Design of negative feedback amplifiers

Design Procedure Negative Feedback Amplifiers

1. Measure the load signal (V or I)

a. Voltage should be measured across (in parallel with) the load
b. Current should be measured through (in series with) the load

2. Design a network that generates of copy of the source signal (V or I) from the measured load signal

a. The transfer of this network is the reciprocal of the desired source-to-load transfer

3. Subtract the copy from the source signal

a. In case of a voltage source signal, the signal source and the output of the feedback network should be connected anti-series
b. In case of a current source signal, the signal source and the output of the feedback network should be connected anti-parallel

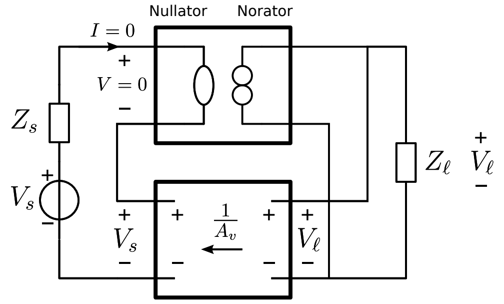
4. Nullify the difference

a. In case of a voltage source signal, a nullator closes the loop of the above anti-series connection
b. In case of a current source signal, a nullator is placed in parallel with the above anti-parallel connection
c. In case of a voltage load signal, a norator is placed in parallel with the load
d. In case of a current load signal, a norator closes the loop of the series connection of the load and the input of the feedback network

Single-loop negative Feedback Amplifiers

Source	Load	Amplifier	A, B, C, D	Feedback method
V	V	Voltage amplifier	A, 0, 0, 0	Output voltage sensing / parallel feedback Input voltage comparison / series feedback
V	I	Transadmittance	0, B, 0 ,0	Output current sensing / series feedback Input voltage comparison / series feedback
I	V	Transimpedance	0, 0, C, 0	Output voltage sending / parallel feedback Input current comparison / parallel feedback
I	I	Current amplifier	0, 0, 0, D	Output current sensing / series feedback Input current comparison / parallel feedback

Nullor is ideal controller

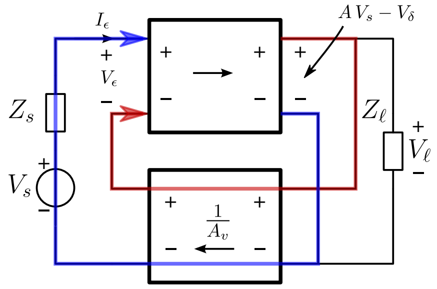


Feedback techniques

Implementation of the feedback network

1. Nonenergic feedback
   a. Transformers, gyrators
   b. Short and open circuit
2. Passive feedback
   a. Non-dissipative: inductors and/or capacitors
   b. Dissipative: includes resistors
3. Active feedback
   a. Feedback network comprises an amplifier

Negative feedback



Multiple-loop feedback

Similar procedure.

Performance improvement by negative feedback

Static and dynamic drive capability

The static and dynamic drive capability of a feedback amplifier at best equal those of its controller.
Voltage drop across and current flow through feedback elements cause a reduction of the static and dynamic drive capability.

Power efficiency

The power efficiency of a feedback amplifier at best equals that of its controller. Power losses and energy storage in feedback elements generally cause a reduction of the power efficiency.

Noise

The equivalent-input noise sources of a feedback amplifier at best equal those of its controller.
Passive feedback elements generally increase the contribution of the controller input noise sources to the total source-referred noise.
Noise sources associated with passive and active feedback elements also deteriorate the signal-to-noise ratio.

Zero error (offset) and drift

The (temperature-dependent) equivalent-input offset voltage and offset current of DC feedback amplifiers at best equal those of its controller.
Passive feedback elements generally increase the contribution of these sources to the total source-referred (zero error) offset.
Offset sources associated with active feedback elements also contribute to the total source-referred (zero error) offset.

PSRR

Injection of power supply noise into the signal path reduces the PSRR. Increasing the gain of the stages that preceed the power supply noise injection, helpd to increase the PSRR.

Static inaccuracy

The static gain inaccuracy of a feedback amplifier at best equals the reciprocal value of the static (DC) **loop gain**.
Generally the feedback network itself also contributes to the static gain inaccuracy.

Static weak nonlinearity

The static differentiial gain error of a feedback amplifier at best equals the quotient of the static differentail gain error of the loop gain and the loop gain in the quiescent operating point.

Bandwidth

The bandwidth of a negative feedback amplifier equals that of its **servo function** (also: discrepancy factor).
For an all-pole feedback amplifier with **n dominant poles** it equals the n-th root of its n-th order **loop gain-poles product**.

Dynamic weak nonlinearity

The frequency dependent differentiial gain error of a feedback amplifier at best equals the quotient of the frequency-dependent differentail gain error of the loop gain and the magnitude of the frequency dependent loop gain in the quiescent operating point.

Ideal Gain

1. The ideal gain is the source-to-load transfer in the case of a nullor as controller
2. Practical controllers have a finite gain and bandwidth
   a. Negative (corrective) feedback if the transfer from the positive output of the controller to its positive input is inverting:

Modeling of circuits with feedback

Facilitate a two-step design approach:

1. Design of the ideal gain with nullators and norrators (perfect controllers)
2. Design of the controller: the deviation of the behavior from that with the ideal controller:
   - Servo function (Montagne)
   - Discrepancy factor (Middlebrook)

Asymptotic-gain feedback model

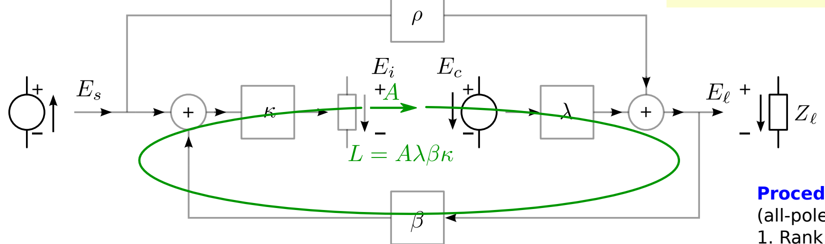
A : Loop gain reference variable: gain of the controlled source selected as loop gain reference
β : Feedback factor (reciprocal value of ideal gain)
ρ : Direct transfer

Asymptotic-gain: $A_{f\infty} = \lim_{A\to\infty} \frac{E_\ell}{E_s}$

Loop gain: $L = A \left.\frac{E_i}{E_c}\right|_{E_s=0}$

Gain: $A_f = \frac{E_\ell}{E_s} = A_{f\infty}\frac{-L}{1-L} + \frac{\rho}{1-L}$

Servo Function: $S = \frac{-L}{1-L}$ Measure for deviation of the gain from the asymptotic gain as a result of a lack of loop gain



Proper selection of the loop gain reference:
- Asymptotic-gain = ideal gain
- Direct transfer can be ignored
- Servo function is a measure for the difference between the actual gain and the ideal gain.

Servo bandwidth with all-pole loop gain function

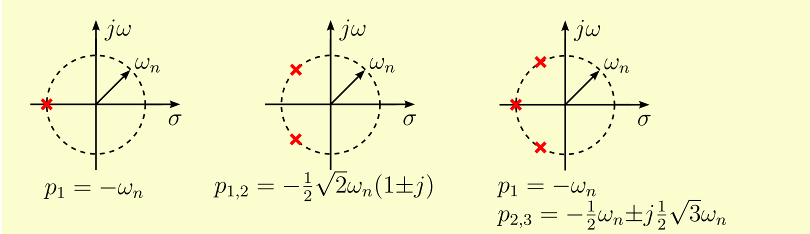
All-pole loop gain function
- Finite DC loop gain
- n poles

All-pole servo function
- Coefficient of the highest order of 's' in the denominator is the product of all the poles and the DC loop gain; (assume |L|>>1)

$$L(s) = \frac{L_{DC}}{\prod_{i=1}^{n}\left(1-\frac{s}{p_i}\right)}$$

$$S(s) = \frac{-L_{DC}}{1-L_{DC}}\frac{1}{1+\ldots+(-1)^n\frac{s^n}{(1-L_{DC})\prod_{i=1}^{n}p_i}}$$

MFM or Butterworth filter response



$p_1 = -\omega_n$ $\quad$ $p_{1,2} = -\frac{1}{2}\sqrt{2}\omega_n(1\pm j)$ $\quad$ $p_1 = -\omega_n$, $p_{2,3} = -\frac{1}{2}\omega_n \pm j\frac{1}{2}\sqrt{3}\omega_n$

$H_1(s) = \frac{1}{1+s/\omega_n}$ $\quad$ $H_2(s) = \frac{1}{1+s\sqrt{2}/\omega_n+s^2/\omega_n^2}$ $\quad$ $H_3(s) = \frac{1}{1+2s/\omega_n+2s^2/\omega_n^2+s^3/\omega_n^3}$

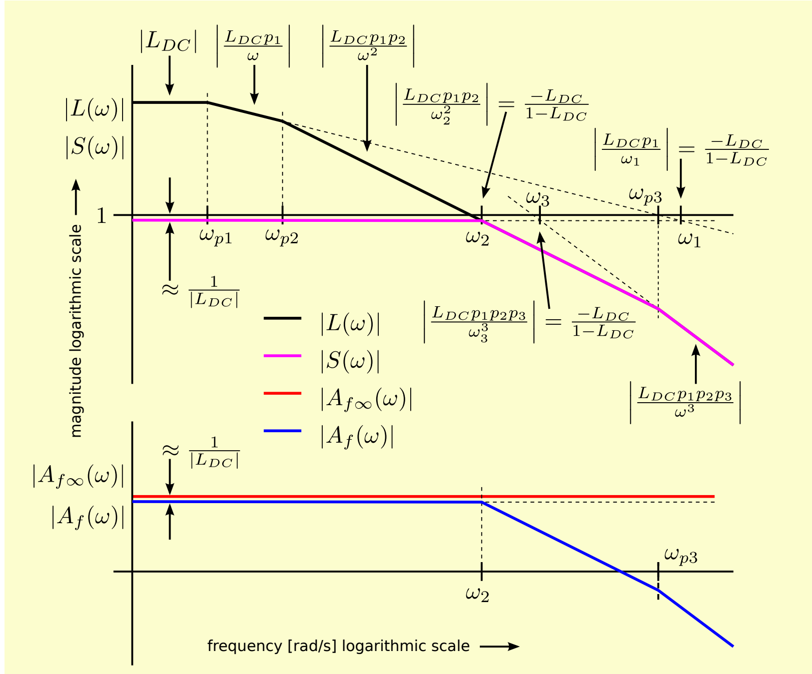
-3dB bandwidth of n-th order MFM filter

$$\omega_n = \sqrt[n]{\left|\prod_{i=1}^{n}|p_i|\right|}$$

-3dB bandwidth of servo function with n-th order MFM filter characteristic

$$\omega_n = \sqrt[n]{\left|(1-L_{DC})\prod_{i=1}^{n}p_i\right|} \approx \sqrt[n]{\left|L_{DC}\prod_{i=1}^{n}p_i\right|}$$

Loop gain-poles product (LP product)
- The bandwidth requirement for the amplifier sets a requirement for the product of the DC loop gain and the dominant poles of the loop gain.
- The controller should be selected or designed such that its contribution to this product is sufficiently large, while the number of poles should be kept as small as possible.



1. Dominant poles are those that contribute to the bandwidth of the servo function
2. The bandwidth of the servo function can be approximated as the frequency of intersection of the asymptotic approximation of the magnitude characteristic of the loop gain and unity (assuming dominant poles only)

Procedure for finding dominant poles

(all-pole loop gain only)
1. Rank poles of the loop gain (ascending order of absolute frequency)
2. For increasing order i, calculate the achievable MFM bandwidth from the i-th order LP product
   - If the bandwidth decreases with the order i, or if the frequency of the i-th pole exceeds the (i-1) order MFM bandwidth, the i-th pole is not dominant and the (i-1) order MFM bandwidth is the maximum achievable MFM bandwidth.

Non dominant poles

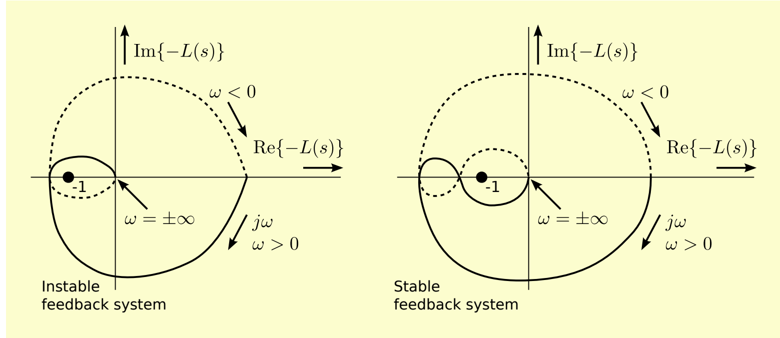
The influence of the non-dominant poles on the achievable MFM bandwidth cannot always be ignored.

Frequency stability

After the bandwidth has been designed, the amplifier may be instable, and in general will not have the desired frequency response.
A system is stable if all the roots of the characteristic equation (poles) are located in the left half of the complex plane. Note: a delay function has an infinite number of poles and zeros.

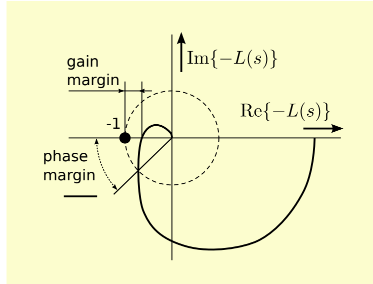
Routh-Hurwitz stability criterion
- Mathematical test using the Routh Array

Nyquist stability criterion
- The number of clockwise encirclements of the point (-1,0) in the complex plane by the polar (contour) plot of -L equals the number of right half plane poles of the gain minus the number of right half plane poles of the loop gain

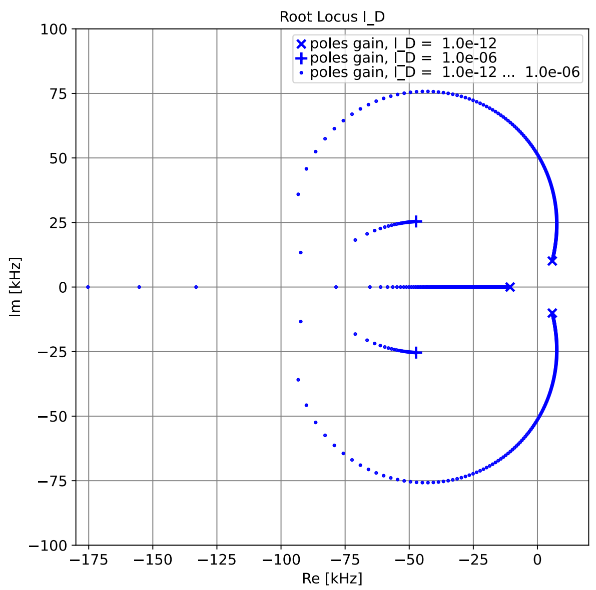


Gain margin and phase margin
Simple stability criterion for all-pole second-order feedback systems

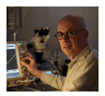
- The distance of the contour to -1 should be kept sufficiently large.
- 2-nd order MFM: PM = 60 degrees
- No unique correspondence between frequency response and gain and phase margin in higher order systems or in feedback systems that have zeros in the loop gain.



Root-locus technique
Graphical method, tracing out the paths (branches) of the poles of the servo function while increasing the DC or mid-band loop gain from zero to infinity.
1. The number of branches equals the number of poles
2. Poles are either real or complex conjugated
3. A branch starts at a pole of the loop gain
4. A branch ends at a zero of the loop gain. If there are n poles and m zeros, we assume m-n zeros at infinity
5. Parts of the real axis left from an odd number of poles plus zeros are part of a branch
6. If there are n poles and n zeros n-m branches go to infinity. The angle of their asymptotes is $\theta_i = \frac{2i+1}{n-m}\pi,\ i=0,1,2,\ldots$
7. The asymptotes intersect the real axis at $\sigma = \frac{\sum_{k=1}^{n}p_k - \sum_{i=1}^{m}z_i}{n-m}$
8. Break away and arrival points on the real axis are found from solving $\frac{d}{ds}L(s) = 0$
9. The angles between the branches at the break-away or arrival points are equally spaced over 360 degrees
10. Each point on a branch satisfies |L(s)|=1 and arg(-L(s))=180 degrees
Note: L=-Hk; difference between Black's model and the asymptotic gain model.



With SLiCAP we can draw root-locus plots with an arbitrarily selected root locus variable



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