Lead and Lag Controller Design

Angelos Kamariadis
School of Electrical and Computer Engineering
National Technical University of Athens
Athens, Greece
angeloskamariadis@gmail.com

***Abstract*—This project presents a comprehensive design and analysis of lead and lag controllers for a second-order temperature control system using both time-domain and frequency-domain techniques. Initially, the limitations of proportional control are demonstrated through Root Locus analysis, motivating the need for a phase lead compensator to meet transient performance specifications, including damping ratio and settling time. A systematic lead controller design is carried out by selecting desired dominant poles and applying angle and magnitude conditions, followed by validation through simulation and Bode diagram analysis. The impact of controller-induced zeros on overshoot is examined, and a reference filter is introduced to mitigate undesirable transient effects, restoring near-ideal second-order behavior. Subsequently, a lag controller is designed to improve steady-state accuracy by increasing the low-frequency gain while carefully addressing the trade-off between error reduction and dynamic performance. The placement of the lag compensator pole-zero pair is analyzed to balance phase margin preservation and response speed. Finally, robustness is assessed under model uncertainty, demonstrating that the designed system maintains stability within specified perturbation bounds. All designs and analyses are supported by numerical simulations implemented in Python.**

***Index Terms*—Lead Controller, Lag Controller, Bode Diagram, Nyquist Diagram, Root Locus, Robust Stability, M-Circles, Resonant Peak, Python**

I. LEAD CONTROLLER DESIGN

A. Root Locus Design

Initially, we consider a temperature control system, with the transfer function:

$$G(s) = \frac{1}{(s+1)(s+2)}$$

We observe that the system has two poles, at positions $p_1 = -1$ and $p_2 = -2$, while there are no zeros.

For the above system, the Root Locus was plotted using Python and its appropriate libraries. The Root Locus is depicted below.

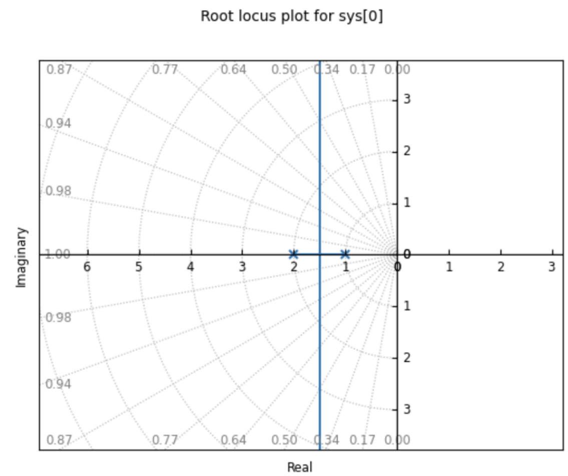

Fig. 1. Root Locus

To delineate the desired region of the Root Locus, which allows us to have a damping ratio (ζ) greater than or equal to 0.4 and a settling time (t_s) less than 1 second, we use the following formulas:

$$\zeta = \cos\theta \implies \theta = \cos^{-1}\zeta \implies \theta = \cos^{-1}(0.4) \implies \theta \approx 66.4^\circ$$

$$t_s \approx \frac{4}{\zeta\omega_n} \implies \zeta\omega_n > 4, \quad \zeta\omega_n = \sigma$$

where σ is the absolute value of the real part of the pole.

Therefore, the desired region is the space located both to the left of the vertical line $\sigma = -4$, and inside the cone of angle $\theta = \pm 66.4^\circ$. This region is shown below.

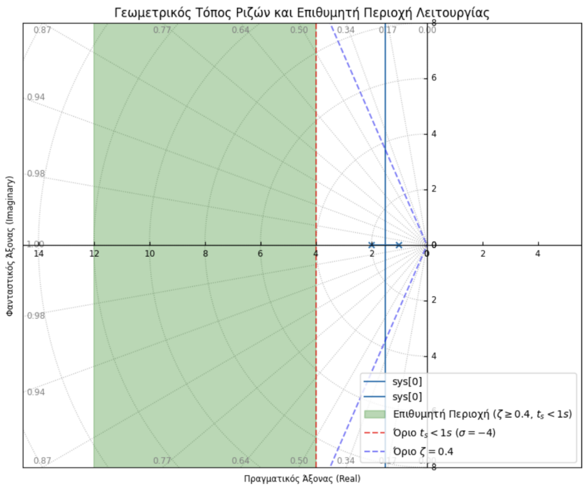

Fig. 2. Root Locus and Desired Operating Region

Can we achieve this with proportional feedback?

No, unfortunately, we cannot achieve this with only proportional feedback. The Root Locus with a proportional controller (P Controller) never passes through the desired specifications region. To satisfy these specifications, we would need to pull the Root Locus to the left, which requires a controller that adds zeros to the system, such as a proportional-derivative controller (PD Controller) or a phase lead controller (Lead Controller).

B. Lead Controller Design

The transfer function of the phase lead controller has the following form:

$$C_{lead}(s) = K_s \frac{s+z_c}{s+p_c}$$

For the design of the controller, we must first choose the pair of dominant poles, for which the specifications are $\zeta \geq 0.4$ and $t_s < 1$ s.

A safe choice is one where the poles over-satisfy the specifications. One such choice is:

$$\implies s = -5 \pm j5 \implies \zeta = \frac{5}{\sqrt{5^2+5^2}} \approx 0.707 > 0.4$$
$$\implies t_s \approx \frac{4}{\zeta\omega_n} = \frac{4}{5} < 1$$

Then, we arbitrarily choose a zero, which will "pull" the Root Locus to the left, and we calculate the position of the pole p_c.

Choosing the zero at position $z_c = 3$, the pole p_c is calculated with the following procedure.

$$\begin{aligned}
& s = -5 \pm j5,\ p_1 = -1 \text{ and } p_2 = -2 \\
& \implies \theta_{p_1} = 180^\circ - \arctan\left(\frac{5}{4}\right) = 128.66^\circ \\
& \implies \theta_{p_2} = 180^\circ - \arctan\left(\frac{5}{3}\right) = 120.96^\circ \\
& \implies \angle G(s) = -\theta_{p_1} - \theta_{p_2} = -249.62^\circ \\
& \quad \angle G(s) + \angle C(s) = -180^\circ \\
& \implies \varphi = \angle C(s) = 69.62^\circ \\
& \quad \theta_z = 180^\circ - \arctan\left(\frac{5}{2}\right) = 111.80^\circ \\
& \quad \theta_z - \theta_p = \varphi \implies \theta_p = 42.18^\circ \\
& \implies \tan\theta_p = \frac{5}{p_c - 5} \\
& \implies p_c \approx 10.5
\end{aligned} \tag{1}$$

Therefore, with the arbitrary choice of the zero at position $z_c = 3$, we forced the pole p_c to be located at position $p_c = 10.5$.

Finally, we calculate the gain K_c, such that: $|C(s) \cdot G(s)| = 1 \implies$

$$\begin{aligned}
& \implies \left| K_c \frac{s+3}{s+10.5} \cdot \frac{1}{(s+1)(s+2)} \right| = 1 \\
& |s+1| = \sqrt{(-4)^2+5^2} \approx 6.4 \\
& |s+2| = \sqrt{(-3)^2+5^2} \approx 5.8 \\
& |s+10.5| = \sqrt{5.517^2+5^2} \approx 7.45 \\
& |s+3| = \sqrt{(-2)^2+5^2} \approx 5.3 \\
& \implies K_c = \frac{6.4 \cdot 5.8 \cdot 7.4}{5.3} \implies K_c = 51.62
\end{aligned}$$

Now, regarding the overshoot, we calculate the expected overshoot by assuming that our system will behave like an ideal 2nd-order system, without zeros.

The relation that gives the overshoot, as a function of the damping ratio (ζ), is:

$$M_p \approx e^{-\frac{\pi\zeta}{\sqrt{1-\zeta^2}}} \cdot 100\%$$

$$\implies M_p \approx e^{-\frac{\pi \cdot 0.707}{\sqrt{1-0.707^2}}} \cdot 100\% \approx 4.3\%$$

By simulating the system with the controller we designed, using Python and its appropriate libraries, we found that the actual overshoot is noticeably larger. This is shown in the following graph.

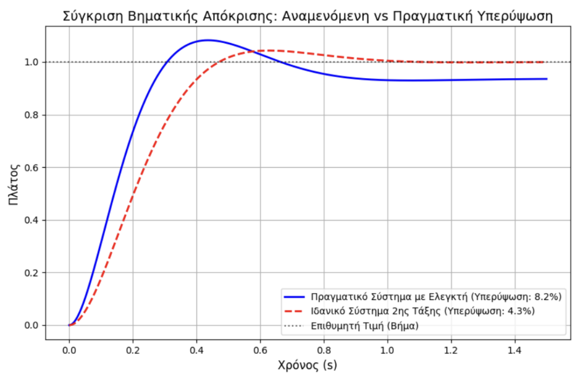

Fig. 3. Comparison of Step Response: Expected vs Actual Overshoot

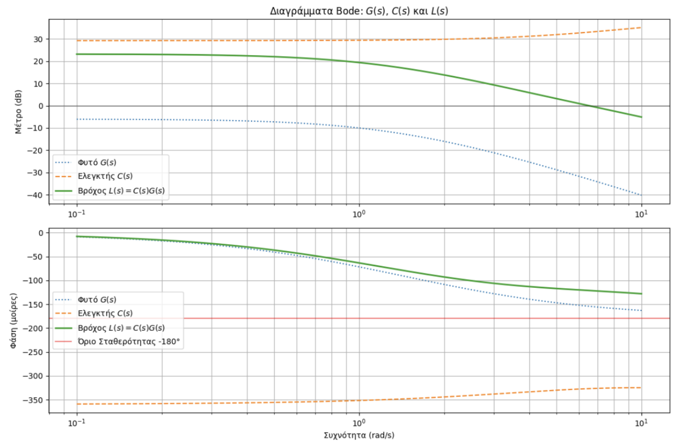

Fig. 4. Bode Plots: $G(s)$, $C(s)$ and $L(s)$

We observe that the system presented an actual overshoot of the order of 8.2%, while the theoretical, ideal overshoot was calculated at 4.3%. This difference is due to two main reasons:

The introduction of the zero in the closed loop

Because our controller has a zero, this zero remains as a zero in the closed-loop system as well. A zero in the left half complex plane acts like a derivative term (derivative action), which speeds up the system, i.e., it decreases the rise time, while simultaneously and unavoidably increasing the overshoot, pushing the response higher than expected.

The fact that the system increases in order (3rd order)

The controller we designed introduced an additional pole, with the result that the overall closed-loop system is no longer second order. The formulas for overshoot and settling time are approximate and strictly apply only to pure second-order systems. Therefore, the third pole and the zero alter the ideal behavior.

C. Bode Plot Design

We will use the controller we calculated previously, with a zero at position $z_c = 3$, a pole at position $p_c = 10.5$, and a gain of $K_c = 51.6$.

From the above Bode plots, we clearly observe the dual contribution of the lead controller C(s). On the one hand, its high gain raises the magnitude of the overall loop L(s) so that it crosses 0 dB at a relatively high gain crossover frequency (around 6-7 rad/s), making the system's response faster. On the other hand, in the phase plot we observe the lead in action; while the original system G(s) dangerously dips towards $-180°$, the positive phase introduced by the controller lifts the final phase of the loop L(s). The result is that exactly at the frequency where the magnitude crosses 0 dB, the green phase line is significantly far from the red $-180°$ limit line, providing a wide and safe Phase Margin. This exact safety distance in the Bode plot is what ensures the desired damping ratio and prevents severe oscillations in the time domain.

How much phase does the controller add at the gain crossover frequency?

The phase it adds is equal to approximately 50 - 60 degrees of positive phase at that point, ensuring a strong Phase Margin.

What is the phase margin?

The phase margin is approximately 65 degrees.

How is the phase margin related to the damping ratio which we also see in the Root Locus?

The phase margin is directly related to the damping ratio. In the time domain (i.e., in the Root Locus), the specification was to have a damping ratio less than or equal to 0.4, which determined the desired safety region within which the poles had to be located, so that the system would not exhibit large overshoot and oscillations. In the frequency domain (i.e., in the Bode plots), stability and overshoot are expressed through the phase margin. Therefore, when in the Root Locus we require a damping ratio less than or equal to 0.4 (for good damping), this is equivalent in the Bode plot to the requirement for a phase margin of approximately 40 degrees or greater. The phase lead controller we designed is exactly the same mechanism that lifts the phase in the Bode plot away from $-180°$, increasing the phase margin. In conclusion, these are two different visual representations of the exact same improvement in the stability and damping of the system!

D. Closed-Loop Transfer Function and Reference Filter

The calculation of the transfer function was performed using Python and its appropriate libraries, and is as follows:

$$T(s) = \frac{51.62s + 154.86}{s^3 + 13.517s^2 + 85.171s + 175.894}$$

Next, we experimented with a reference filter, which is placed outside the loop, before the controller, and its mission is to cancel the zero present in the closed-loop transfer function of the system. To achieve this, the filter must have a pole at position -3 and a unity DC gain.

Based on the above, the (transfer) function that describes the operation of the filter is:

$$F(s) = \frac{3}{s+3}$$

Below, the plots are depicted, from which we draw some important conclusions.

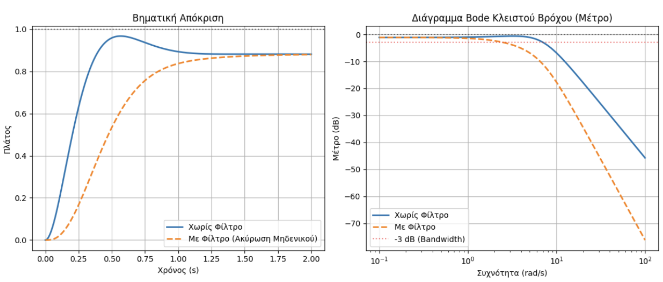

Fig. 5. Step Response (left) and Closed-Loop Bode Plot Magnitude (right): Without vs With Filter

How does the step response change?

Without the filter, the step response is very fast, but exhibits large overshoot. With the filter, the response becomes smoother, slightly slower, but the overshoot almost completely disappears.

Is the overshoot as predicted by the damping ratio?

It depends on whether or not the reference filter is used, as the classical theoretical formula for overshoot applies strictly only to ideal 2nd-order systems that have absolutely no zeros. In the initial closed-loop system without the filter, the actual overshoot is large and differs radically from the expected 4.3% predicted by our design because the zero at position -3 introduced by the controller adds a derivative action which accelerates the response and derails its peak. Conversely, when we introduce the reference filter with the appropriate pole, it mathematically cancels the specific zero, restoring the system's dynamics to the ideal standard 2nd-order behavior. As a result of this cancellation, the step response is smoothed out and the final overshoot is exactly at 4.3%, perfectly verifying the prediction of our initial damping ratio.

What is the reason for their differences?

The classical formulas connecting the damping ratio with the overshoot apply strictly only to ideal 2nd-order systems without zeros. The zero adds derivative action, skyrocketing the overshoot. The reference filter cancels this zero, bringing the system closer to the ideal standard behavior we designed in the Root Locus.

Closed-Loop Bode Plots & Tradeoffs

When we compare the closed-loop Bode plots, we observe that without a filter, the bandwidth is larger, while simultaneously an obvious peak appears above 0 dB. With the filter, the bandwidth is smaller and there is absolutely no peak.

Tradeoffs (with filter)*: Slower response, larger rise time, but perfectly smooth transition without oscillations.

Tradeoffs (without filter)*: Reacts much faster to changes, presents large overshoot and is more sensitive to high-frequency noise

Guided by the above tradeoffs, the design with the reference filter is preferable.

Thermal systems have large inertia, while they often feature only a heating element. If the temperature exhibits overshoot (for example if it goes to 110°C while we wanted 70°C), the controller will turn off the heater, but the system will take a very long time to cool down naturally, ruining the process. In such systems, the priority is to limit the probability of overshoot occurrence, something which we achieve with the use of the feedback filter.

For a more complete comparison of the two designs, what other plots could we choose?

For a more complete comparison of the two designs, we could choose the plot of the control signal, i.e., the signal that the controller produces and sends to the actuator. The reason for this choice lies in the fact that all signals can theoretically take infinite values, but in practice, a heater has a maximum power. The zero of the controller causes the so-called "Derivative Kick", a huge and sudden power demand the moment we apply the step reference signal. If we plot the control signal, we will see that without the filter, the controller asks for an unfeasibly large initial power that will lead the actuator to a state of saturation. With the reference filter, the change in the signal reaching the controller is smoother, resulting in the control signal being much smoother and realistic for the limits of our hardware.

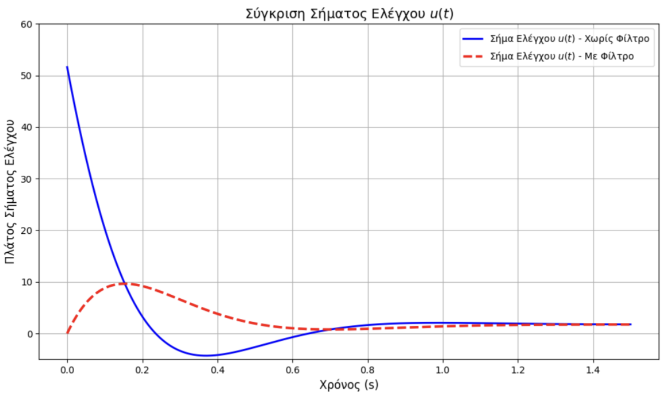

Fig. 6. Comparison of Control Signal $u(t)$

As can be seen in the above graph, although both systems eventually reach the same desired result, the design with the filter is the one that makes the control signal realistic, smooth, and safe for the actual factory equipment.

E. Crossover Frequency greater than 100 rad/sec and Phase Margin greater than 90°

Mathematically speaking, if we assume that the ideal scenario without physical constraints exists, we could achieve a crossover frequency above 1000 rad/sec and a phase margin greater than 90°, using very high gains and multiple stages of phase lead. However, this is an extremely bad idea. On the one hand, because a phase margin over 90° corresponds to a damping ratio greater than unity, a fact that would make the system overdamped, completely negating the initial goal of high speed. On the other hand, because maintaining such a vast bandwidth would act as a "magnifying glass" for high-frequency sensor noise and simultaneously excite the unseen, unmodeled dynamics of the system at these high frequencies, leading it with mathematical precision to uncontrollable instability.

II. LAG CONTROLLER DESIGN

A. Steady-State Error for Step Input

Assuming that we use the system from the 1st part, which has no pure integrator/pole at 0, for a step input, the steady-state error is calculated from the position constant:

$$K_p = \lim_{s \to 0} L(s) = \lim_{s \to 0} C_{lead}(s)G(s)$$

Using the parameters: $K_c = 51.62$, $z_c = 3$ and $p_c = 10.517$

$$\implies K_p = \lim_{s \to 0} K_c \frac{s + z_c}{s + p_c} \cdot \frac{1}{(s+1)(s+2)}$$

Substituting the values into the formula, we get:

$$\implies K_p \approx 7.36$$

The steady-state error is given by the formula:

$$e_{ss} = \frac{1}{1 + K_p} \implies e_{ss} \approx 0.1196 \equiv 11.96\%$$

We conclude that the system stabilizes at approximately 0.88, instead of 1.

B. Lag Controller Design

The Lag controller has the form:

$$C_{lag}(s) = \frac{s + z_{lag}}{s + p_{lag}}$$

with $\alpha = \frac{z_{lag}}{p_{lag}}$ being the DC Gain, which describes the amplification at low frequencies.

For the controller to reduce the steady-state error for a step input to a value less than 0.5%, it must be:

$$e_{ss} \leq 0.005 \implies e_{ss} = \frac{1}{1 + K_{p,lag}} \leq 0.005 \implies K_{p,lag} \geq 199$$

$$\implies \alpha = \frac{z_{lag}}{p_{lag}} \approx 27.03$$

For safety, we choose $\alpha = 30$.

Tradeoffs resulting from the placement of the pole-zero pair for a given α

For a given $\alpha = \frac{z_{lag}}{p_{lag}}$, which achieves the desired amplification that guarantees a steady-state error for a step input of less than 0.5%, there are two possible scenarios for the placement of the pole-zero pair. The zero at position $|\pm 300|$ and the pole at $|\pm 10|$, or the zero at position $|\pm 0.3|$ and the pole at $|\pm 0.001|$.

- 1st Scenario: $z_{lag} = 0.3$ and $p_{lag} = 0.001$

Pros: At the crossover frequency, the Lag controller will have a phase of almost 0 degrees. This is perfect, because it means it will not spoil the phase margin and the speed we achieved with the Lead at all!

Cons: The step response will acquire a slow tail. The system will quickly reach 88% (thanks to the Lead), but the remaining 12% to reach 99.5% will take a very long time to cover, due to the huge time constant of the very small pole.

- 2nd Scenario: $z_{lag} = 3$ and $p_{lag} = 0.1$

Pros: The slow tail disappears. The system will close the error much faster.

Cons: If we place the pair so close to the crossover frequency, the Lag controller will introduce negative phase (delay) exactly where it shouldn't! This will affect the phase margin, dramatically reducing the damping ratio. The step response will suddenly be filled with huge oscillations or the system may even become unstable.

To balance this tradeoff, it is necessary to place the zero of the Lag approximately one decade behind the crossover frequency. Thus, the error is corrected relatively quickly, without causing catastrophic oscillations!

III. Understanding and Relationships of Diagrams

A. *Correlation of Pole Positions - Overshoot and Resonance in Bode*

The transfer function of the system is:

$$G(s) = \frac{1}{s^2 + \alpha s + 1}$$

Comparing it with the general form: $G(s) = \frac{1}{s^2+2\zeta\omega_n s+\omega_n^2}$, we draw the following conclusions.

Initially, $\omega_n^2 = 1$ and $\alpha = 2\zeta\omega_n \implies \zeta = \frac{\alpha}{2}$

Thus, for $\alpha = 1.5 \implies \zeta = 0.75 \implies$ the positions of the poles will be:

$$s_{1,2} = -\zeta\omega_n \pm j\omega_n\sqrt{1-\zeta^2} = -0.75 \pm j0.661$$

Since, therefore, the poles are complex conjugates and have an imaginary part, like any system with complex poles, this specific one too will be underdamped and by definition will present overshoot and oscillation in the step response.

Regarding resonance, in the Bode plot the magnitude presents a peak (resonant peak), only if the condition holds:

$$\zeta < \frac{1}{\sqrt{2}}$$

In our case we have $\zeta = 0.75$. Since $0.75 > 0.707$, the system does not present a peak! The magnitude starts from 0 dB (for $\omega = 0$) and simply decreases continuously as the frequency increases. It does not amplify any frequency above 1 (the 0 dB).

How do we explain the fact that while the system does not amplify any frequency, the step response in the time domain exhibits overshoot?

The magnitude $|G(j\omega)|$ alone is not sufficient to explain the overshoot; the phase is also needed. The overshoot in the time domain does not arise because the system amplifies certain frequencies, but from the phase shift, which forces the frequency components of the input to stack up and sum in such a way that they create a transient peak before stabilizing.

B. *Finding Phase Margin and Gain Margin using Nyquist and Bode Plots*

We consider a unity feedback system with an open-loop transfer function:

$$G(s) = 100\frac{1}{(s+1)(s+2)}\frac{900}{s^2+2s+900}$$

The term: $\frac{900}{s^2+2s+900}$, corresponds to a 2nd-order subsystem with frequency $\omega_n = 30$ rad/s and damping $\zeta = 0.033$. The system, as is evident, based on what was mentioned previously, presents a resonant peak.

Below, the Bode and Nyquist plots for the above system are depicted, which were designed using Python and its appropriate libraries.

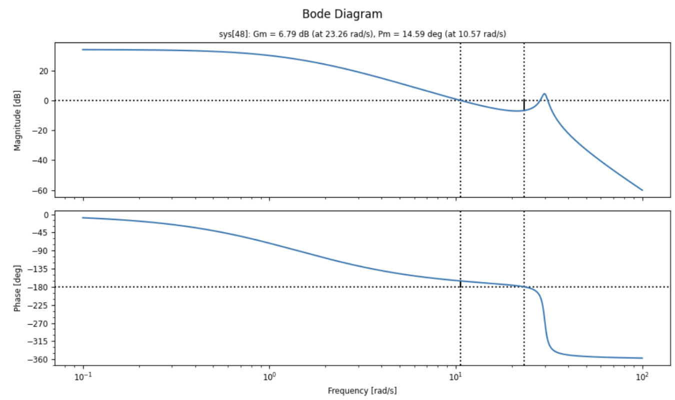

Fig. 7. Bode Diagram

The gain margin and phase margin have been calculated as 6.79 dB and 14.59 degrees, respectively.

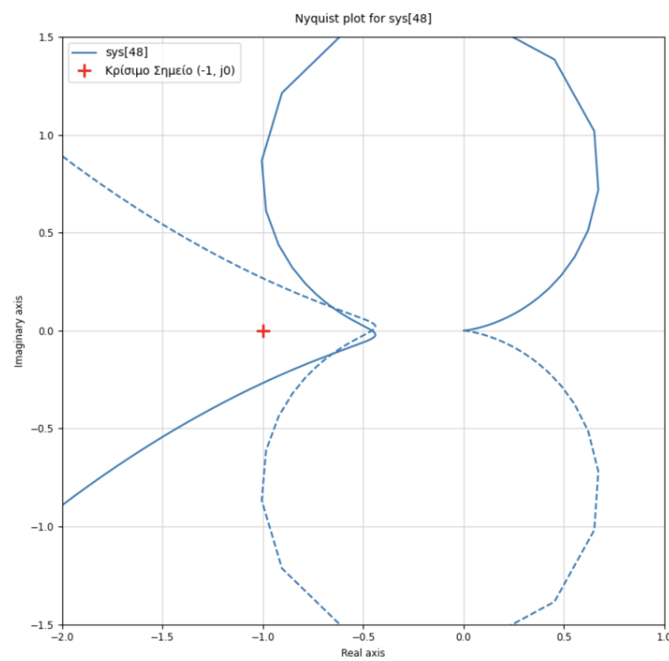

Fig. 8. Nyquist plot (Note: The red cross indicates the Critical Point at $-1 + j0$)

From the Nyquist plot we calculate the gain and phase margin, which are respectively 6.9 dB and 37°.

In conclusion, the Bode plot is inadequate or even misleading for evaluating margins in systems with strong resonances. In such cases, confirmation with the Nyquist plot is absolutely necessary to see the actual distance of the system from instability.

C. *Second-order System with feedback*

We consider a second-order system with the following transfer function:

$$G(s) = \frac{5}{s^2 + s + 1}$$

The system is connected in a unity feedback configuration. To compute the stability margin, we use:

$$s_m = \inf_{\omega} |G(j\omega)|$$

which corresponds to the minimum distance of the Nyquist plot from the critical point $-1 + j0$.

Equivalently, we use:

$$s_m = \inf_{\omega} |1 + G(j\omega)|$$

First, we compute $1 + G(j\omega)$:

$$G(j\omega) = \frac{5}{(j\omega)^2 + (j\omega) + 1}$$

Thus,

$$1 + G(j\omega) = 1 + \frac{5}{(j\omega)^2 + (j\omega) + 1}$$

Combining into a single fraction:

$$1 + G(j\omega) = \frac{(j\omega)^2 + (j\omega) + 1 + 5}{(j\omega)^2 + (j\omega) + 1}$$

Since $(j\omega)^2 = -\omega^2$, we obtain:

$$1 + G(j\omega) = \frac{(6 - \omega^2) + j\omega}{(1 - \omega^2) + j\omega}$$

To simplify calculations, we consider the squared magnitude:

$$|1 + G(j\omega)|^2 = \left| \frac{(6 - \omega^2) + j\omega}{(1 - \omega^2) + j\omega} \right|^2$$

Using $|a + jb|^2 = a^2 + b^2$, we get:

$$|1 + G(j\omega)|^2 = \frac{(6 - \omega^2)^2 + \omega^2}{(1 - \omega^2)^2 + \omega^2}$$

Let:

$$x = \omega^2$$

Then the function becomes:

$$f(x) = \frac{(6 - x)^2 + x}{(1 - x)^2 + x}$$

Expanding:

$$f(x) = \frac{x^2 - 11x + 36}{x^2 - x + 1}$$

The stability margin is determined by finding the minimum value of $f(x)$ for $x \geq 0$, which corresponds to:

$$s_m = \sqrt{\min f(x)}$$

To find the extrema, we differentiate:

$$\frac{df(x)}{dx} = \frac{(2x - 11)(x^2 - x + 1) - (x^2 - 11x + 36)(2x - 1)}{(x^2 - x + 1)^2}$$

Setting the numerator equal to zero:

$$(2x - 11)(x^2 - x + 1) - (x^2 - 11x + 36)(2x - 1) = 0$$

This simplifies to:

$$10x^2 - 70x + 25 = 0$$

Solving the quadratic equation:

$$x_{1,2} = \frac{70 \pm \sqrt{(-70)^2 - 4 \cdot 10 \cdot 25}}{20}$$

$$x_{1,2} = \frac{7 \pm \sqrt{39}}{2}$$

Numerically:

$$x_1 \approx 0.377, \quad x_2 \approx 6.622$$

Evaluating $f(x)$ at these points, we find that the minimum occurs at x_2.

The stability margin is:

$$s_m = \sqrt{f(x_2)} = \sqrt{\frac{x_2^2 - 11x_2 + 36}{x_2^2 - x_2 + 1}}$$

Substituting $x_2 \approx 6.622$:

$$s_m \approx 0.428$$

The minimum distance of the Nyquist plot from the critical point is approximately:

$$s_m \approx 0.428$$

which confirms the stability margin of the system.

D. Relation to M-Circles

M-circles in the complex plane represent loci where the magnitude of the closed-loop transfer function remains constant.

The stability margin also defines a circle centered at the critical point $-1 + j0$ with radius s_m. The Nyquist plot is tangent to this circle but never enters it.

This circle corresponds to the maximum sensitivity of the system.

E. Relation to Resonant Peak

For a sinusoidal input, the steady-state output amplitude equals the magnitude of the closed-loop transfer function.

The maximum value of this magnitude is known as the *resonant peak*. Its relation to the stability margin is:

- As s_m decreases (Nyquist plot approaches -1), the resonant peak increases sharply.
- A sufficiently large stability margin ensures controlled amplification and prevents excessive oscillations.

F. Robust Stability with Uncertainty

Assume the perturbed system:

$$\tilde{G}(s) = G(s) + D(s), \quad |D(j\omega)| \leq 0.1$$

Using the Nyquist stability criterion, robust stability is guaranteed if:

$$|D(j\omega)| < |1 + G(j\omega)|, \quad \forall\omega$$

Since:

$$\max|D(j\omega)| = 0.1, \quad \min|1 + G(j\omega)| = s_m \approx 0.428$$

we have:

$$0.1 < 0.428$$

Thus, the perturbation is insufficient to drive the Nyquist plot through the critical point.

The system is robustly stable, since the uncertainty remains strictly within the stability margin. Therefore, the closed-loop system preserves stability under the given perturbation.

IV. Conclusion

This project presented the design of lead and lag controllers for a second-order system using Root Locus, Bode, and Nyquist methods. A lead compensator was developed to satisfy transient performance requirements, improving speed and phase margin, while a reference filter was introduced to reduce overshoot caused by the controller zero. A lag compensator was then designed to significantly reduce steady-state error without substantially affecting system stability. The results highlight the trade-offs between transient response, steady-state accuracy, and robustness, as well as the complementary role of time- and frequency-domain analysis in control system design.

References

[1] Richard C. Dorf, Robert H. Bishop, *Modern Control Systems*, 12th ed. Pearson, Upper Saddle River, NJ, 2011.

[2] Vardoulakis Antonios - Ioannis, *Introduction to Mathematical Theory of Signals, Systems and Control*, 2 ed. A. Tziola and Sons Publications.