Derive Controller Requirements from Amplifier Specifications

Noise

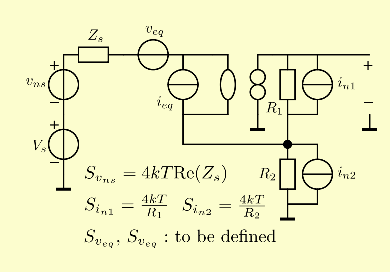
Noise contributions

Source
- Noise associated with the signal source

Feedback network
- If the feedback networks comprise a dissipative element

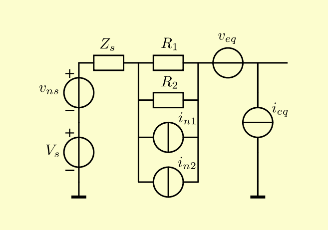
Controller
- Equivalent input voltage noise
- Equivalent input current noise
- Inflence on SNR may be enlarged by passive feedback elements

Power supply
- Can be modeled as part of equivalent input noise sources, sets requirement for PSRR.



Nonenergic feedback elements

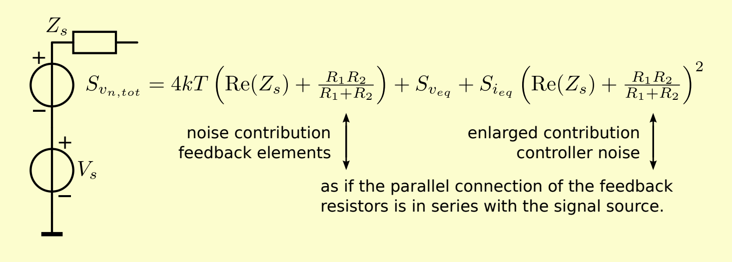
Equivalent input noise sources of the amplifier equal those of the controller



Effect of passive feedback elements

1. **Voltage amplifier**
   As if the parallel connection of the feedback elements is in series with the signal source.
2. **Transimpedance amplifier**
   As if the feedback element is in parallel with the signal source.
3. **Transadmittance amplifier**
   As if the feedback element is in series with the signal source.
4. **Current amplifier**
   As if the series connection of the feedback elements is in parallel with the signal source.

$$F = \frac{\text{Total source referred noise}}{\text{Source noise}}$$



$$S_{v_{n,tot}} = 4kT\left(\mathrm{Re}(Z_s) + \frac{R_1R_2}{R_1+R_2}\right) + S_{v_{eq}} + S_{i_{eq}}\left(\mathrm{Re}(Z_s) + \frac{R_1R_2}{R_1+R_2}\right)^2$$

Controller noise budgets

Solve spectra of controller equivalent-input noise sources from total source-referred noise.
Show stopper for spectrum of equivalent-input voltage noise: assume spectrum of equivalent-input current noise zero.
Show stopper for spectrum of equivalent-input current noise: assume spectrum of equivalent-input voltage noise zero

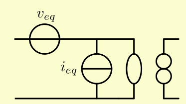
Application of OpAmp controllers
- Spectral density equivalent-input noise voltage
- Spectral density equaivalent-input noise current

Transistor-level design of controllers
- Device type selection and design operating conditions (PCA technology)
- Design of input stage device and operating conditions (IC technology)

Modeling of controller noise

- Spectral density of equivalent-input voltage noise source.
- Spectral density of equivalent-input current noise source.

"Noisy nullor" concept:
Ideal controller, except for noise



Biasing and DC operating point stability

Causes for bias errors

Tolerances and temperature dependency of device parameters cause piece to piece variation and temperature dependency of the DC operating point of the amplifier.

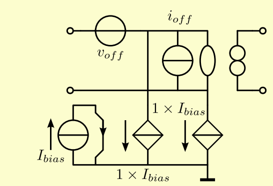
Modeling of controller bias errors

- Probablity function of equivalent-input DC offset voltage source
- Probablity density function of equivalent-input DC offset current source
- Probability density function of input bias current

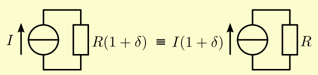
Application of OpAmp controllers
- Input offset voltage
- Input common-mode voltage range
- Input bias current
- Input offset current

Transistor-level design of controllers
- Device type selection and design opearting conditions (PCA technology)
- Design of input stage device and operating conditions (IC technology)
- Biasing design may conflict with noise design. Noise design considerations dominate over bias design considerations.
- Error-reduction techniques can be applied to reduce biasing errors.

"Nullor with bias and offset"
Ideal controller, except for :
- input offset voltage
- input offset current
- input bias current



"Simplified statistical DC analysis"
Replace variance of resistance with variance of current through that resistor



Controller bias error budgets

Use SLiCAP statistical DC analysis methods to derive budgets for statistical properties of offset and bias sources from the budget for the amplifier's operating point variance.

Power efficiency

Contributions to power losses and energy storage

Load
- Power losses and energy storage in the load

Feedback network
- Power losses in the feedback networks if they comprise dissipative elements
- Energy storage in the feedback networks if they comprise reactive elements.

Controller
- Energy storage and power losses are always present.
- Signal-dependent effects can be modeled with controlled voltage and current source and impedances at the output port.
- Quiescent losses and energy storage can be modeled with controlled sources and impedances in the power supply port.
- Inflence on power efficiency may be enlarged by passive feedback elements.

General design rule

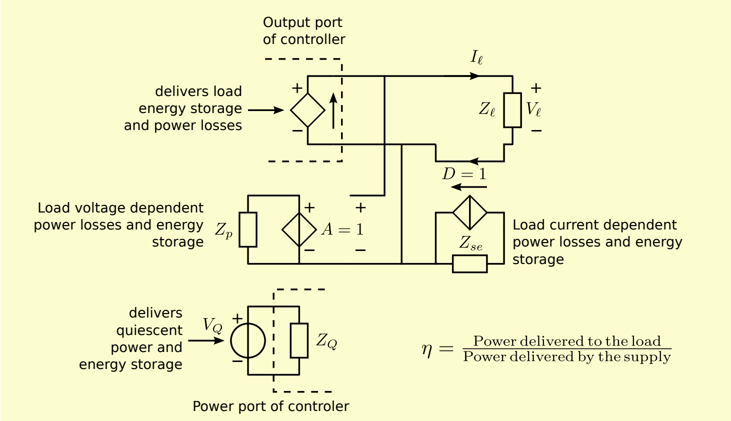
Impedances in series or admittances in parallel with the signal path should be kept as small as possible:
- They increase energy storage and/or power losses.
- Increase of energy storage almost always results in:
  - Deterioration of overdrive recovery
  - Increase of power losses.

Effect of passive feedback elements

1. **Voltage amplifier**
   As if the series connection of the feedback elements is in parallel with the load.
2. **Transimpedance amplifier**
   As if the feedback element is in parallel with the load.
3. **Transadmittance amplifier**
   As if the feedback element is in series with the load.
4. **Current amplifier**
   As if the parallel connection of the feedback elements is in series with the load.

Nonenergic feedback elements

Power losses and energy storage in the amplifier equal those of the controller.



$$\eta = \frac{\text{Power delivered to the load}}{\text{Power delivered by the supply}}$$

Voltage and current handling capability

Proper operating conditions for all devices at all signal values and rates of change

Controller output port requirements
- Differential-mode and common-mode static voltage and current drive capability
- Differential-mode and common-mode voltage and current slew-rate cababliity

Controller requirements can be defined for specified signals using a nullor as controller.

Application of OpAmp controllers
- Voltage slew rate
- Maximum load current
- Voltage headroom (difference between +/- output voltage and +/- power supply voltage)

Transistor-level design of controllers
- Output stage device type selection (PCA technology)
- Output device design (IC technology)

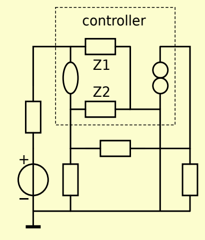
Static (gain) inaccuracy

Contributions to static (gain) inaccuracy

- Inaccuracy in transfer feedback networks.
- Impedances between the terminals of the controller and the ground.
- Finite DC loop gain: $\delta_{A_f} \approx \frac{1}{L_{DC}}$

$$A_f = A_{f\infty}\frac{-L}{1-L} + \frac{\rho}{1-L}$$

$$A_f \approx A_{\mathrm{ideal}}\frac{-L}{1-L}$$



Adverse effects of Z1 and Z2 on the inaccuracy and the weak nonlinearity of the gain cannot be reduced by increasing the loop gain.

Weak nonlinearity

Contributions to weak nonlinearity

- Weak nonlinearity in the transfer of the feedback networks
- Elements between the terminals of the controller and the ground.
- Weak nonlinearity of the loop gain:

$$\epsilon_{A_f}(e_\ell) \approx \frac{\epsilon_L(e_\ell)}{1-L_Q}$$

Differential error to gain ratio

- The controller should contribute as little as possible to the differential error to gain ratio of the loop gain.
- Increasing the loop gain without increasing its differential gain error reduces the contribution of the controller to the differential gain error of the amplifier.

Transistor-level controller design
A fgure Of Merit for a stage (in the feedback loop):
The smallest possible differentail error to gain ratio

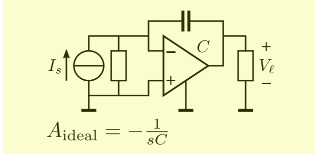
Bandwidth

Contributions to dynamic behavior of the transfer

Feedback network
- The feedback network fixes the intended dynamic behavior of the amplifier

Loop gain
- Finite static loop gain combined with poles and zeros of the loop gain

Parasitic impedances
- Impedances at between the terminals of the controller and the ground.



Desired frequency characteristic and bandwidth

- **The desired freqyency characteristic** of a negative feedback amplifier is the frequency characteristic fixed by its feedback network.
- **The bandwidth of the servo function** is a measure for the frequency range over which the gain approximates the ideal gain.

$$A_f(s) \approx A_{\mathrm{ideal}}(s)\frac{-L(s)}{1-L(s)}$$

$$S(s) = \frac{-L(s)}{1-L(s)}$$

Servo bandwidth with all-pole loop gain function

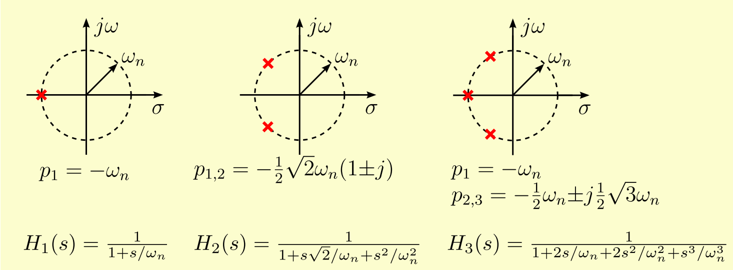
All-pole loop gain function
- Finite DC loop gain
- n poles

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



-3dB bandwidth of n-th order MFM filter

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

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

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

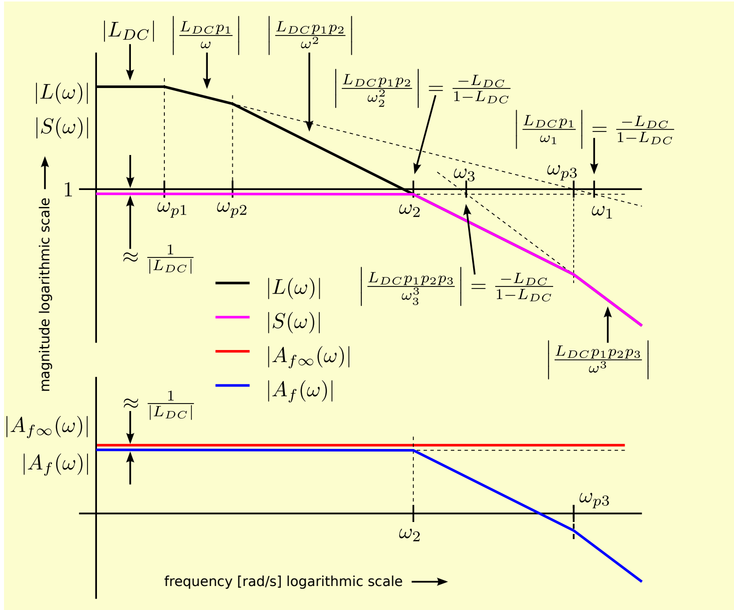
Loop gain-poles product (LP product)

- 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.

Application of OpAmp controllers
- Parasitic capacitances at the input and output terminals
- DC voltage gain, input and output resistance
- Poles and zeros of the gain (GB product for first-order approximations)

Transistor-level controller design
- Stages that maximally contribute to the LP product
- Maximum contribution of a transistor stage to the LP product is its cut-off frequency ω_T

Determination of LP product



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. Their frequency should be at least two times the MFM frequency, obtained according to the above procedure.

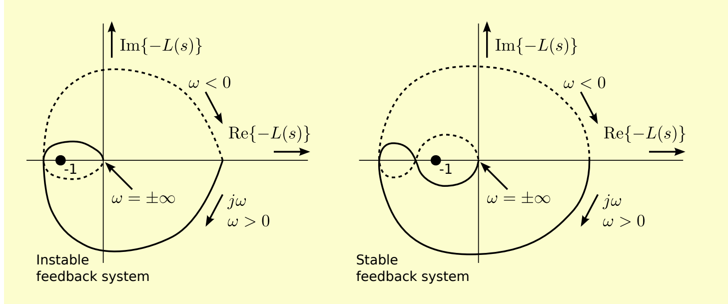
Frequency stability

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.

Methods for stability analysis

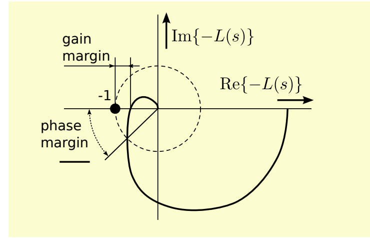
Routh-Hurwitz stability criterion
- Mathematical test using the Routh Array

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



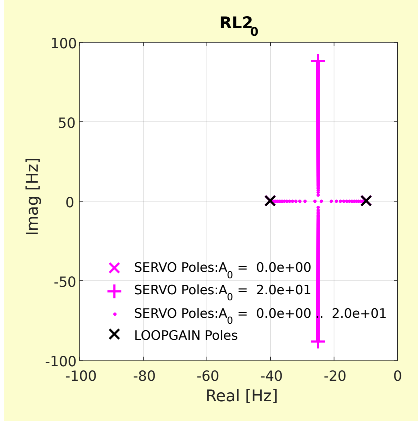
Gain margin and phase margin
Simple stability criterion for all-pole second-order feedback systems

- 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.



From the root locus plots we can learn that all-pole, higher order, negative feedback systems may have poles in the right half plan and be instable. Insertion of zeros in the left half plane may be benificial for the stability.

Frequency compensation techniques will be applied to obtain the desired pole-zero pattern without affecting the bandwidth.

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