal frequency of operation. So, it will be nearly scaling of the responses, you will have corresponding to this normalized transfer function.

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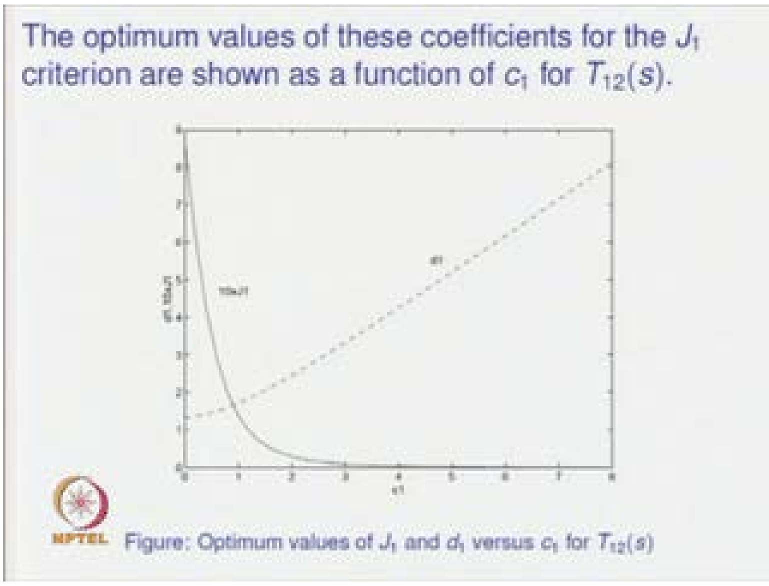

Now, we have got the optimum values for the coefficient of the standard transfer function d 1 and c 1. So, the standard transfer function given by equation number nine, has got two coefficients. One is c 1, and another is d 1; one can find optimum values of d 1 corresponding to certain c 1. So, for every c 1 there exists some optimum d 1, and with the help of some optimization function or with the help of some performance index, particularly the integral square time error criterion that is given by J 1. It is possible to get optimum values for d 1 for different values of c 1.

So, these plot shows the optimum values for d 1 given by the dotted line for different values of c 1 given by the x axis. So, c 1 is varying from 0 to 8, and d 1 is having different values when c 1 assumes values from 0 to 8. Also this figure shows the J 1 value, when the ISTE integral square time error criterion optimization is used to find the values for coefficients d 1 and c 1. So, this figure basically shows the optimum values of J 1, and d 1 versus c 1 for some transfer function T 1 2(S), what we mean by T 1 2(S), Y r dash(S) n is equal to T 1 two S is equal to c 1 S n plus 1 divided by S n square plus d 1 S n plus 1.

Now, what benefit we get from getting the delay free part expressed in this standard form, then we can find the values d 1 corresponding to or for certain c 1, and we can make use of beta c 1, d 1 to estimate model parameters parameters of the controllers G c, G c 1, and G c 2.

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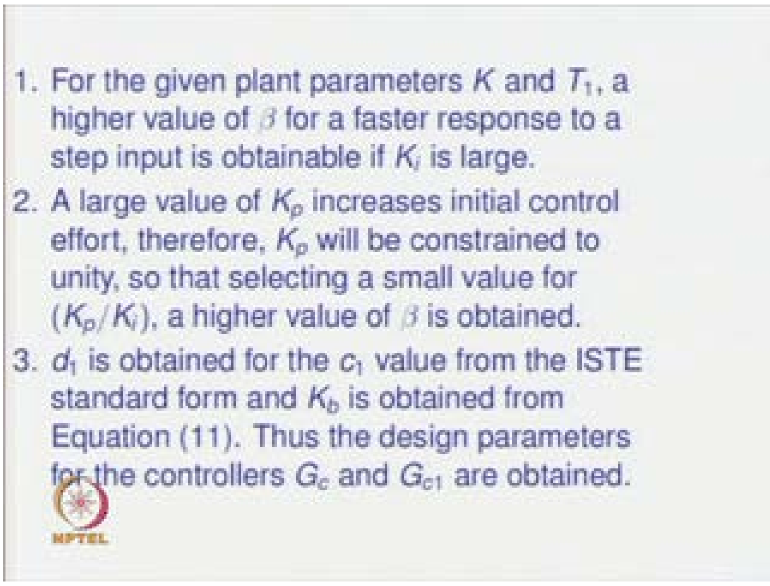

For the given plant parameter K, and T 1; we have got the plant dynamics given by equation 7, please keep in mind the plant parameters are k theta and T 1. So, for given k theta, and T 1 particularly for the given plant parameters K, and T 1; a higher value of beta results in a faster response to a step input. If K i is large, if k i is large - a large value of K p increases initial control effort, as you know higher proportional gain not only increases the control effort rather can leads to instable or unstable operation or instability in the closed loop system, for that we will constraint the K p value to unity.

So, that selecting a small value for K p upon K i or selecting a small value for K p upon K i or indirectly speaking one upon K i, a higher value of beta is obtained. Then d 1 is obtained for the c 1 value from the ISTE standard form, and K b is obtained from equation 11. So, 11 has got unknowns like K p and K b, because all other things are known. And of course, beta beta is or we are choosing beta, that way since K p has been constant to 1, the only unknown unknown that is in equation number 11 is K b, which can be estimated easily provided, we have certain value for d 1 - provided d 1 is known.

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The controller G_{c2}, whose prime job is to reject unwanted load disturbances, is designed on the basis of stabilisation of the characteristic equation

$$1 + GG_{c2}e^{-\theta s} = 1 + \frac{KK_l e^{-\theta s}}{T_1 s - 1} = 0 \qquad (12)$$

Based on the optimum phase margin criterion for the stabilisation of an unstable FOPDT process, it is easy to obtain

$$K_l = \sqrt{\frac{T_1}{\theta K^2}} \qquad (13)$$

with the constraint $\theta / T_1 < 1$.

Now, using these the steps has been explained now the way you should proceed with to design the controller parameters. The controller G c 2 prior to that what we have done, the steps for designing G c, and G c 1 are given in the steps 1, 2, 3 two. The steps for designing G c, and G c 1 are given here, whereas for designing the remaining controller G c 2, we have to reset to some other technique. The controller G c 2 which job is to reject load disturbances is designed on the basis of stabilization of the characteristic equation 1 plus GG c 2 e to the power minus theta S.

Now, substitution of G, and expression for G c 2 gives us an expression 12 or equation 12 which has got the expression 1 plus KK l e to the power minus theta S. Please keep in mind in the upper in the numerator of this expression, you have got K l not K i 1 plus KK l e to the power minus theta S upon T 1 S minus 1 is equal to 0. Now, based on the (( )) optimum phase margin criterion, now how can you find the optimum phase margin criterion; how how can you find use the optimum phase margin criterion here.

Again get it this expression expressed in the form of the frequency domain expression, and then you can find the phase angle of this function - whole function 1 plus KK l e to the power minus J omega theta upon J omega T 1 minus 1. Then optimized that phase angle with respective to the frequency, omega, and then you will get a condition, and that condition will be obviously, K l is equal to square root of T 1 upon theta K square.

So, the equation in 12 will give will have optimum or maximum phase angle provided K l is equal to square root of T 1 upon theta K square; it is not difficult to obtain this expression 13. So, based on the optimum phase margin criterion for the stabilization of unstable first order plus dead time process, we obtain the parameter of the controller G c 2, but this controller is designed with a constraint given by theta upon T 1 less than 1. The normalized dead time of the unstable first order plus dead time process has to be less than one; if it is not so, then the formula given in equation 13, may not be correct.

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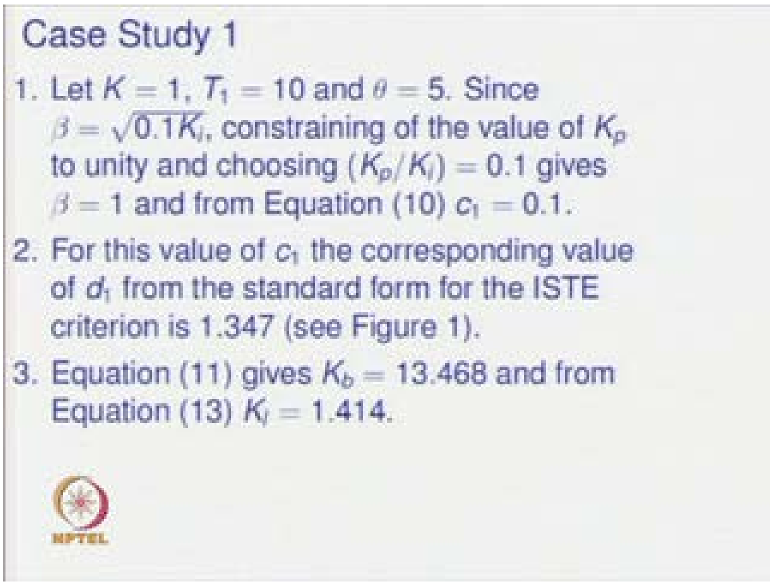

Let us go to case study one, where we will have we consider a an unstable first order plus dead time system with steady state gain K equal to 1, time constant T 1 equal to 10, and time delay theta equal to 5. The normalized dead time theta upon T 1 is 5 upon 10 is equal to 0.5, which is less than 1. Therefore, there will be no difficulties in designing the controller G c 2. Since beta is equal to given by square root of 0.1 times K i, how do you find this one; we know beta is equal to K K i divided by T 1. So, when K equal to 1, T 1 equal to 10, and K p equal to 1; then, beta becomes square root of 0.1 times K i. Now, constraining K p to unity, and choosing K p upon K i as 0.1 gives beta equal to 1, we are choosing the value of K i indirectly. So, that way the controller G c, which has got parameters K p, and K i has already been designed with proper choices for K p and K i.

So, this shows that k i is equal to 10, which gives us ultimately beta is equal to 1, and from equation 10, we get c 1 equal to 0.1, because you look at equation 10; c 1 is beta times K p upon K i, and since beta equal to one K p upon K i is 0.1. Therefore, c 1 also will be equal to 0.1. So, thus we get c 1 is equal to 0.1. Now, for these value of c 1, the corresponding value of d 1 from the figure, which figure from the standard form figure for the standard form from this figure. So, when c 1 equal to 0.1, we do get exact value of d 1 - optimum value of d 1 as the value of d 1 as 1.347. So, d 1 is equal to 1.347.

Now, equation 11 can be used to estimate the unknown parameter of the controller G c 1 which is nothing but K b is equal to 13.468, and again from equation 13. We obtain K l the controller that improves the disturbance rejection is equal to 1.414; thus all the three controllers are designed, where the p i controller has got the parameters K p equal to 1, K i is equal to 10. And now the second controller G c 1 is equal to having value 13.468, and G c 3 is having value one 1.414.

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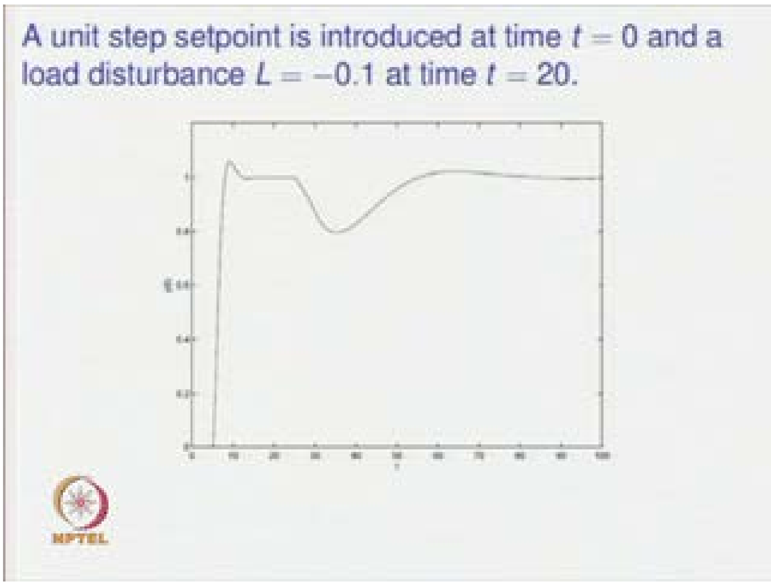

So, when this controller parameters are put, and simulation is carried out simulation of the modified smith predictor structure with the unstable first order plus dead time unstable process is carried out. We do get the response from a unit step set point input, that is introduced at time t equal to zero, and a load disturbance L of magnitude minus 0.1 at time t equal to 20. So, we do get responses corresponding to the set point and disturbance inputs, and the responses are shown over here.

Here, we see that we have got a quite good acceptable time response for the controlled system or the for the closed loop system. Why it is acceptable, if I look at carefully the plot, the over shoot is not beyond 10 percent; it is rather quite low, the over shoot is less, the raise time is high. The settling settling time is also its small. So, we have got all desired time domain performance parameters from the closed loop system. And of course, the disturbance response is not satisfactory, not so satisfactory as it is for the set point input response.

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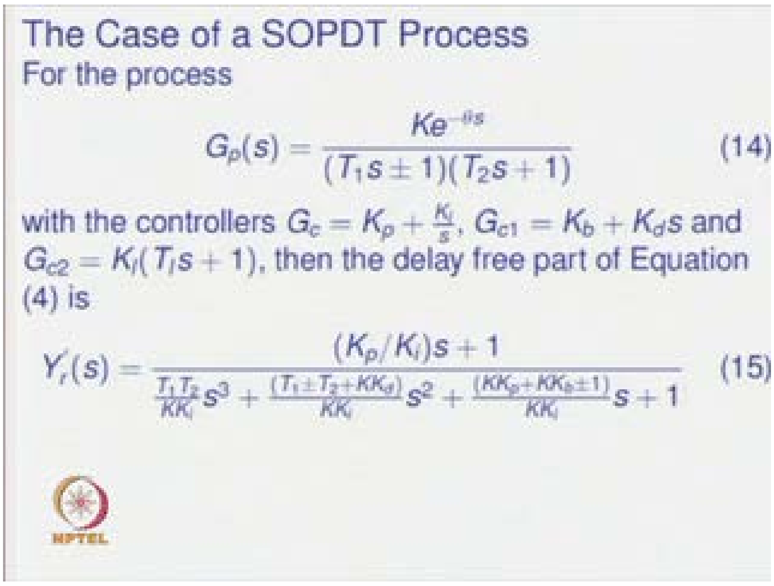

Let us go to the case of a second order plus dead time process, we will attempt to design the three controllers G c, G c 1, and G c 2 for this second order plus dead time process, where we can have stable or unstable poles with with the controllers choice of G c is equal to K p plus K i upon S, G c 1 is equal to K b plus K d S, please see the difference. Now, we are using the stabilizing controller as a PD controller.

So, the stabilizing controller G c 1 has got derivative action as well. So, it assumes the form now, G c 1 is equal to K b plus K d S, and the disturbance rejection controller takes the form of G c 2 is equal to K l times T l S plus 1. Now, the delay free part of equation 4, Y r dash S can be written as K p upon K i S plus one divided by T 1 T 2 upon K K i S cubed plus T 1 plus minus T 2 plus KK d upon K K i S square plus K K p plus K K b plus minus 1 divided by K K i S plus 1. How do you get Y r dash upon S, Y r dash upon S is given by the delay free part is given by the ratio of GG c upon 1 plus G times G c plus G c 1. When you substitute those GG c, and G c 1 then the delay free part becomes this one as shown in equation number 15.

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Normalisation with $\beta = (\frac{KK_i}{T_1 T_2})^{1/3}$ gives

$$Y_r'(s_n) = \frac{c_1 s_n + 1}{s_n^3 + d_2 s_n^2 + d_1 s_n + 1} \quad (16)$$

where

$$c_1 = (K_p/K_i)\beta \quad (17)$$

$$d_2 = \frac{(T_1 \pm T_2 + KK_d)}{T_1 T_2 \beta} \quad (18)$$

$$d_1 = \frac{(KK_p + KK_b \pm 1)}{T_1 T_2 \beta^2} \quad (19)$$

Now, normalization with beta is equal to K K i upon T 1 T 2 cube root gives the standard third order transfer function Y r dash S n is equal to C1 S n plus 1 divided by S n cube plus d 2 S n square plus d 1 S n plus one. Where C 1 is equal to K p upon K i times beta d 2 is equal to T 1 plus minus T 2 plus KK d upon T 1 T 2 beta, and d 1 is equal to K K p plus K K b plus minus 1 divided by T 1 T 2 beta square. So, 16, 17, 18, 19 gives us the normalized third order transfer function involving three coefficients; and the coefficients are C 1, d 2, d 1. Like the earlier case with the optimization of IST ISTE criterion or with the use of ISTE criterion with the optimization of the function, it is possible to obtain optimum values for d 2 and d 1 coefficients with respect to C 1 or for various C 1.

How to obtain is how to obtain optimum values for the coefficients for different order of transfer functions have been already discussed in one of our lecture, you need not worry. Simply, you have to transport the final plot, you have for either second order or third order standard transfer function.

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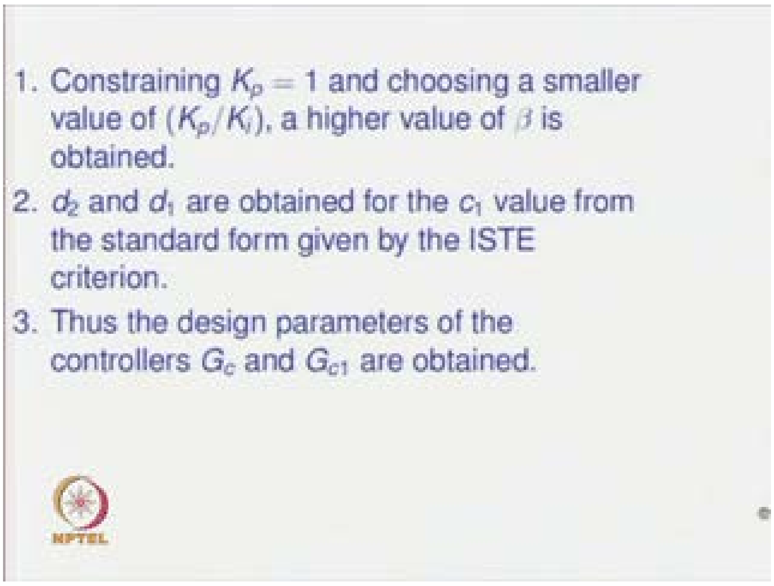

Now, I will explain you the procedure the way we will design the three three controllers. So, initially constraining K p to 1, again do we constraint K p to 1 to reduce the control effort. And choosing a smaller value of K p upon K 1 K i, a higher value of beta is obtained. I believe you are getting this concept why we are going for a higher value of beta, now d 2 and d 1 are obtained for the C 1 value from the standard form given by the ISTE criterion.

So, I am not going to show you further one plot. Now, directly the d 2, and d 1 values for a given c 1 will be used in our further analysis. Thus the design parameters of the controllers G c, and G c 1 are obtained using what using of course, 17, 18 and 19, and with the choice of a smaller K p upon K I.

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$$1 + GG_{c2}e^{-\theta s} = 1 + \frac{KK_l(T_l s + 1)e^{-\theta s}}{(T_1 s \pm 1)(T_2 s + 1)} = 0 \qquad (20)$$

Since for a stable SOPDT process, $G_{c2} = 0$, Equation (20) is therefore analysed for an unstable SOPDT process only. Letting $T_l = T_2$ to cancel a pole with the zero in Equation (20) gives a characteristic equation of the form shown in Equation (12). Again the optimum value of K_l is obtained by the optimum phase margin criterion given as

$$K_l = \sqrt{\frac{T_1}{\theta K^2}} \qquad (21)$$

Again, if you look at the denominator of the expression for y upon l S, as I had earlier 1 plus G G c 2 e to the power minus theta S. This is to be stabilized, when G is having unstable pole; for stabilization what you have to do is simple technique can be employed, when you write the expression in detail which gives us now 1 plus K K l times T l s plus 1 e to the power minus theta S upon T 1 S plus minus 1 times T 2 S plus 1 is equal to zero; with the choice of T l is equal to T 2, one term in the numerator and denominator will get cancelled.

So, cancel T l S plus 1, and T 2 S plus 1 with the choice of of course, T 2 is equal to T l or T l is equal to T 2. Then it will render again us a characteristic equation of the form 1 plus K K l e to the power minus theta S upon T 1 S plus minus 1. For stable processes you need not worry, we do not use the controller G c 2. So, the parameters K l, T i, T l are zero, but for unstable process what will happen, we will get the characteristic equation now in the form of 1 plus K K l divided by T 1 S minus 1 is equal to 0. Then the way the parameter K l is obtained is already explained to you earlier, and that has been given in equation number 12. So, the optimum value of K l given by this expression K l is equal to square root of T 1 divided by theta K square will be used in this study as well.

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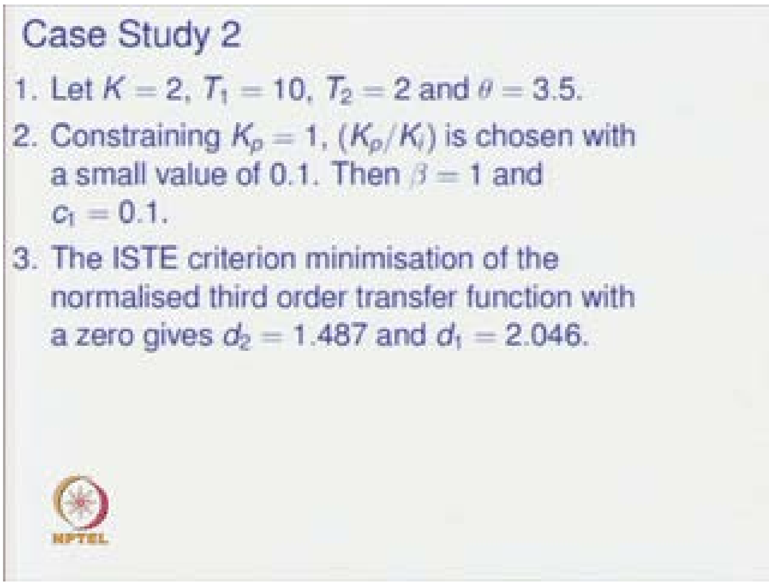

Now, I will go to a specific transfer function of the process which is having the steady state gain to the time constant T 1 as 10, the second time constant T 2 as 2, and theta equal to 3.5. Now, constraining K p equal to 1, and K p upon K i is equal to 0.1, then beta becomes 1. And consequently C 1 becomes 0.1; then corresponding to this C 1 we do get optimum values for the coefficients d 2 and d 1 which are found to be 1.487, and 2.046 respectively. Then we shall make use of the expression 16, 17, 18, as I have said you sorry, 17, 18 and 19 to estimate the unknown parameters of the transfer function model.

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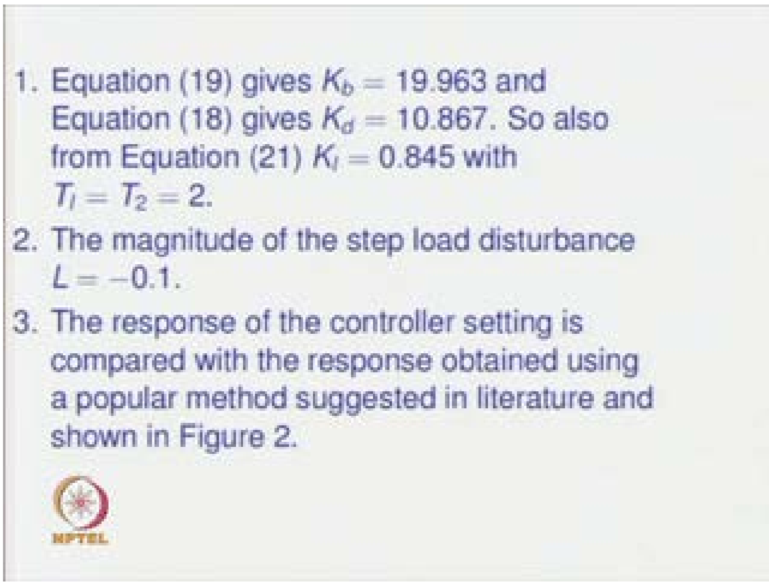

Now equation 19 gives K b is equal to 19.96, and equation 18 gives K d is equal to 10.867. So, also from equation 21, K l is equal to 0.845 with of course, with the choice of T l is equal to T 2 is equal to 2. Then only there will be pole zero cancellations, and ultimately rendering the second order expression into the first order thereby by enabling us to make use of the optimum phase margin criterion to find the parameter K l. So, the parameters of the controller G c 1 are K b and k d, the parameters of the controller G c 2 are K l and T l are thus obtained. The magnitude of the step load disturbance l is now minus 0.1, and the response of the controller setting is compared with the response obtained using a popular method suggested in the literature, and shown in the next figure.

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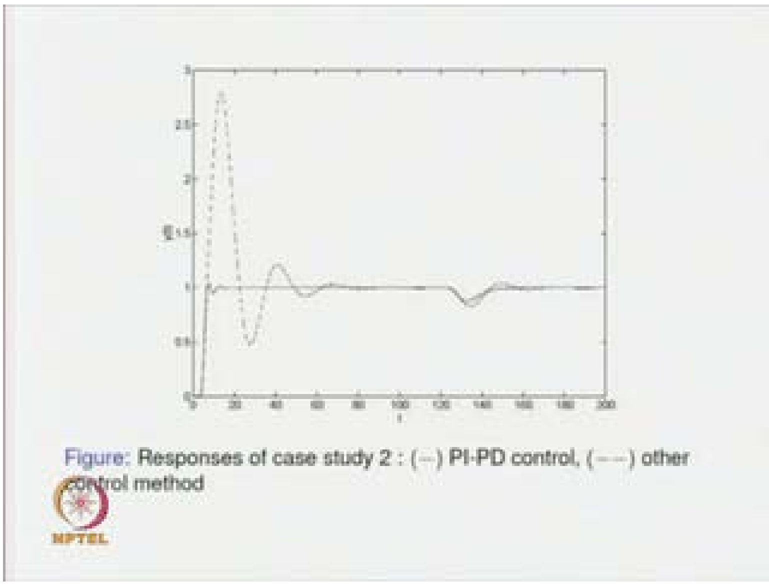

Figure: Responses of case study 2 : (–) PI-PD control, (– –) other control method

Now, in this case study two the solid line shows the responses obtained from the advanced smith predictor, and controller by Majhi et al. Whereas the dotted one shown in this figure is obtained by some popular smith predictor, and controller method. The responses show that the design method, and the advance smith predictor controller proposed by Majhi et al are far superior to that of the other smith predictor controllers, as far as controlling an unstable second order plus dead time process is concerned.

So, as far as controlling an unstable second order plus dead time process is concerned. This the beauty of the advance smith predictor controller, but the control design technique is not complicated; it is it is quite straight forward. You to have very limited number of analytical expressions, and make use of those simple expressions can yield all the parameters of the controllers in the advance smith predictor controller.

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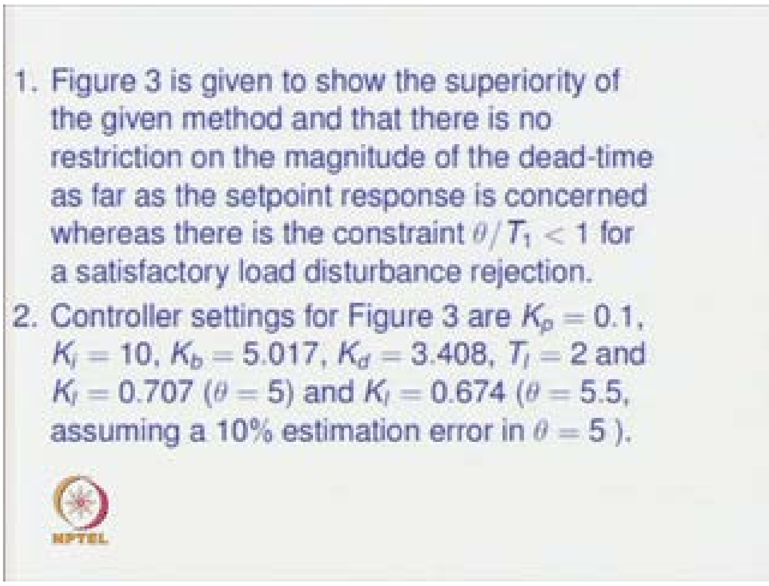

Now, another figure is shown here, the description about the figure let me give now; another figure is given to show the superiority of the given method. And that there is no restriction on the magnitude of dead time as far as the set point response is concerned, whereas there is the constraint theta divided by T 1 less than one for a satisfactory load disturbance rejection. So, controller settings for the next figure are K p equal to 0.1, K i equal to 10, K b is 5.017, K d is 3. 408, T l is equal to 2, and K l equal to 0.707. When the theta is 5, and K l is equal to 0.674, when theta equal to 5.5. What we have done basically, why there are two thetas here.

Now, we have assumed a 10 percent estimation error in theta, you are using some estimation technique; suppose the time delay is estimated erroneously; there is estimation error of 10 percentage in the estimation of time delay. Then it can go to a value of theta is equal to 4.5 or a value of 5.5 depending on under estimation error or over estimation error. So, assuming that we have got over estimation, then in place of theta equal to five we will use theta equal to 5.5, and simulate and see what type of performances we do get from the advanced smith predictor and controller.

Now, the solid line again shows the response we had obtained earlier; this is the one, we had obtained earlier. Now the dotted line shows the robust performance in the face of variation in parameters of the transfer function model. And the two responses show us that we do get robust performances by the controllers, we do get robust performances provided by the controllers although we do use very simple standard form based controller design technique to design the controllers. So, that is the beauty of the control - advanced control technique, and the tuning algorithm schemes.

Now, we will go to the case of a process with an integrator, and a long dead time. Earlier we have handled the control of unstable process, processes; now we go to the case of a process with an integrator, and large time delay. For the integrating process G p(S) is equal to K upon S e to the power minus theta S, when the controllers chosen are of the form G c equal to K p plus K i upon S, G c 1 is equal to K b, and G c 2 equal to K l. Then the delay free part of equation 4 results in Y r dash S is equal to K K p S plus K K i divided by S square plus K K p plus K K b S plus K K i.

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$$Y_r(s_n) = \frac{c_1 s_n + 1}{s_n^2 + d_1 s_n + 1} \qquad (24)$$

where

$$c_1 = (K_p/K_i)\beta \qquad (25)$$

and

$$d_1 = (KK_p + KK_b)\beta^{-1} \qquad (26)$$

Now, dividing both numerators and denominators by K K i, we can get a second order standard transfer function. So, assuming beta equal to root of K K i further results in a second order standard standard second order transfer function or second order standard transfer function of the form Y r Y r dash S n is equal to C 1 S n plus 1 divided by S n square plus d 1 S n plus 1. Again, where C 1 is given by the expression C 1 is equal to K p upon K i times beta, and d 1 is equal to K K p plus K K b times beta inverse.

So, relatively simpler expressions, then the earlier cases are obtain in this case, because we do not have any time constants associated with this process process dynamics or we do not have any time constant in the denominator of equation number 22. Now, how to design controllers? The three controllers for the integrating processes with long dead time, a higher value of beta is obtained again for a smaller value of K p upon K i for the chosen C 1 value, d 1 is obtained again from the standard form for with minimization of the ISTE criterion. The design value of K b is obtained from equation 26, k b yes there is the only unknown here, when we assume that K p is constraint to some value K p equal to one; and k is of course, known and beta has been chosen.

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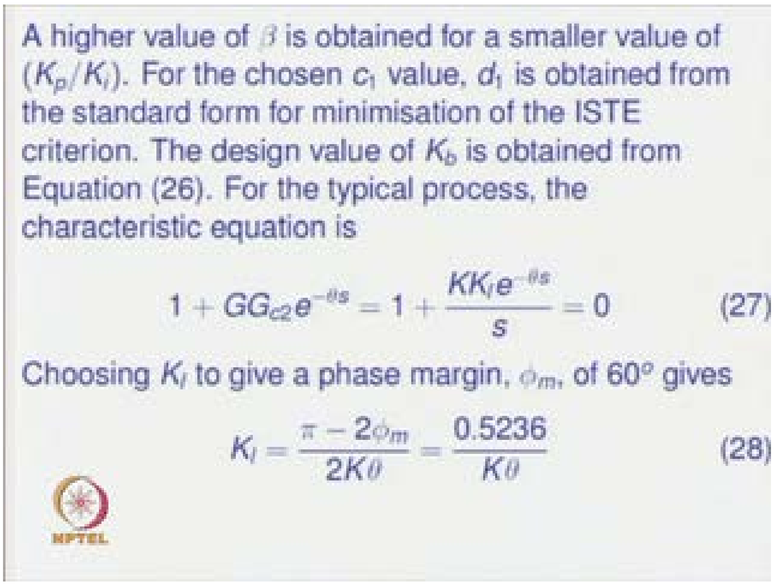

For the typical process now the characteristic equation becomes 1 plus G G c 2 e to the power minus theta s is equal to 1 plus KK l times e to the power minus theta S upon S is equal to 0. Now, choosing K l to give a phase margin of some specified value, suppose phi m phase margin is equal to 60 degree; that condition gives K l or gives an expression for the K l as K l is equal to pi minus two phi m upon two k theta upon substitution of phi m of 60 degree.

That means, pi by 3, you do get in the numerator for K l as 0.5236 or K l is now given expressed as K l is equal to 0.5236 upon K theta; K theta are known to us. As you know K theta are the parameters of the process or parameters of the process model.

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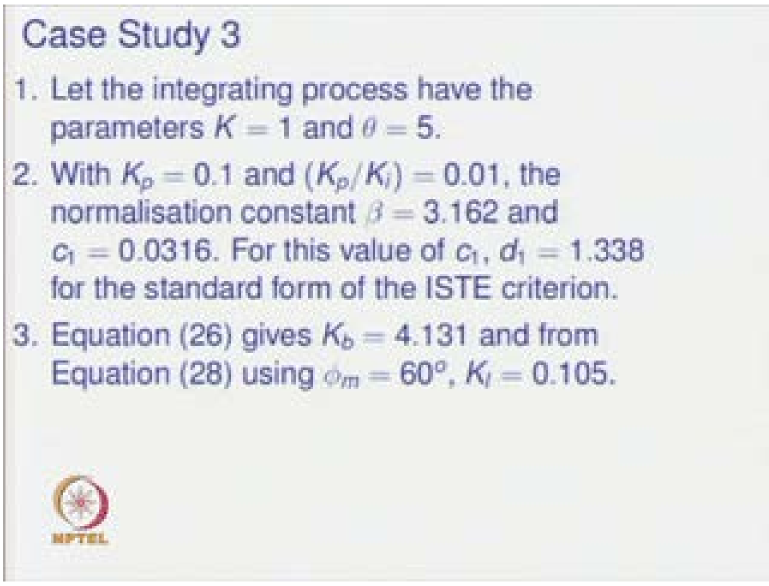

Now, in this case study, we will take the integrating process steady state process gain as K equal to 1, and time delay as theta equal to 5 with K p equal to 0.1, and K p upon K i is 0.01. A further smaller value is used here, the normalization constant beta becomes 3.162. So, in this study we are going for a higher value of beta. So, higher value this higher value of beta results in C 1 equal to 0.0316 for this value of C 1 further using the second order standard transfer function d 1 is obtained as 1.336 for the standard form of the ISTE criterion. Equation 26 gives now K b equal to 4.131, and from equation 28 using the phase margin of 60 degree, we do obtain K l as K l is equal to 0.105. Thus, all the parameters associated with the three controllers for controlling an integrating process are designed.

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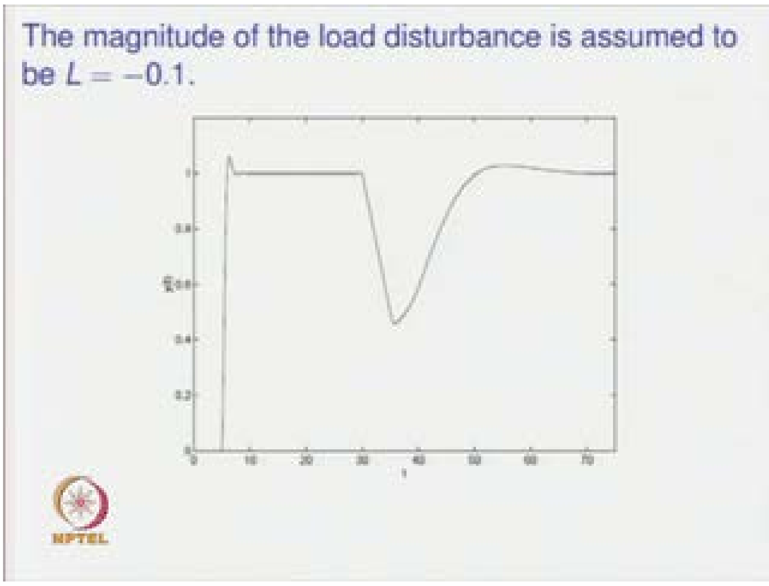

Let us see, that sort of responses we do get from the advanced smith predictor that is controlling an integrating process now. When the magnitude of the static load disturbance l is equal to minus 0.1, we do get the set point response that is given in the left hand side of this plot, and the load disturbance response given in the right side of this plot. Again, it is observed from this plot that, we do get very or quite satisfactory set point input response; although the disturbance response can be improved upon further.

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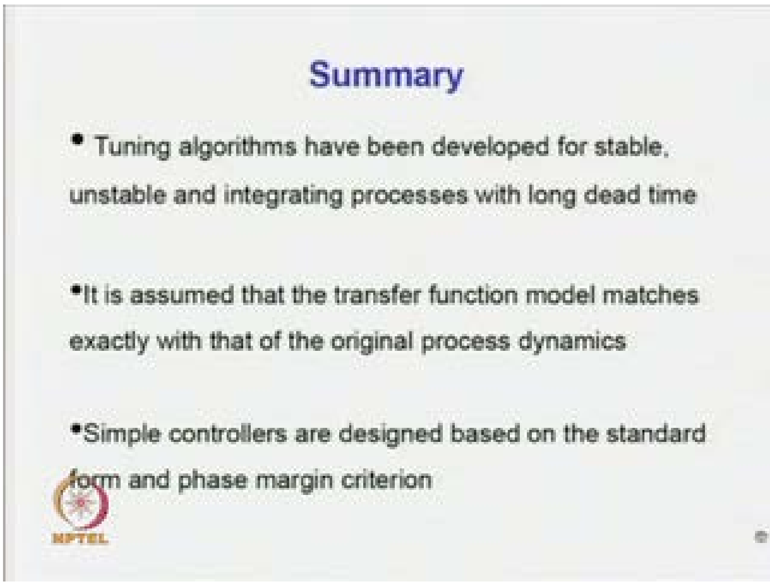

Let me summarize the lecture now. So, the tuning algorithms have been developed for stable, unstable, and integrating processes with long dead time. The tuning algorithms algorithms consider basically the standard transfer function forms, and standard transfer function based controller design techniques. It is assumed that the transfer function model matches exactly with that of the original process dynamics; this is one of the major requirement, we have for our analysis and controller design, and to design robust controller we need to relax upon this condition. Now, simple controllers are designed based on the standard form, and phase margin criterion.

So, G c and G c 1 are designed based on standard forms, and G c 2 is designed based on phase margin criterion; it can be optimum phase margin criterion, it can be ordinary phase margin criterion. Any point to ponder, can we use other methods to design the controllers, yes there exists numerous techniques, such as the direct synthesis method, phase margin methods, optimal controller design methods, loop shaping methods, robust controller design methods and so on; those can be used to design controllers for the advanced smith predictor control structure. Thanks.