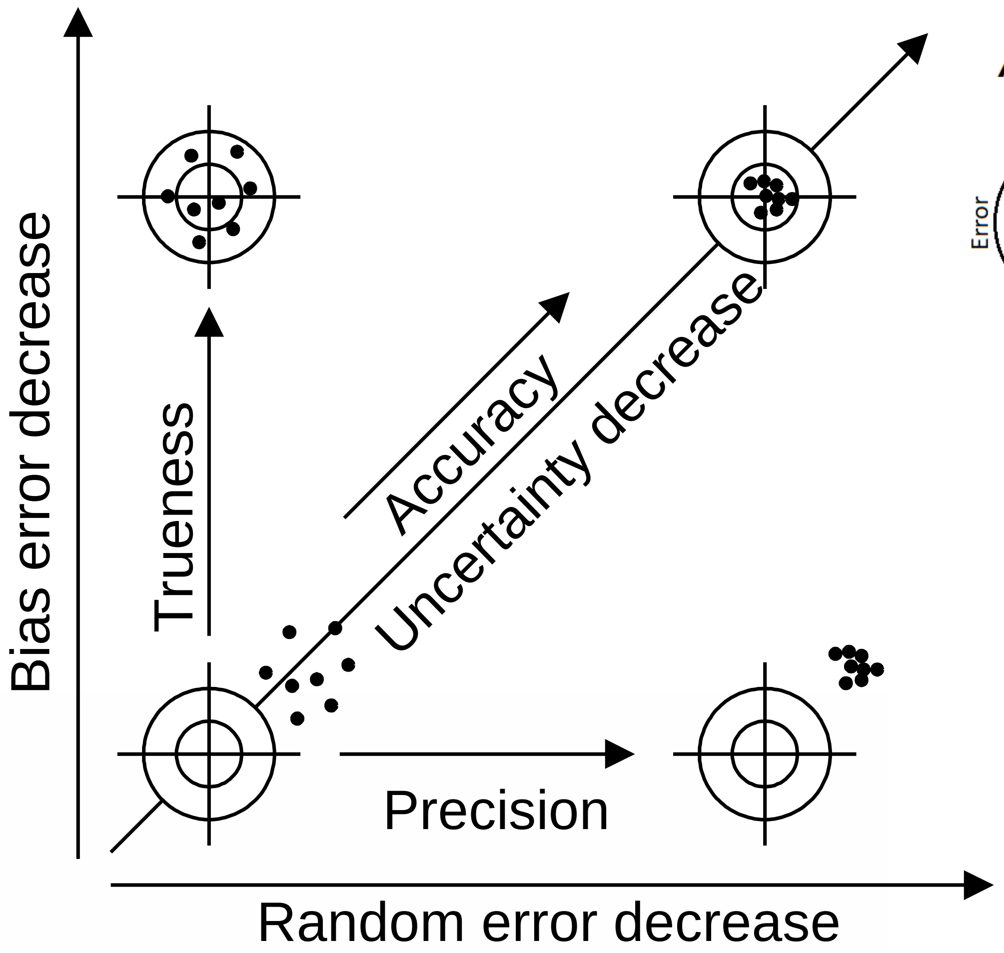
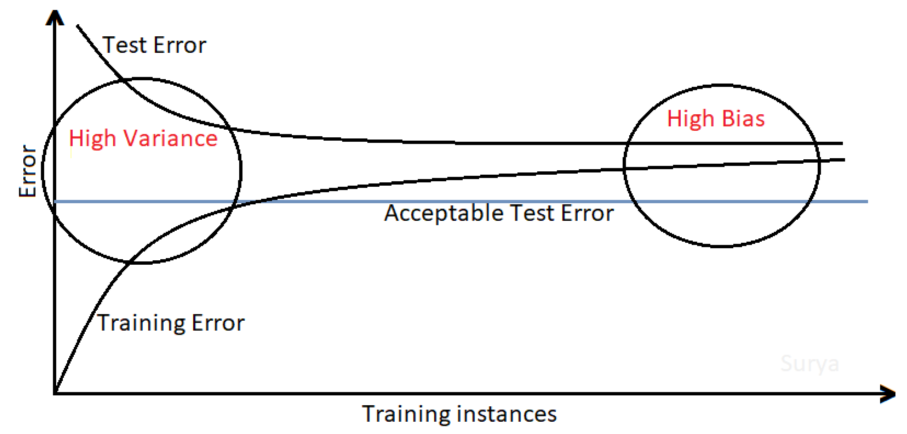
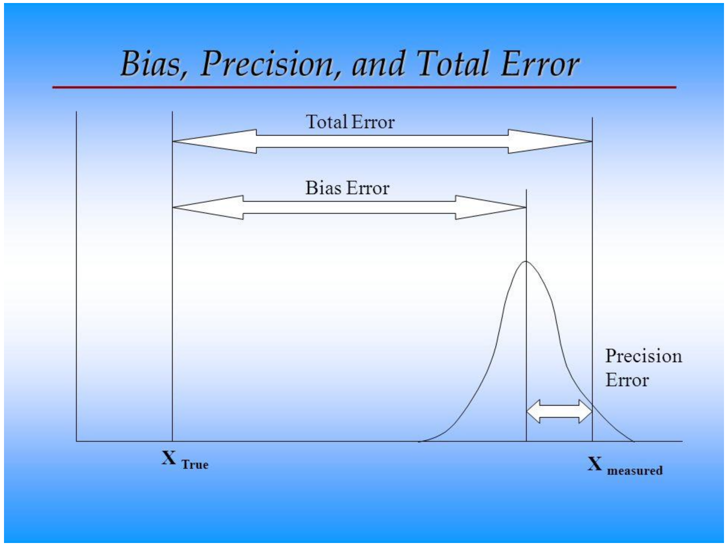
Estimation problem: Amplitude

Prof. L. Svilainis

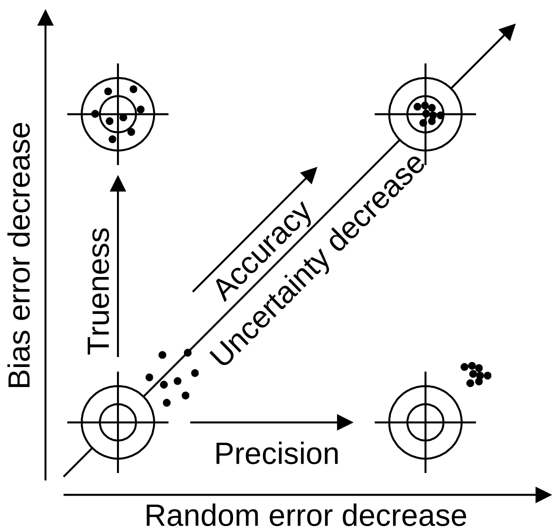
Measurement errors

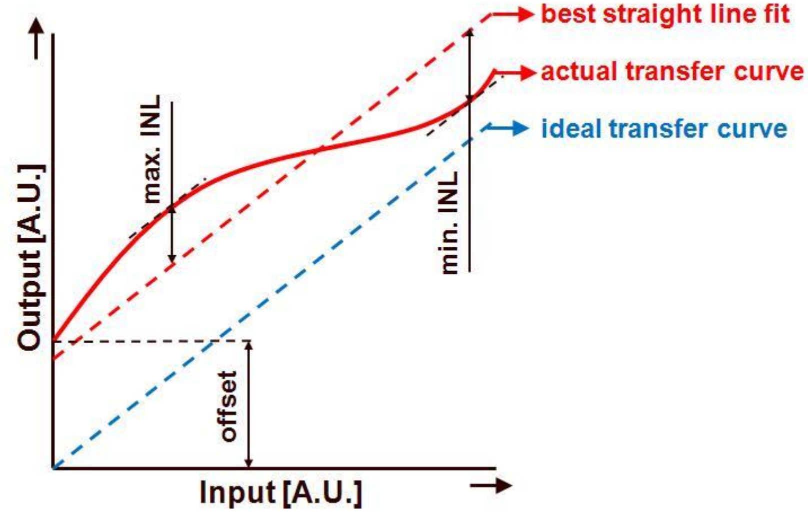


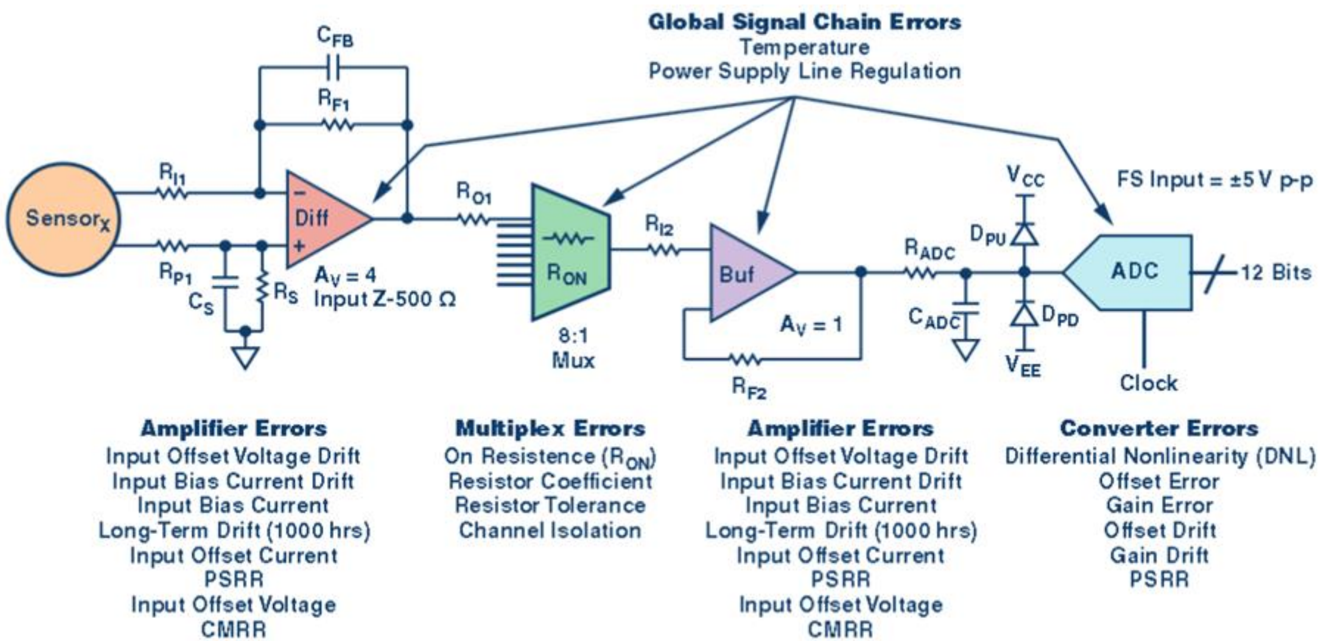




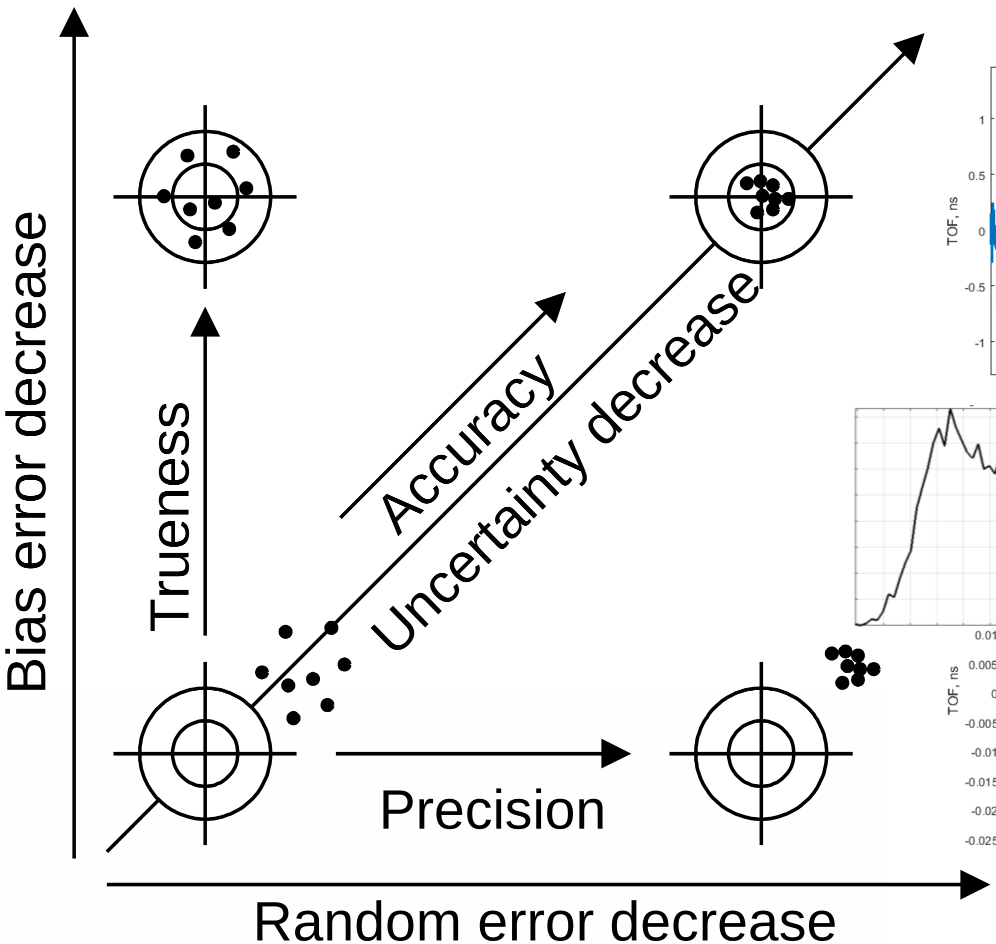
Bias errors

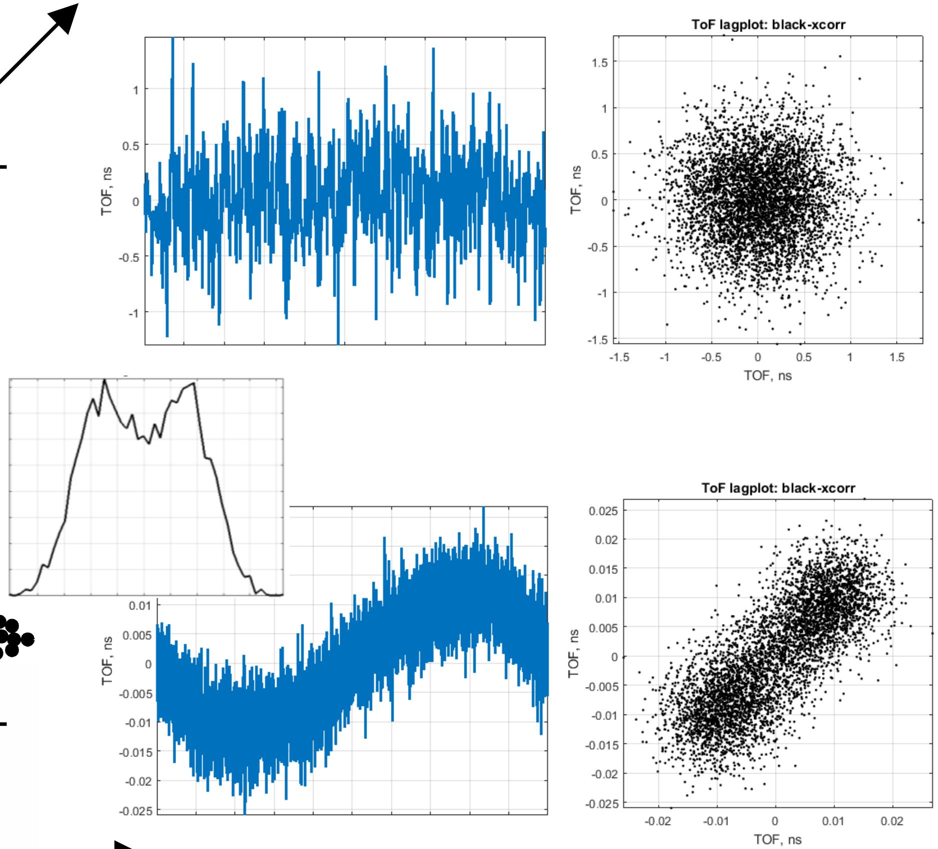

Bias error decrease
Trueness
Accuracy
Uncertainty decrease
Precision
Random error decrease

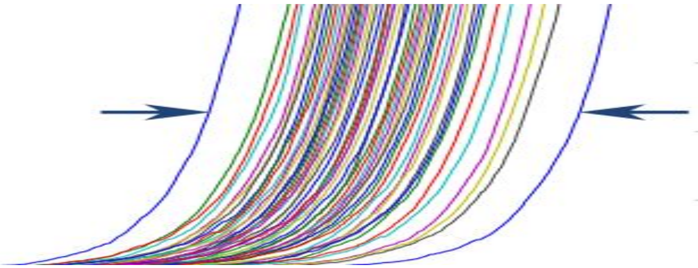

best straight line fit
actual transfer curve
ideal transfer curve
Output [A.U.]
Input [A.U.]
max. INL
min. INL
offset


Global Signal Chain Errors
Temperature
Power Supply Line Regulation
Sensor$_X$
R_{I1}
R_{P1}
C_S
R_S
C_{FB}
R_{F1}
Diff
$A_V = 4$
Input Z-500 Ω
R_{O1}
R_{ON}
8:1
Mux
R_{I2}
Buf
$A_V = 1$
R_{F2}
R_{ADC}
C_{ADC}
V_{CC}
D_{PU}
D_{PD}
V_{EE}
FS Input = ±5 V p-p
ADC
12 Bits
Clock
Amplifier Errors
Input Offset Voltage Drift
Input Bias Current Drift
Input Bias Current
Long-Term Drift (1000 hrs)
Input Offset Current
PSRR
Input Offset Voltage
CMRR
Multiplex Errors
On Resistence (R_{ON})
Resistor Coefficient
Resistor Tolerance
Channel Isolation
Amplifier Errors
Input Offset Voltage Drift
Input Bias Current Drift
Input Bias Current
Long-Term Drift (1000 hrs)
Input Offset Current
PSRR
Input Offset Voltage
CMRR
Converter Errors
Differential Nonlinearity (DNL)
Offset Error
Gain Error
Offset Drift
Gain Drift
PSRR

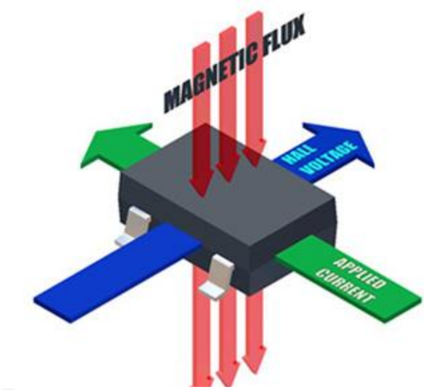
Random errors



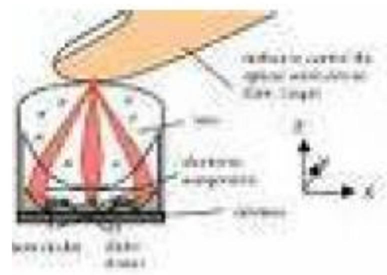


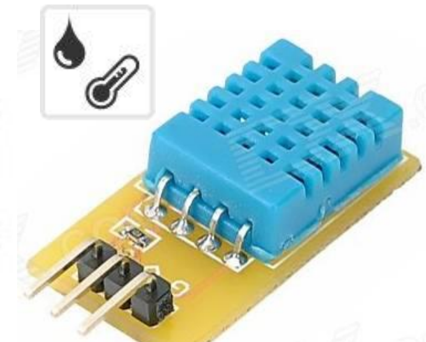


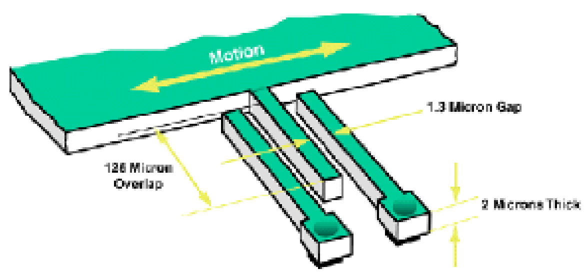
Sensors

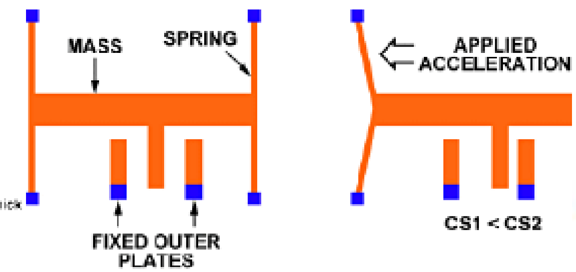

MAGNETIC FLUX
HALL VOLTAGE
APPLIED CURRENT

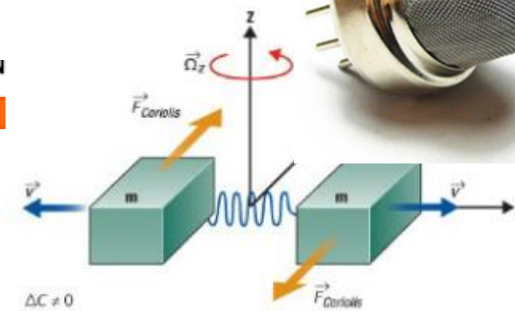

Infineon
3D Magnetic Sensor

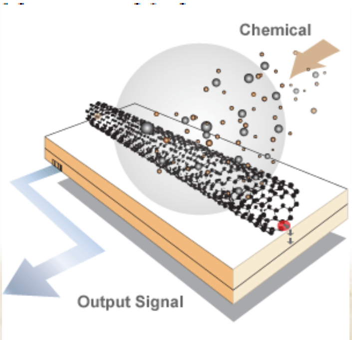


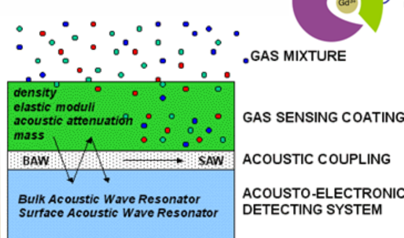


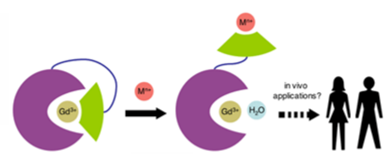

Motion
1.3 Micron Gap
126 Micron Overlap
2 Microns Thick

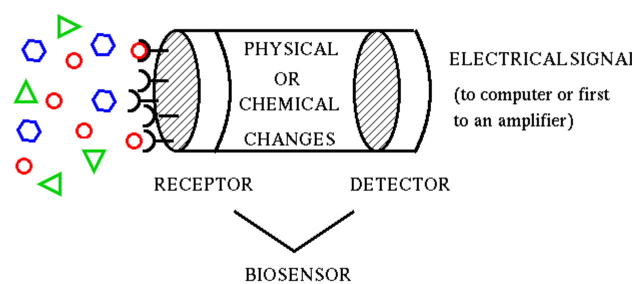

MASS
SPRING
FIXED OUTER PLATES
APPLIED ACCELERATION
CS1 < CS2


z
m
m
x
ΔC ≠ 0

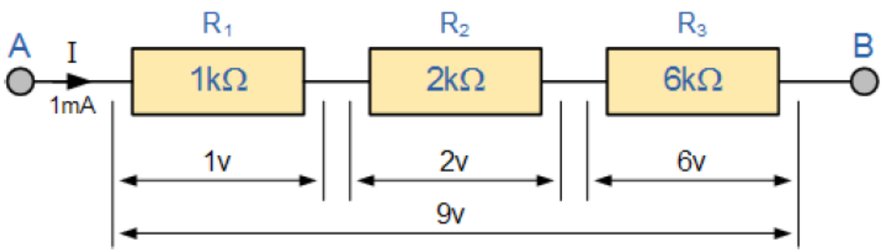

Chemical
Output Signal

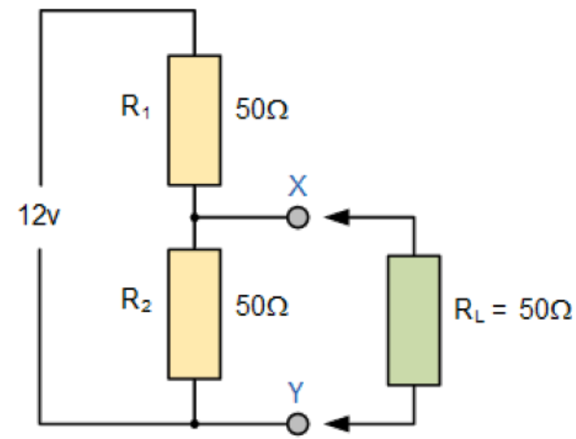

GAS MIXTURE
density
elastic moduli
acoustic attenuation
mass
GAS SENSING COATING
BAW
SAW
ACOUSTIC COUPLING
Bulk Acoustic Wave Resonator
Surface Acoustic Wave Resonator
ACOUSTO-ELECTRONIC DETECTING SYSTEM

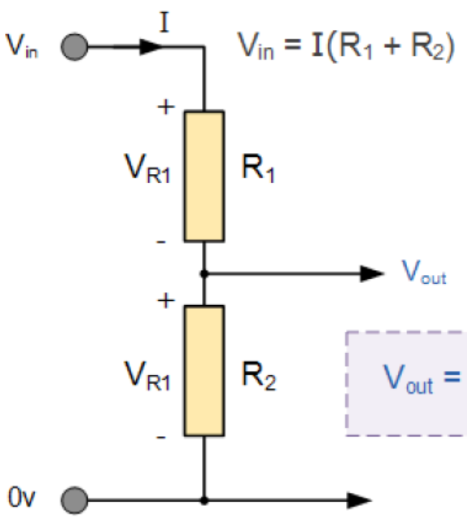



PHYSICAL
OR
CHEMICAL
CHANGES
ELECTRICAL SIGNAL
(to computer or first
to an amplifier)
RECEPTOR
DETECTOR
BIOSENSOR

Amplitude estimation: divider







$$V_{out} = V_{in}\left(\frac{R_2}{R_1 + R_2}\right)$$

a) Without R_L connected

$R_{X\text{-}Y} = 50\Omega$

$$V_{out} = V_{in} \times \frac{R_2}{R_1 + R_2}$$

$$V_{out} = 12v \times \frac{50}{50 + 50} = 6.0v$$

b) With R_L connected

$R_{X\text{-}Y} = 25\Omega$ (Resistors in Parallel)

$$V_{out} = V_{in} \times \frac{R_2}{R_1 + R_2}$$

$$V_{out} = 12v \times \frac{25}{50 + 25} = 4.0v$$

Amplifier

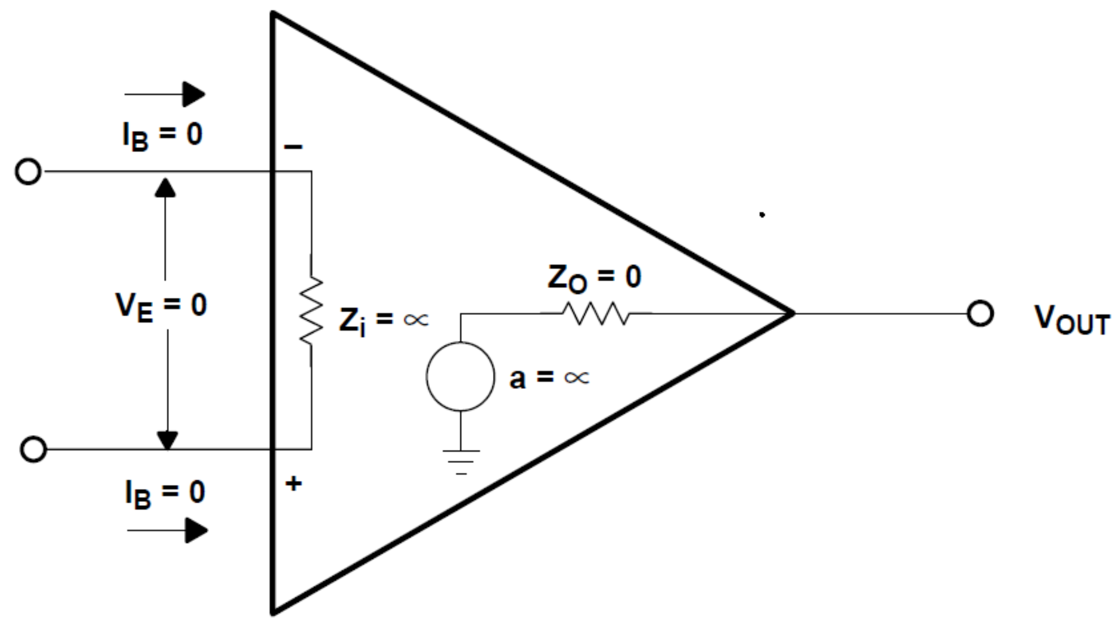
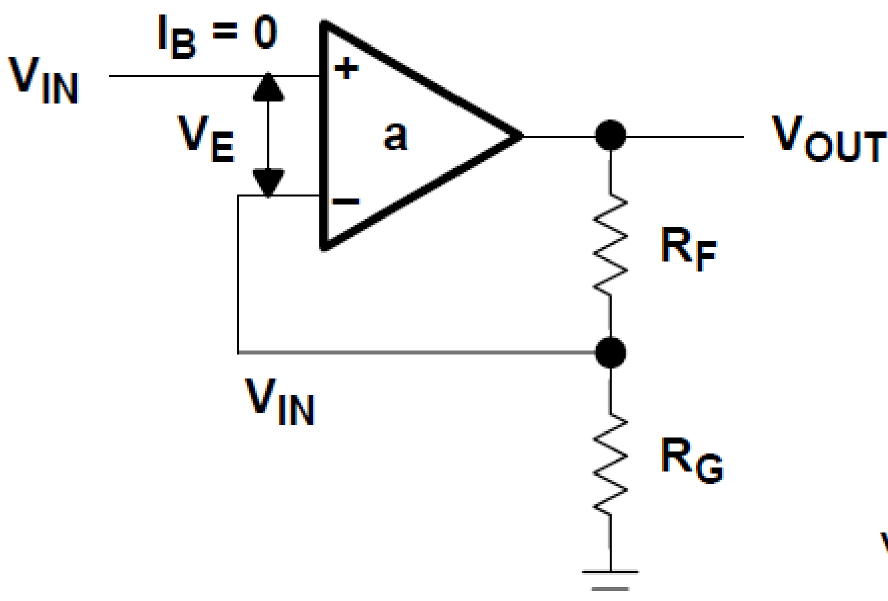


Figure 3–1. The Ideal Op Amp

Amplifier

Noninverting amplifier

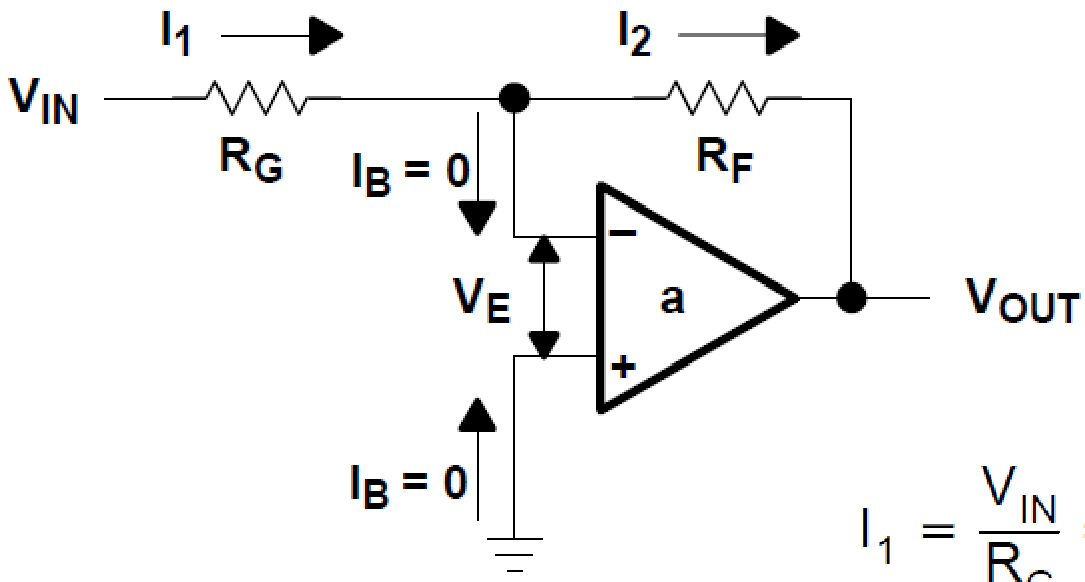


$$V_{IN} = V_{OUT} \frac{R_G}{R_G + R_F}$$

$$\frac{V_{OUT}}{V_{IN}} = \frac{R_G + R_F}{R_G} = 1 + \frac{R_F}{R_G}$$

Amplifier

Inverting amplifier

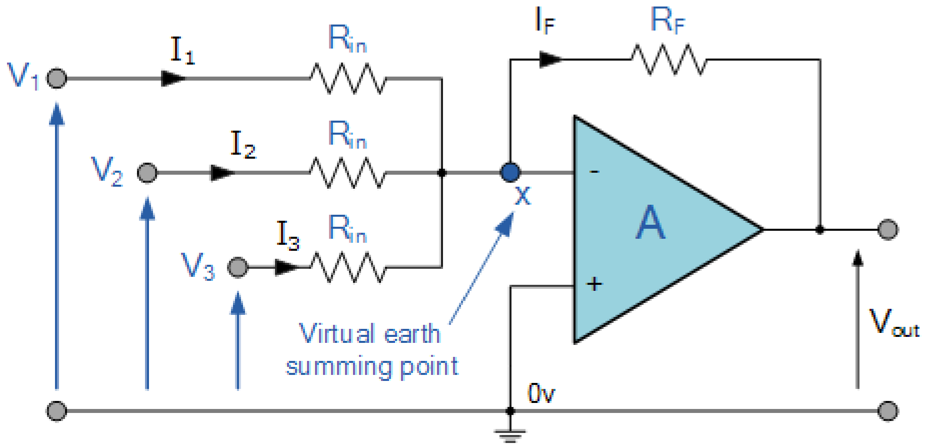


$$I_1 = \frac{V_{IN}}{R_G} = -I_2 = -\frac{V_{OUT}}{R_F}$$

$$\frac{V_{OUT}}{V_{IN}} = -\frac{R_F}{R_G}$$

Amplifier

Current summation



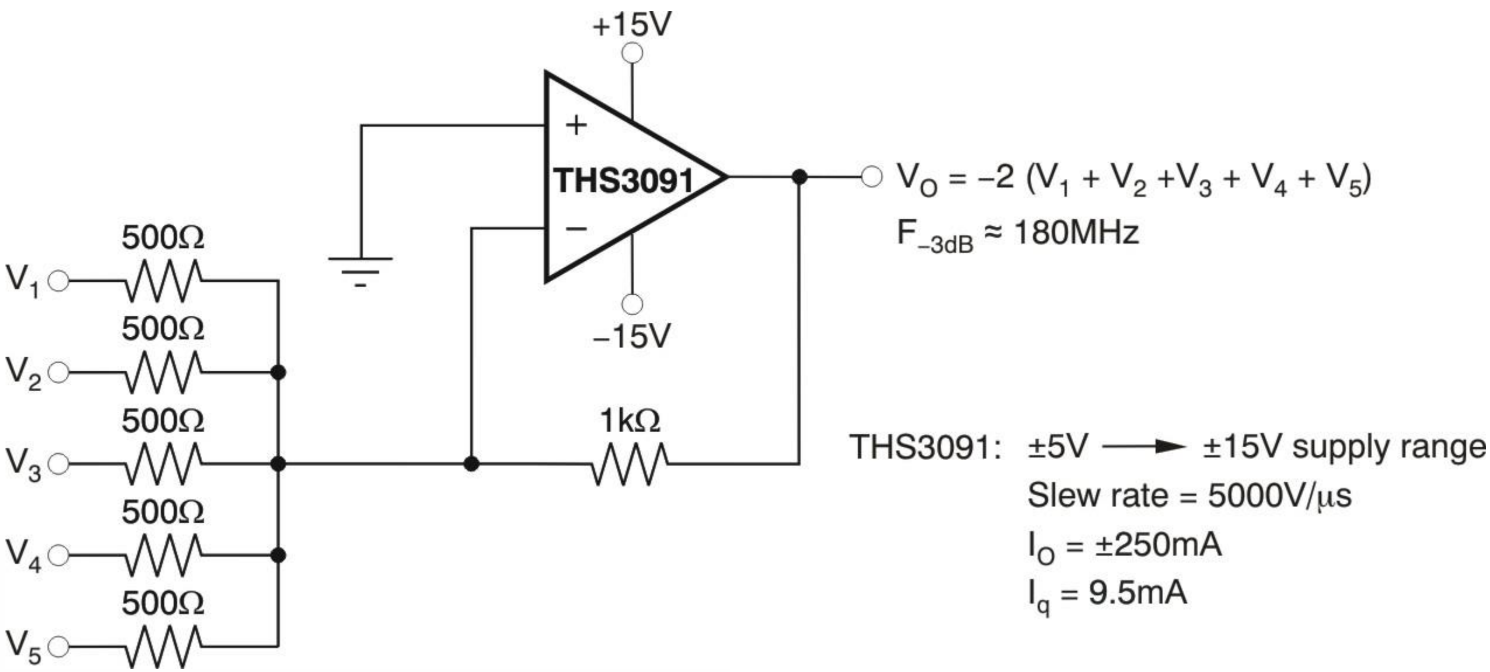
$$I_F = I_1 + I_2 + I_3 = -\left[\frac{V1}{Rin} + \frac{V2}{Rin} + \frac{V3}{Rin}\right]$$

Inverting Equation: $Vout = -\frac{Rf}{Rin} \times Vin$

then, $-Vout = \left[\frac{R_F}{Rin}V1 + \frac{R_F}{Rin}V2 + \frac{R_F}{Rin}V3\right]$

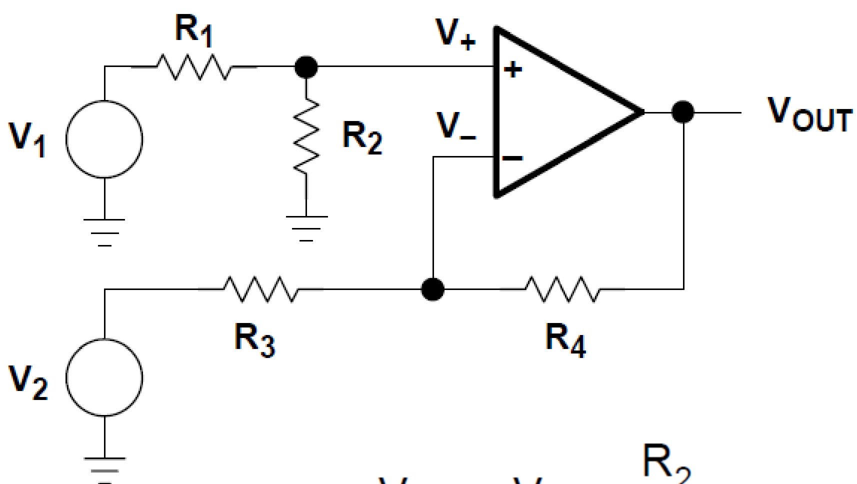
Amplifier

Voltage summation



Amplifier

Differential amplifier



$$R_2 = R_4 \text{ and } R_1 = R_3$$

$$V_{OUT} = (V_1 - V_2)\frac{R_4}{R_3}$$

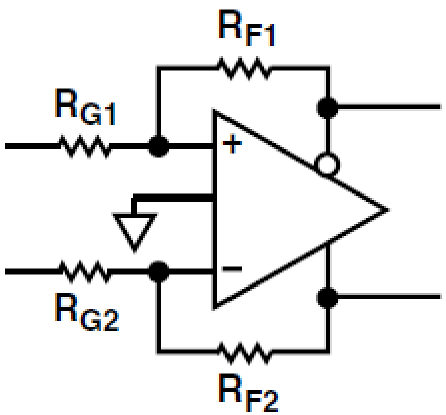
$$V_{OUT2} = V_2\left(-\frac{R_4}{R_3}\right)$$

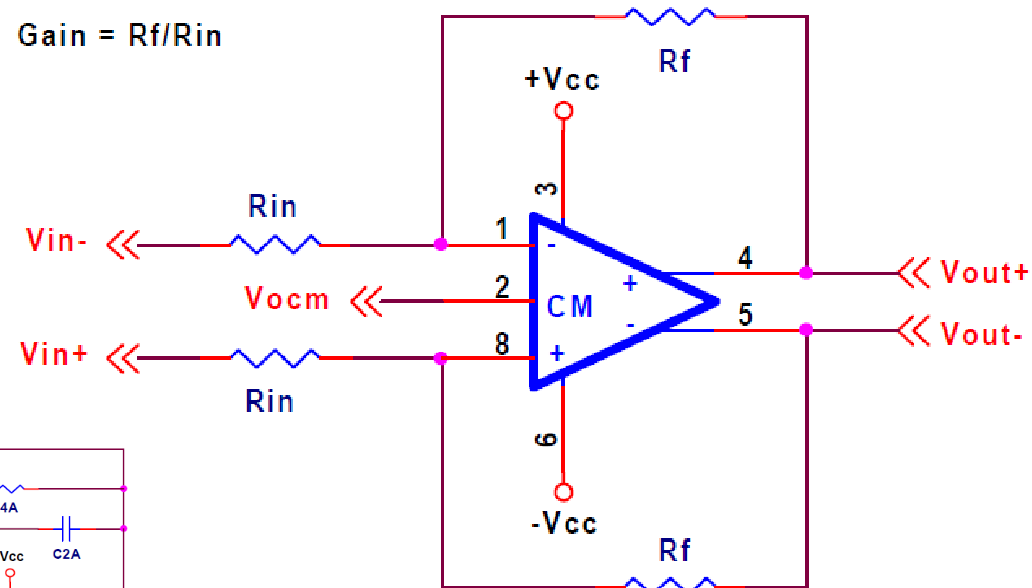
$$V_+ = V_1 \frac{R_2}{R_1 + R_2}$$

$$V_{OUT1} = V_+(G_+) = V_1 \frac{R_2}{R_1 + R_2}\left(\frac{R_3 + R_4}{R_3}\right)$$

Amplifier

Fully differential amplifier





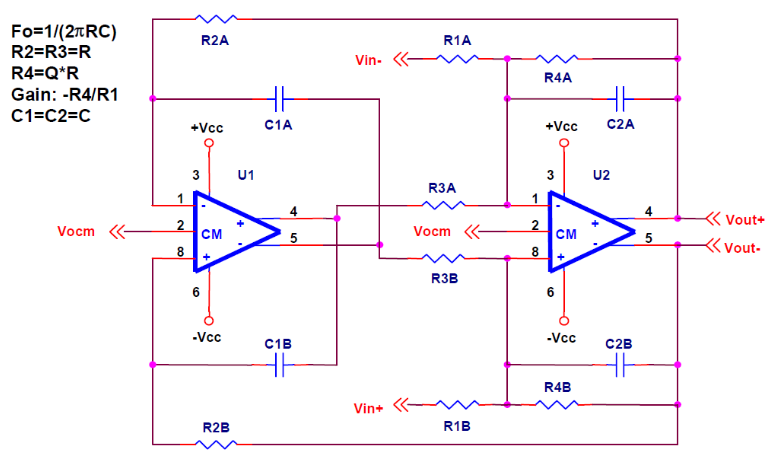
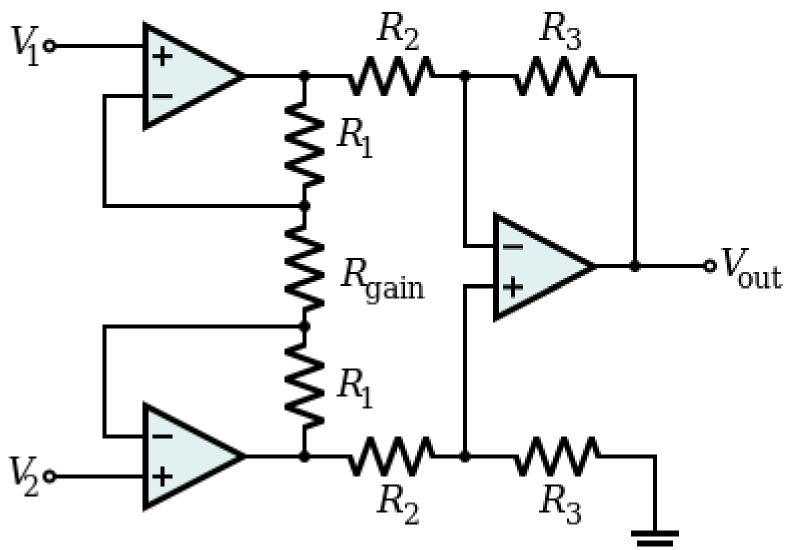


Figure 11: Akerberg Mossberg Band Pass Filter

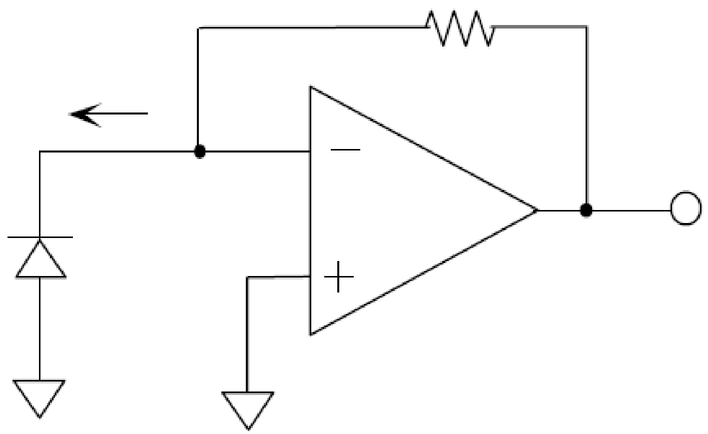
Amplifier

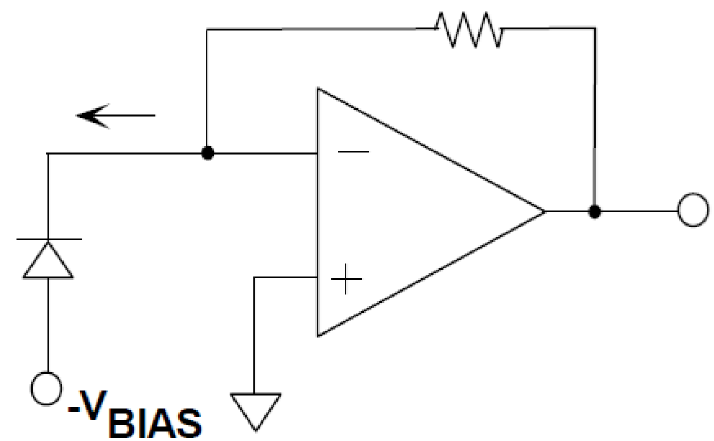
Instrumental amplifier



$$\frac{V_{\text{out}}}{V_2 - V_1} = \left(1 + \frac{2R_1}{R_{\text{gain}}}\right)\frac{R_3}{R_2}$$

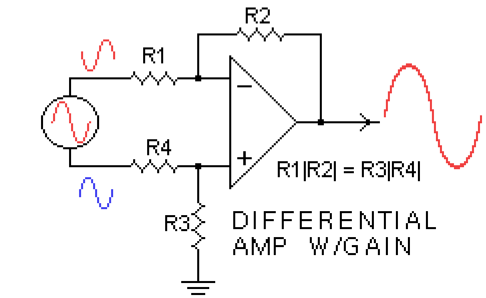
Amplifier

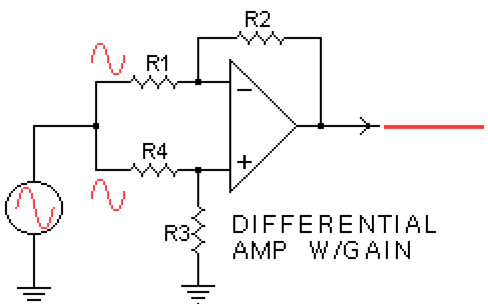




PHOTOVOLTAIC	PHOTOCONDUCTIVE
· Zero Bias	· Reverse Bias
· No Dark Current	· Dark Current Exists
· Precision Applications	· High Speed Applications
· Low Noise (Johnson)	· Higher Noise (Johnson + Shot))

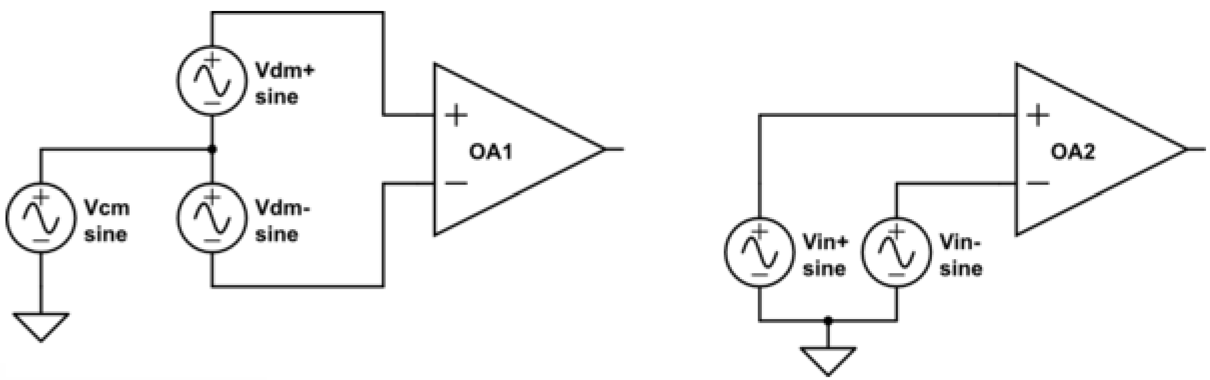
Amplifier





Common-Mode Rejection Ratio

$$CMRR = \Delta V_{COM}/\Delta V_{OS}$$



CMRR, as published in the data sheet, is a dc parameter

when graphed vs. frequency, falls off as the frequency increases.

OpAmps for everyone, TI SLOD006B 2002

Amplifier

Supply Voltage Rejection Ratio

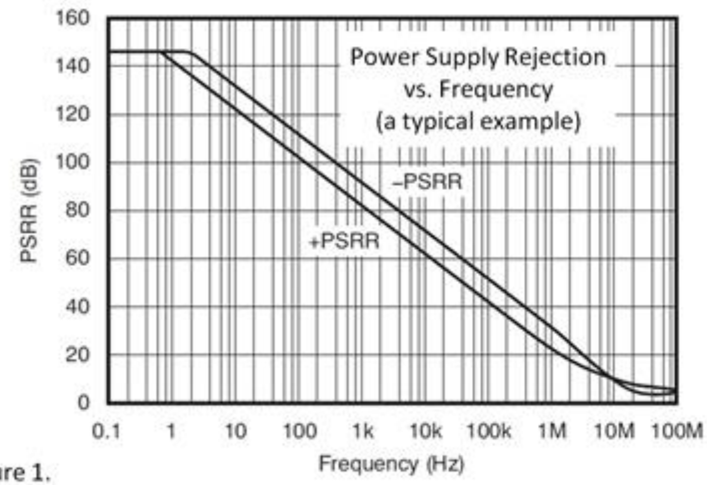
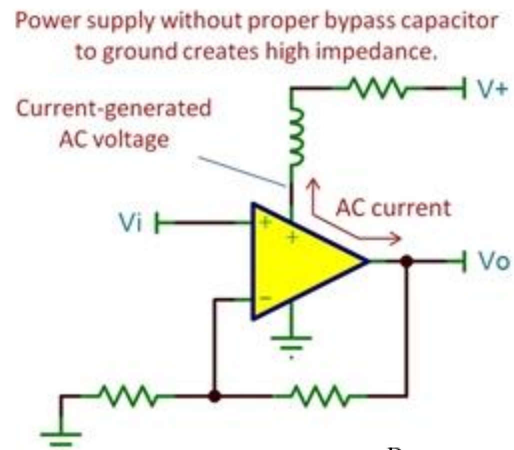
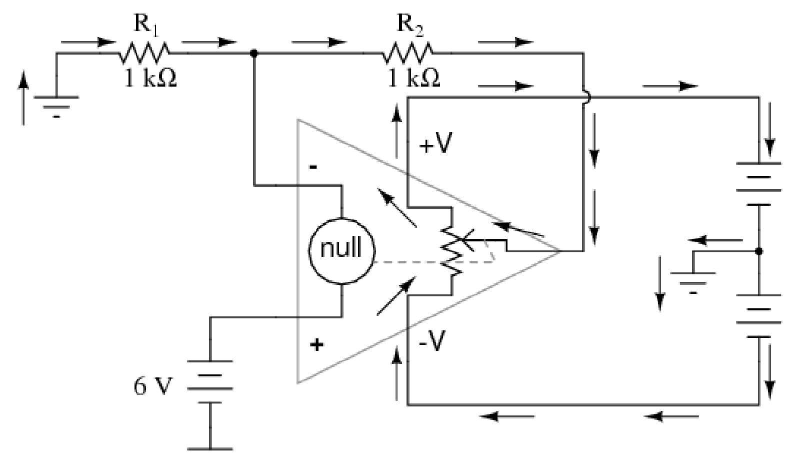


Figure 1.





$$K_{SVR} = \frac{\Delta V_{CC\pm}}{\Delta V_{OS}}$$

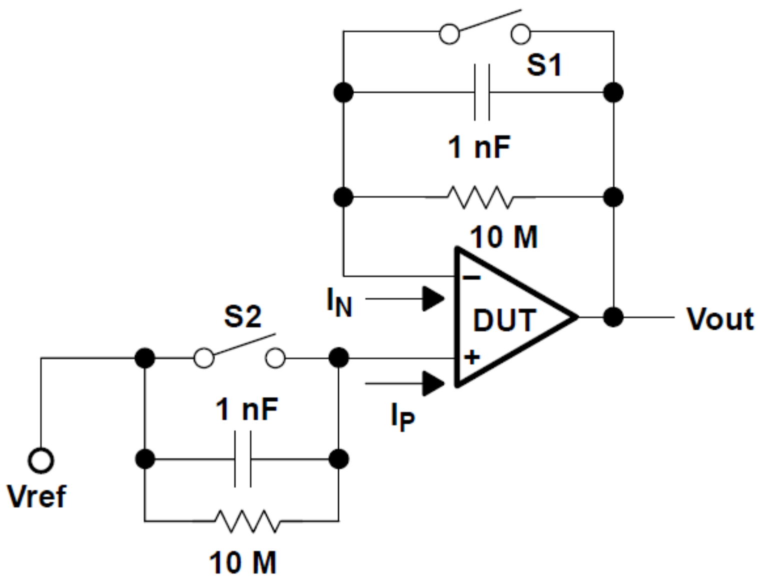
$$K_{SVR} = \frac{\Delta V_{CC}}{\Delta V_{OS}}$$

OpAmps for everyone, TI SLOD006B 2002

Amplifier

ktu 1922

Bias



S1 Closed

$$I_P = \frac{Vout - Vref}{10^7}$$

S2 Closed

$$I_N = \frac{Vout - Vref}{10^7}$$

Test Circuit – I_{IB}

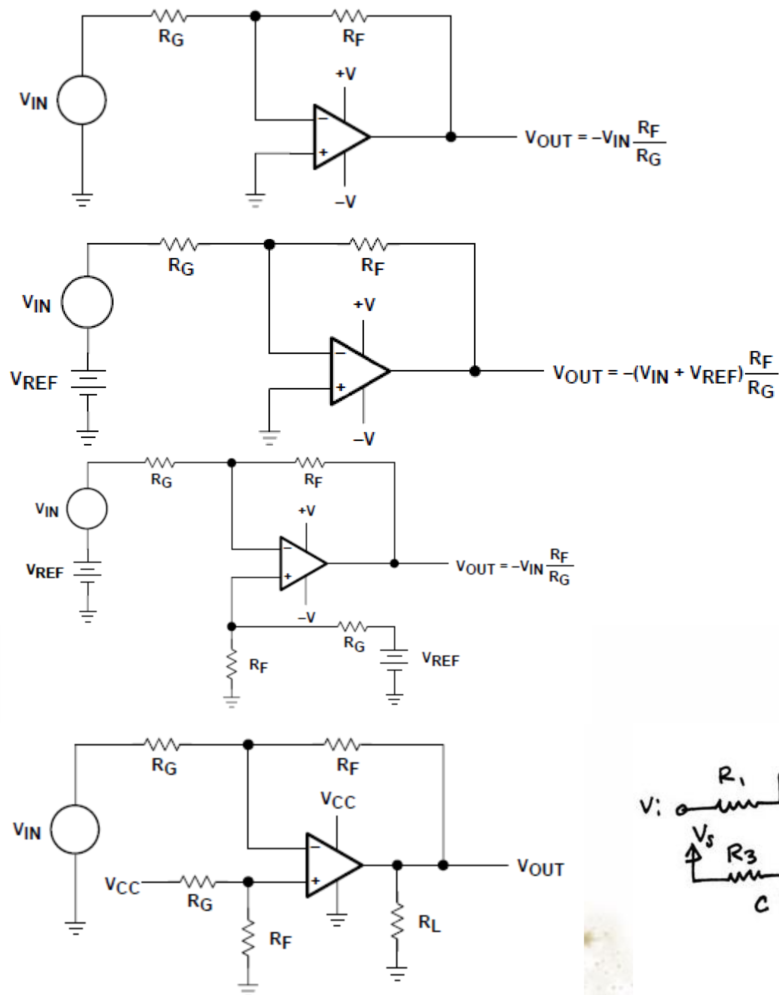
input bias current, I_{IB},

$$I_{IB} = \frac{(I_N + I_P)}{2}$$

input offset current, $I_{IO} = I_N - I_P$.

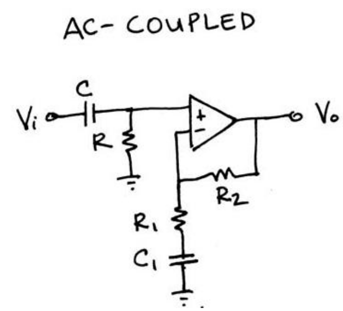
Amplifier

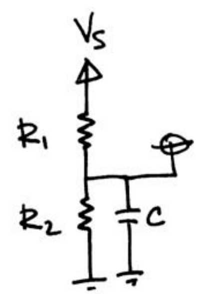
Bias

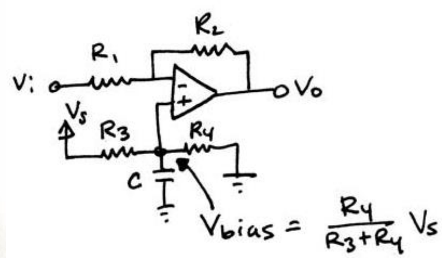


Single-Supply Op Amp Design Techniques

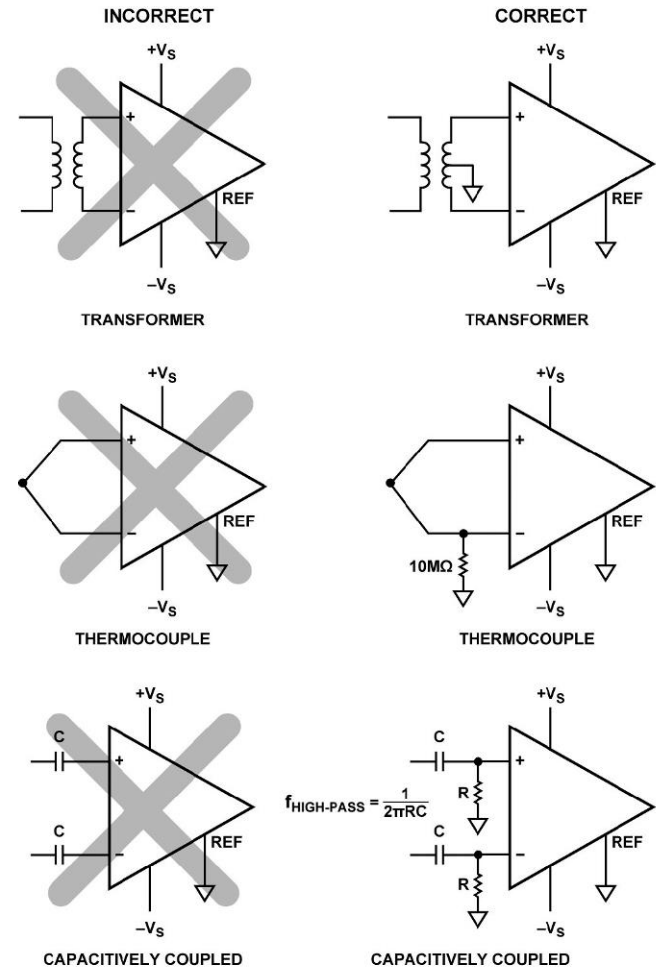
Literature Number SLOA076





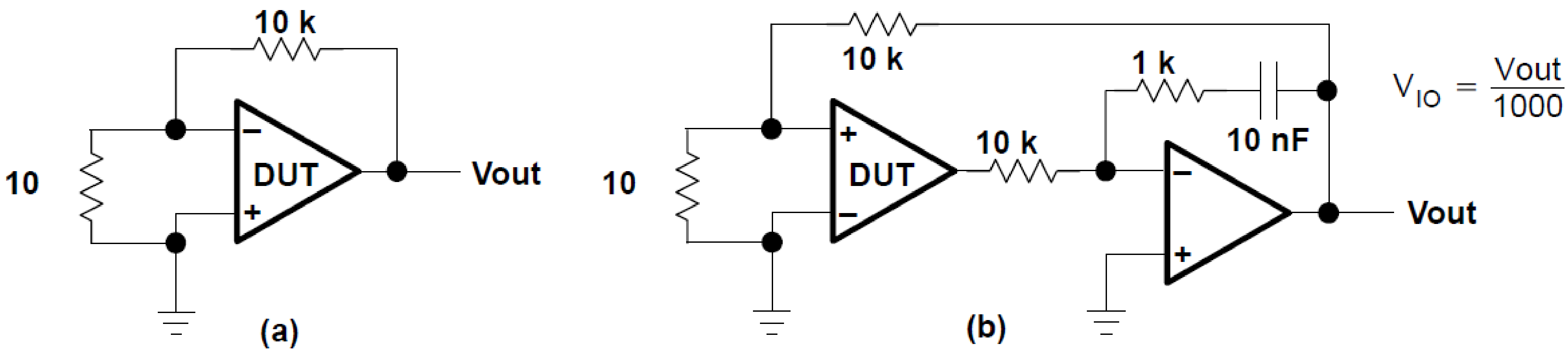


MAS.836
HOW TO BIAS AN OP-AMP

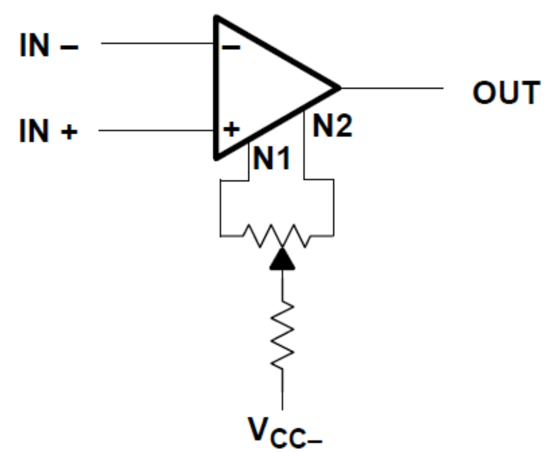


Amplifier

Zero offset



-1. Test Circuits for Input Offset Voltage



Amplifier

Input / output voltage range

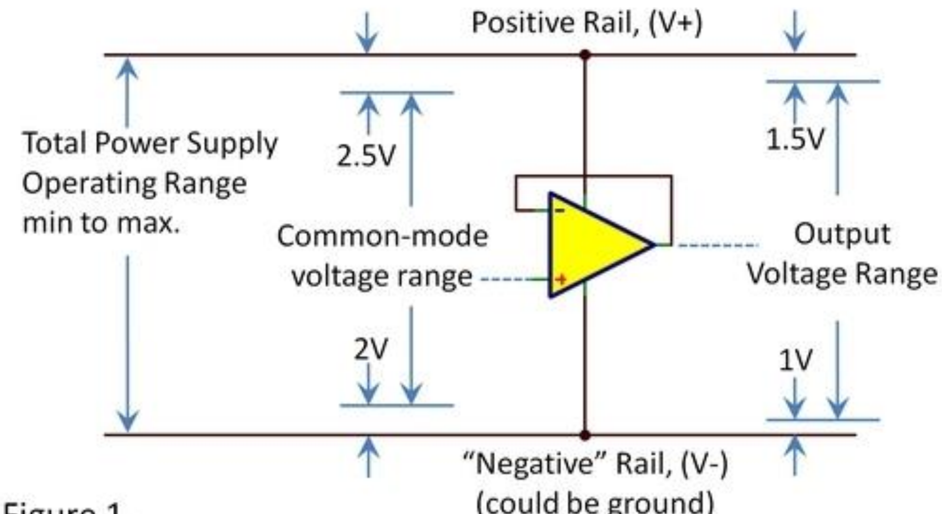


Figure 1.

Voltage range of this G=1 circuit: (V-)+2V to (V+)-2.5V

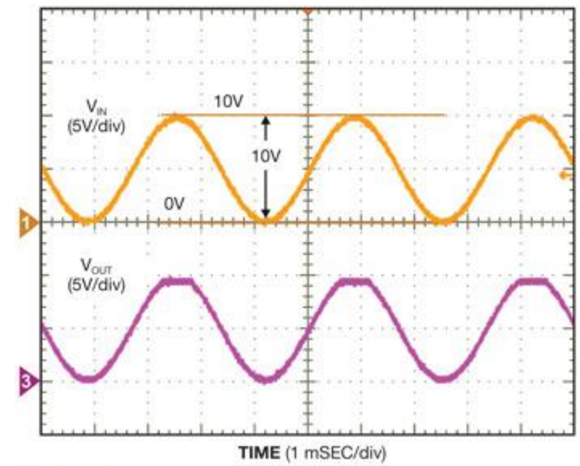


Figure 4 The TLC2272's V_{OUT} shows clipping when V_{IN} (Channel 1) exceeds 9.2V.

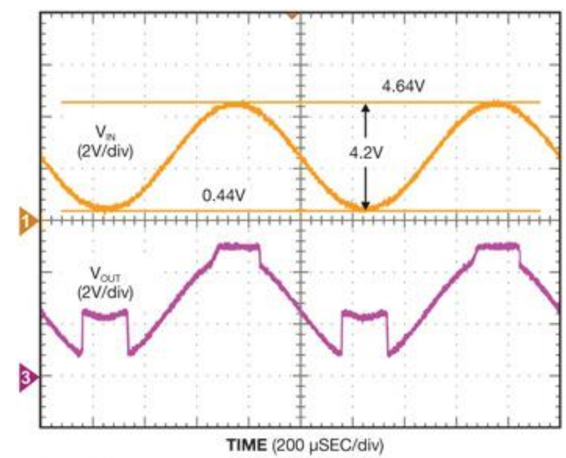
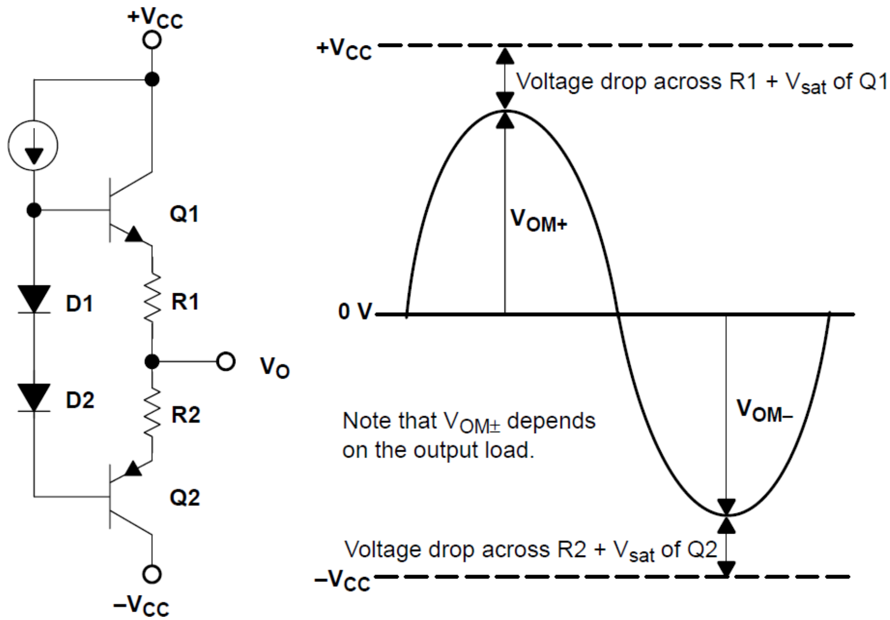


Figure 6 Nonlinear output behavior occurs when the TL971's V_{IN} is 4.2V p-p.

http://www.edn.com/design/analog/4368301/Are-you-violating-your-op-amp-s-input-common-mode-range--item-2

Amplifier

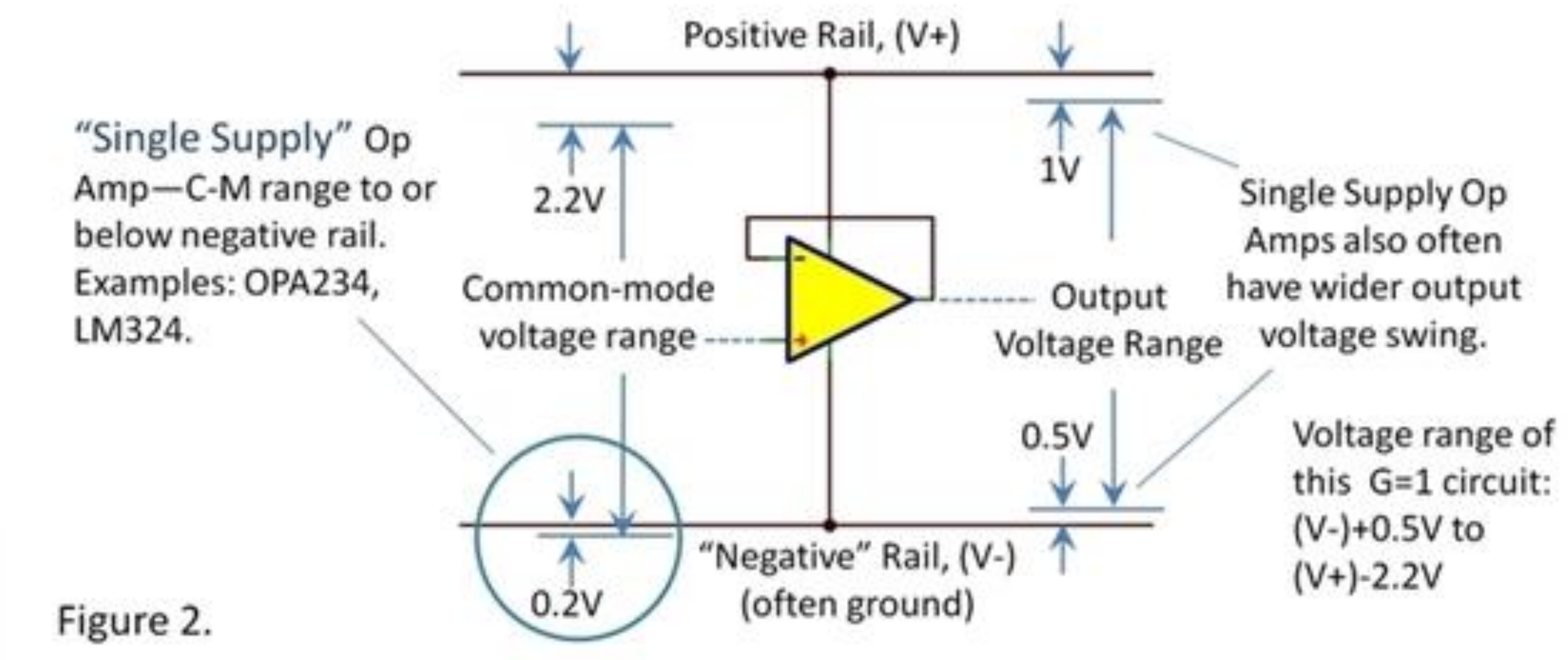
Output stage



Supply current, I_{DD}

Amplifier

Input / output voltage range (SS op amp)



https://e2e.ti.com/blogs_/archives/b/thesignal/archive/2012/05/08/op-amp-voltage-ranges-input-and-output-clearing-some-confusion

Amplifier

Input / output voltage range (RR op amp)

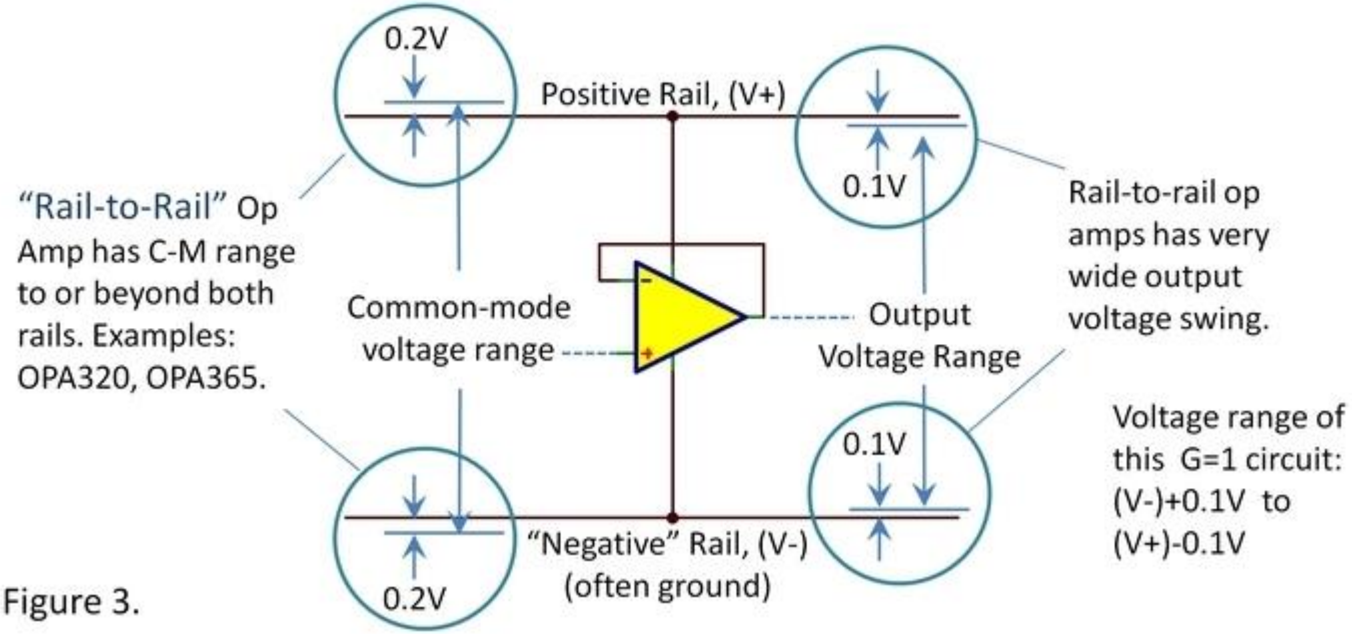
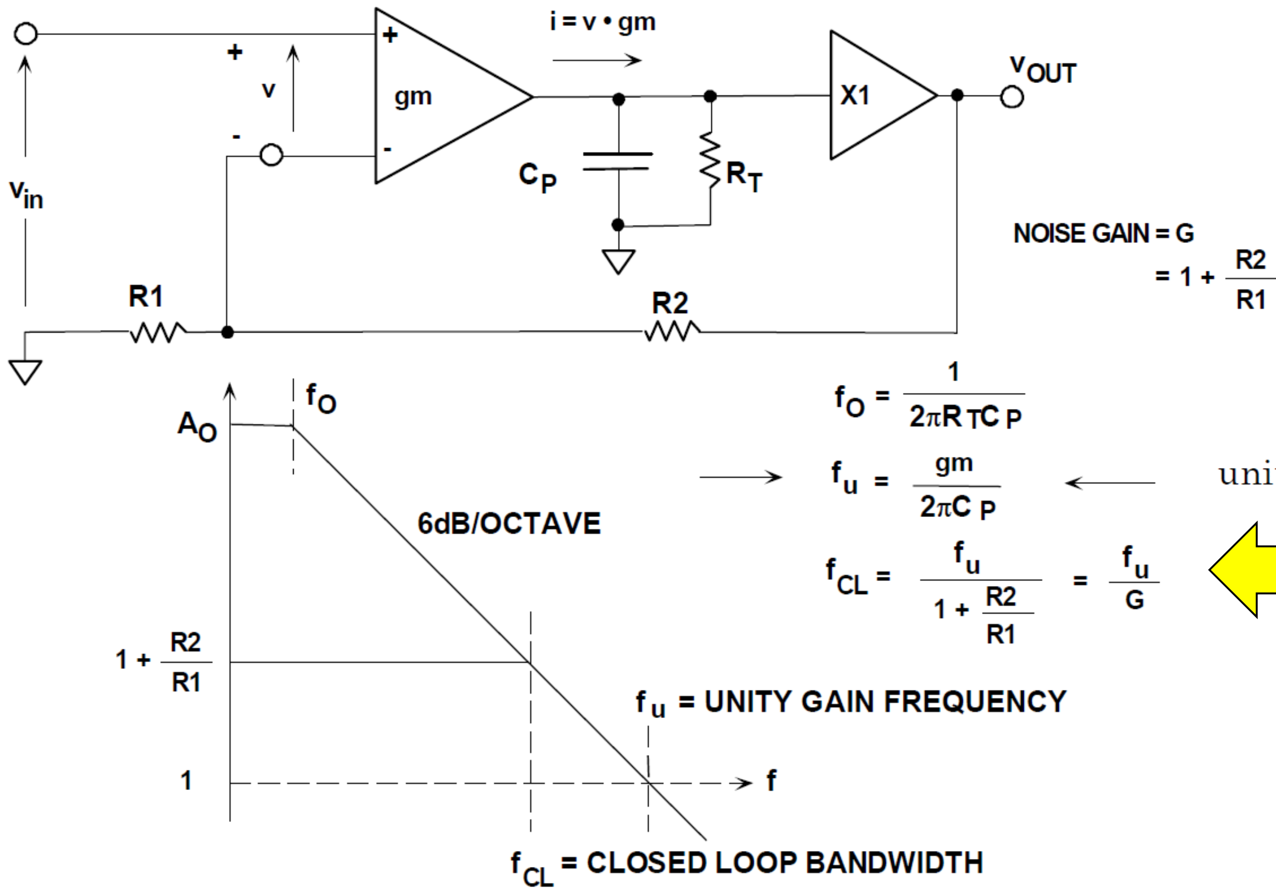


Figure 3.

Amplifier

Small signal bandwidth and gain

MODEL AND BODE PLOT FOR A VFB OP AMP



$$f_O = \frac{1}{2\pi R_T C_P}$$

$$f_u = \frac{gm}{2\pi C_P}$$ ← unity gain-bandwidth

$$f_{CL} = \frac{f_u}{1 + \frac{R2}{R1}} = \frac{f_u}{G}$$

Amplifier

Large signal bandwidth

full-power bandwidth

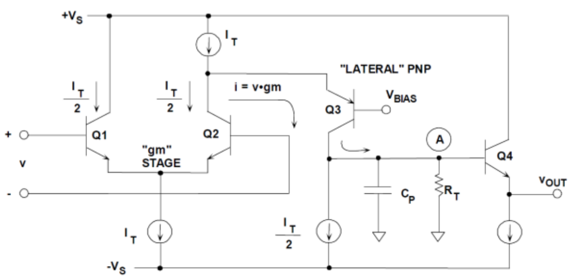
$$\text{FPBW} = \frac{\text{SR}}{2\pi A} = \frac{25\text{V}/\mu\text{s}}{2\pi \cdot 1\text{V}} = 4\text{MHz}$$

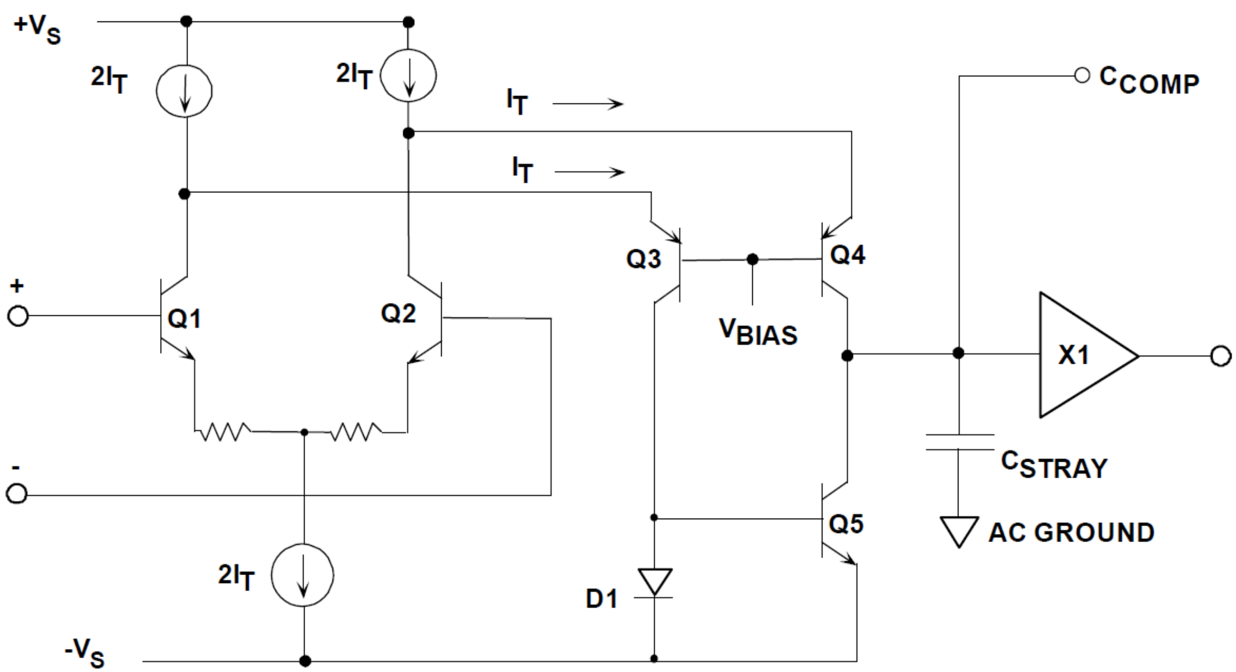
2V peak-to-peak

A is the peak amplitude of the output signal

Amplifier

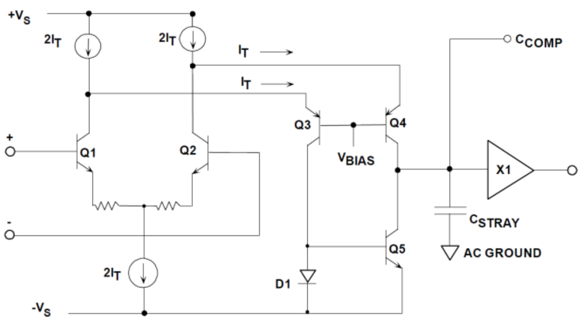
Voltage feedback amplifier

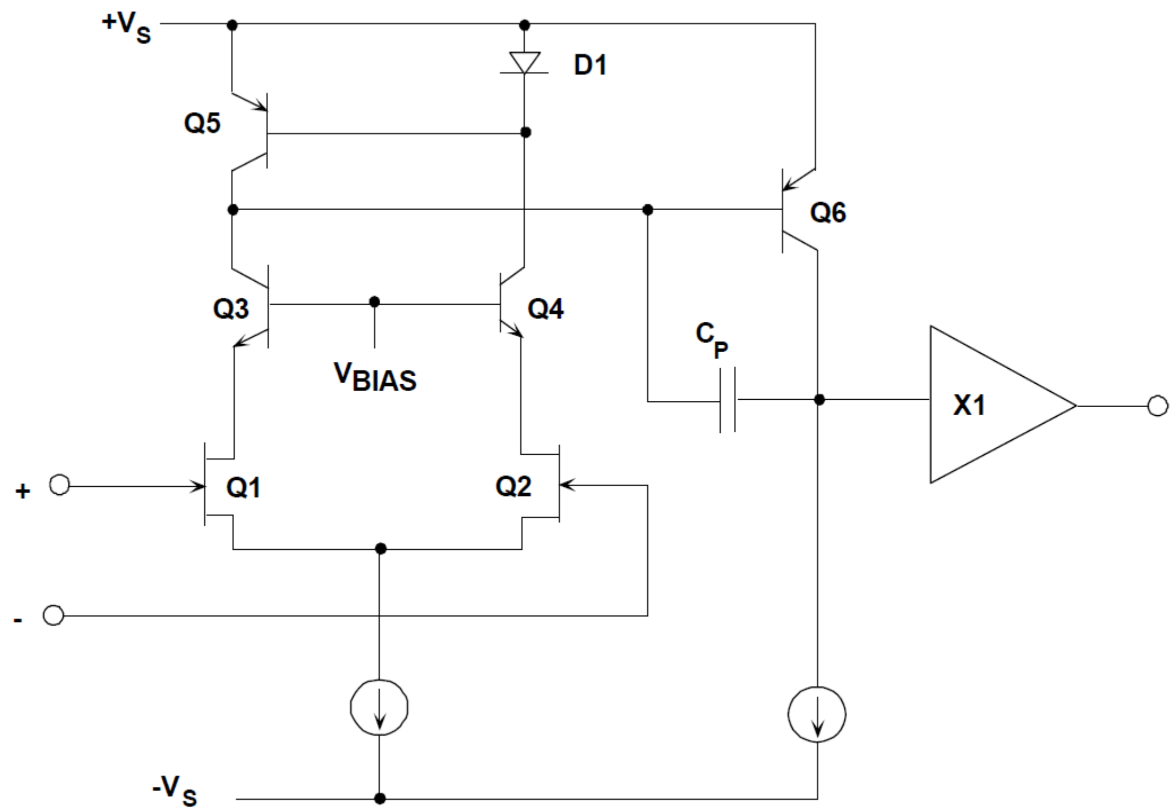




Amplifier

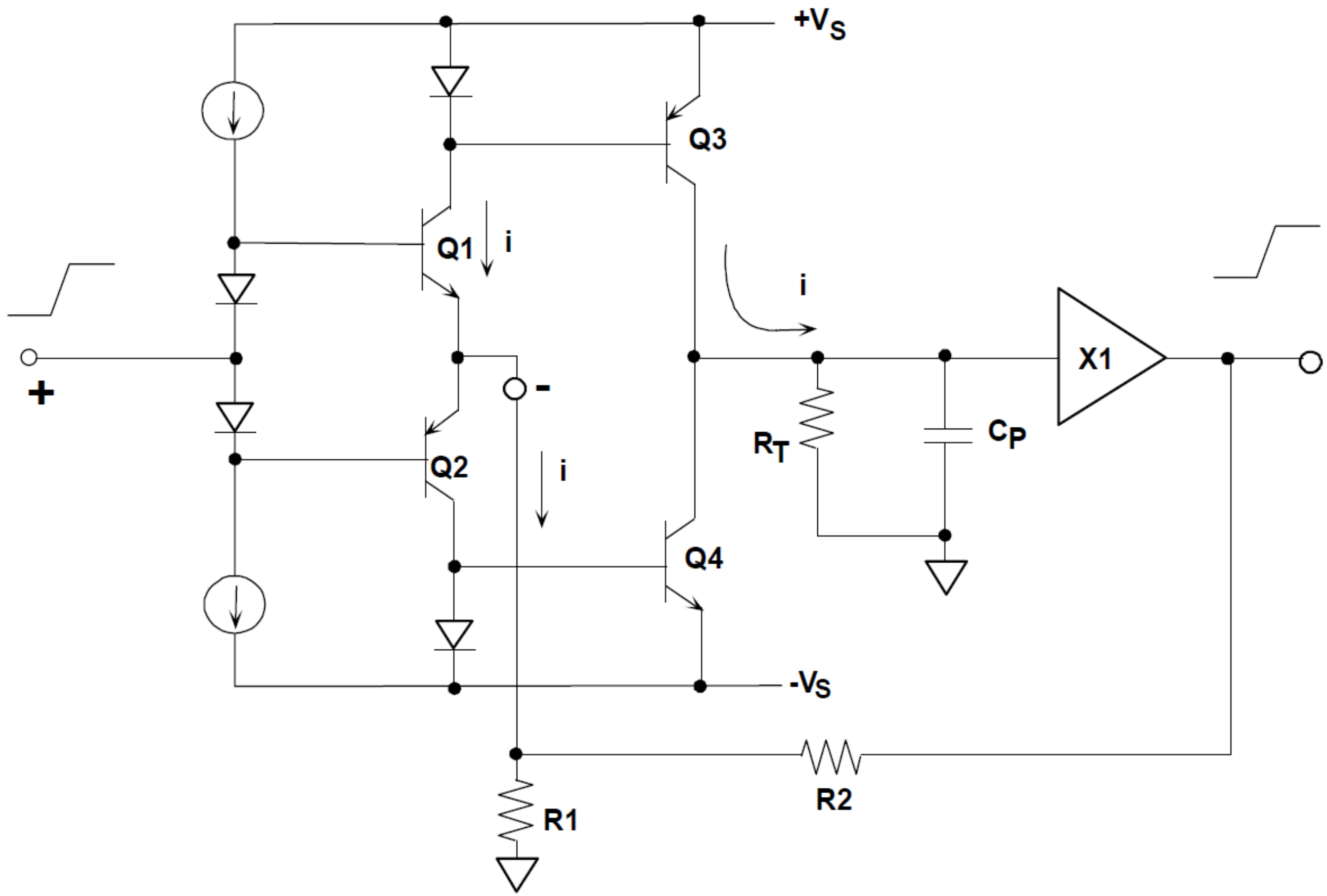
Voltage feedback amplifier: BiFET





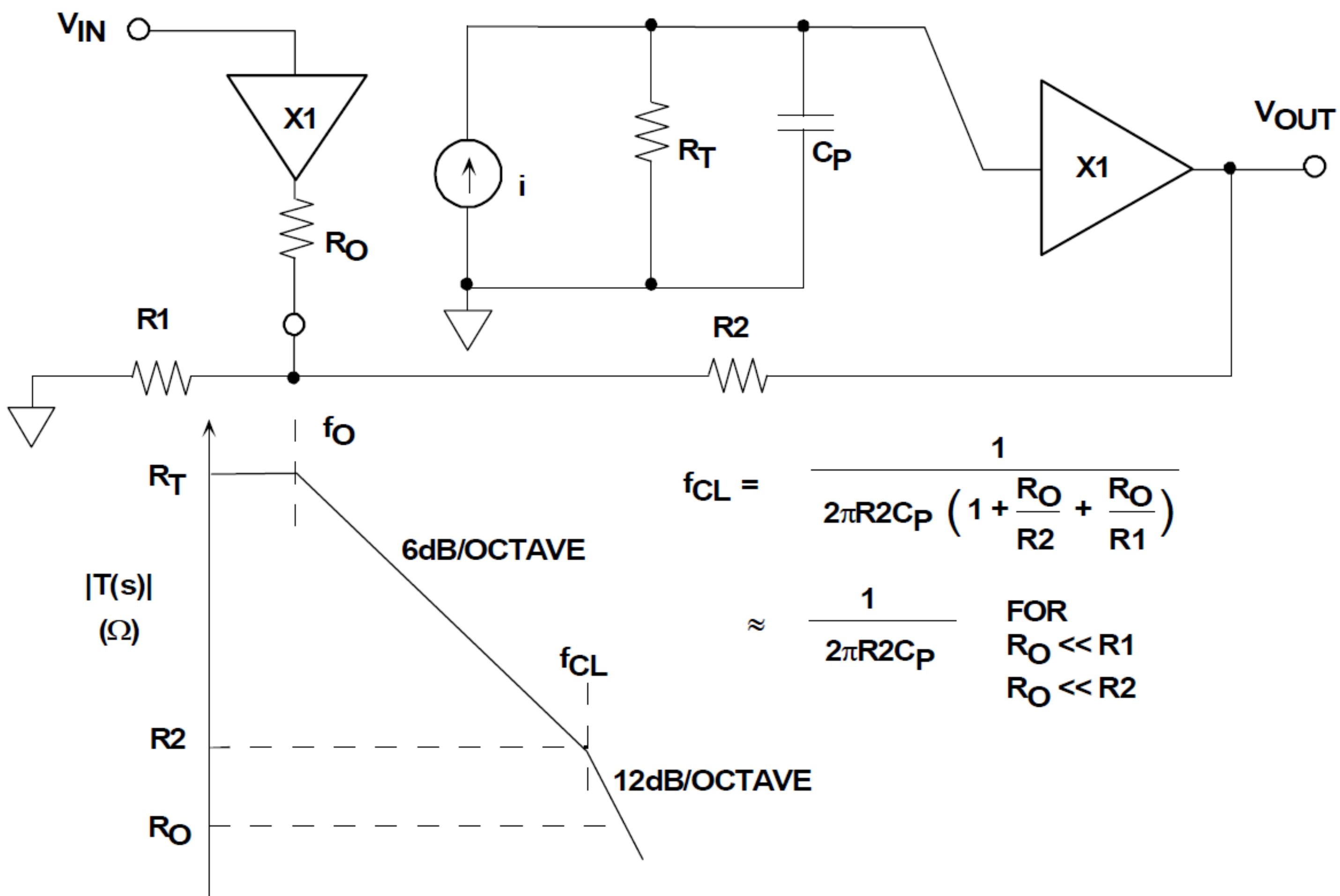
Amplifier

Current feedback amplifier



Amplifier

Current feedback amplifier - CFA

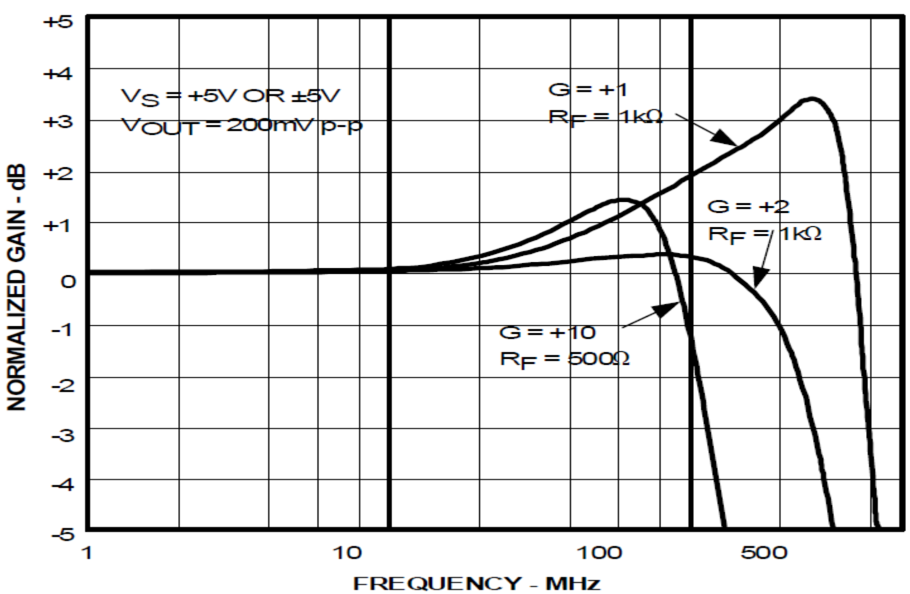


Amplifier

Current feedback amplifier - CFA

keep the amplifier's quiescent power low, yet are capable of supplying *current-on-demand* for wide current excursions required during fast slewing.

AD8011 FREQUENCY RESPONSE
G = +1, +2, +10



Closed-loop bandwidth of a CFB op amp is determined by the internal dominant-pole capacitance and the external feedback resistor, independent of the gain-setting resistor

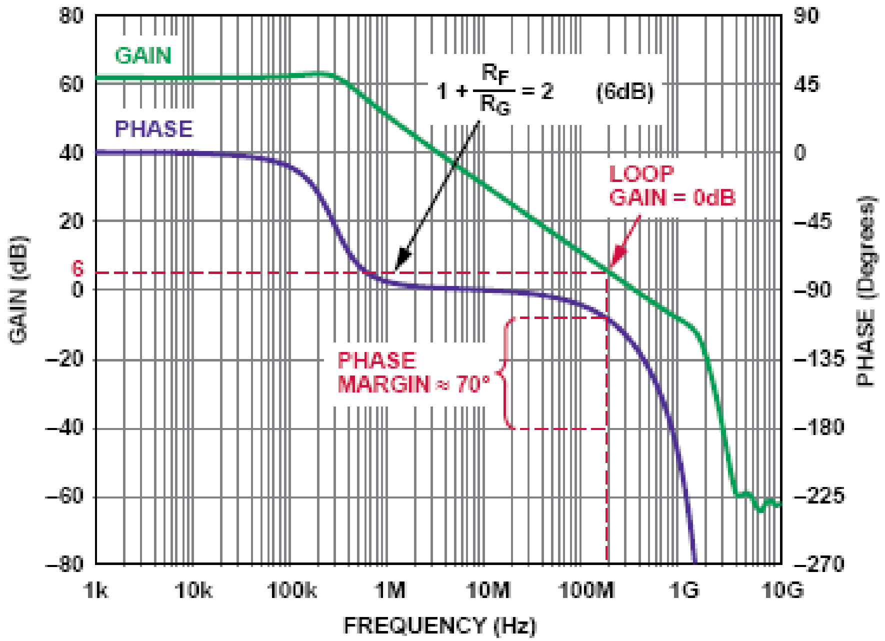
CFB yields higher FPBW and lower distortion than VFB for the same process and power dissipation

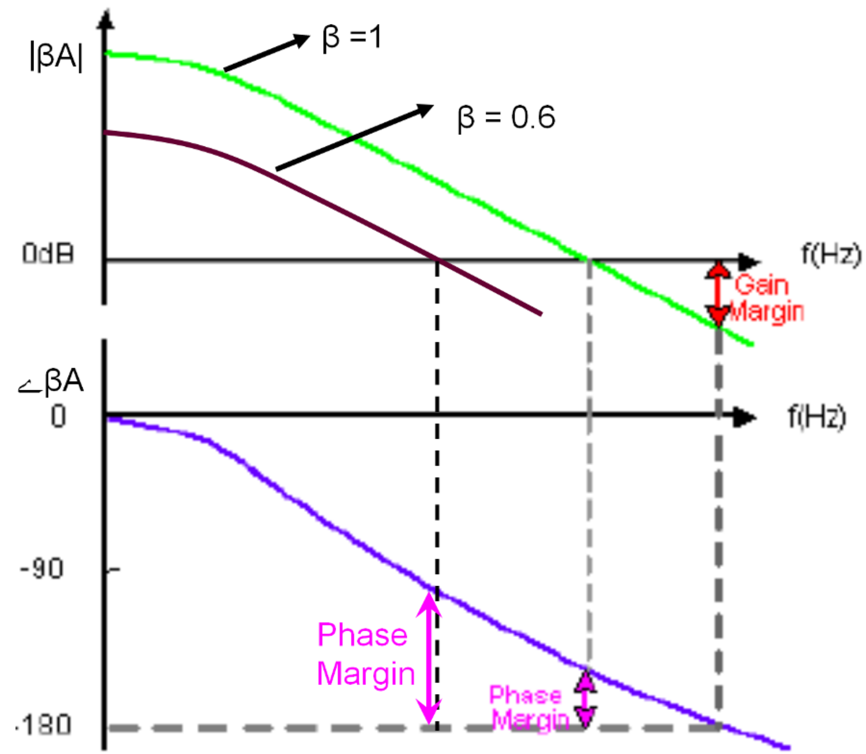
Inverting input impedance of a CFB op amp is low, non-inverting input impedance is high

Examination of this equation quickly reveals that *the closed-loop bandwidth of a CFB op amp is determined by the internal dominant pole capacitor, Cp, and the external feedback resistor R2, and is independent of the gain-setting resistor, R1*. This ability to maintain constant bandwidth independent of gain makes CFB op amps ideally suited for wideband programmable gain amplifiers.

Amplifier

Phase





Phase margin at unity gain (ϕ_m) is the difference between the amount of phase shift a signal experiences through the op amp at unity gain and 180°:

$$\phi_m = 180^\circ - \phi@B1 \qquad (11\text{–}6)$$

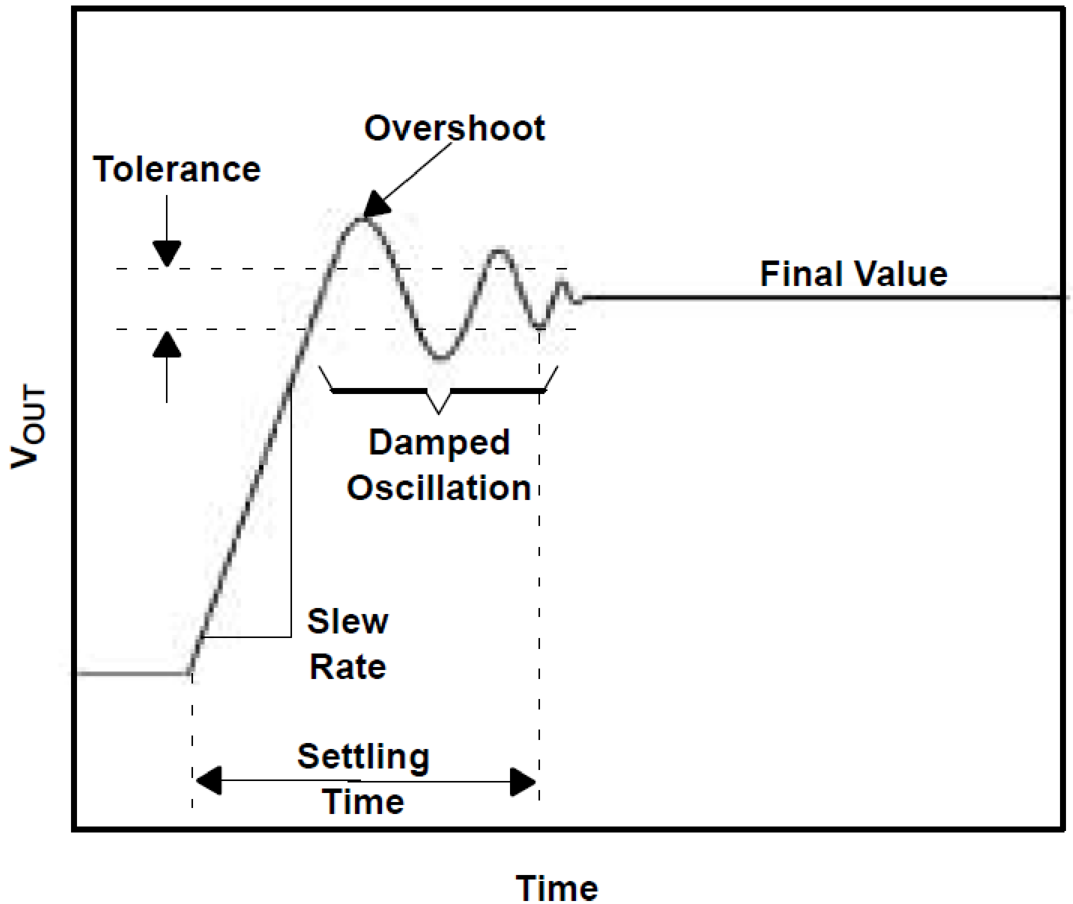
Gain margin is the difference between unity gain and the gain at 180° phase shift:

$$\text{Gain margin} = 1 - \text{Gain @}180^\circ \text{phase shift} \qquad (11\text{–}7)$$

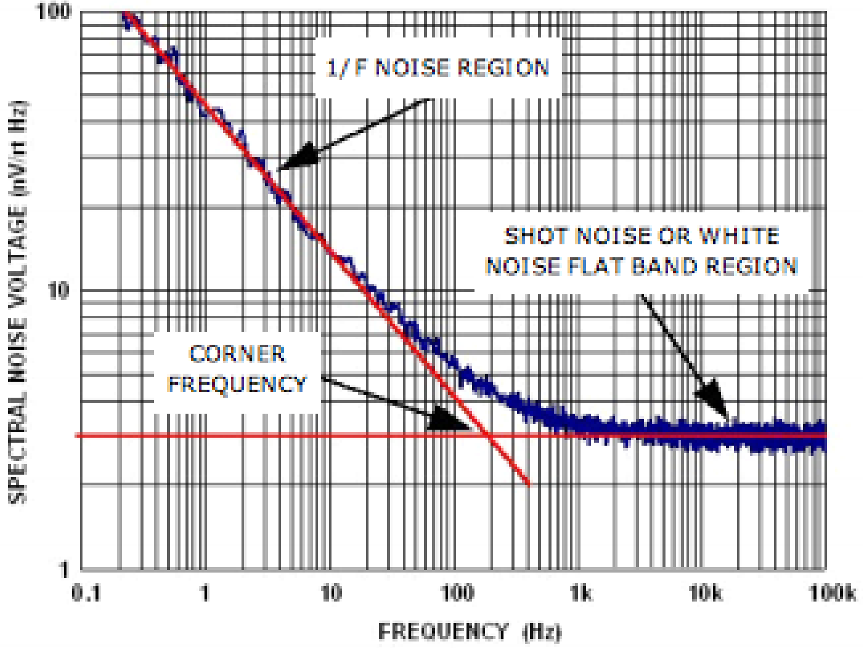
http://microchip.wikidot.com/asp0107:phase-gain-margin

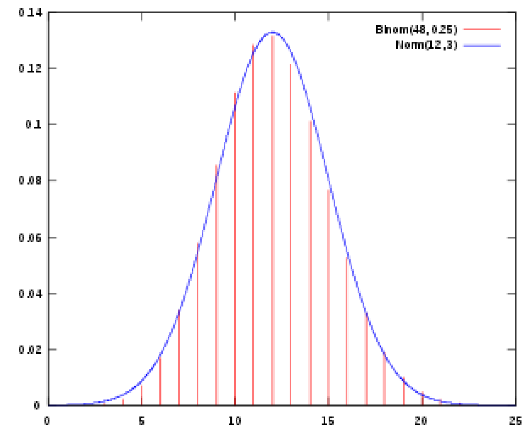
Amplifier

Settling time

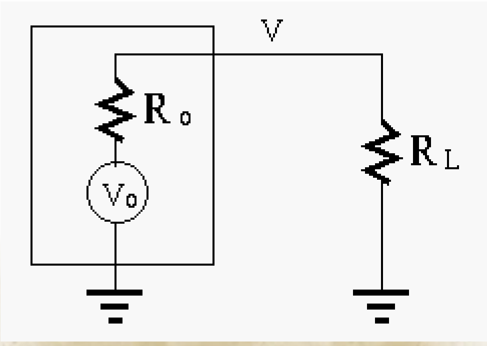


Noise

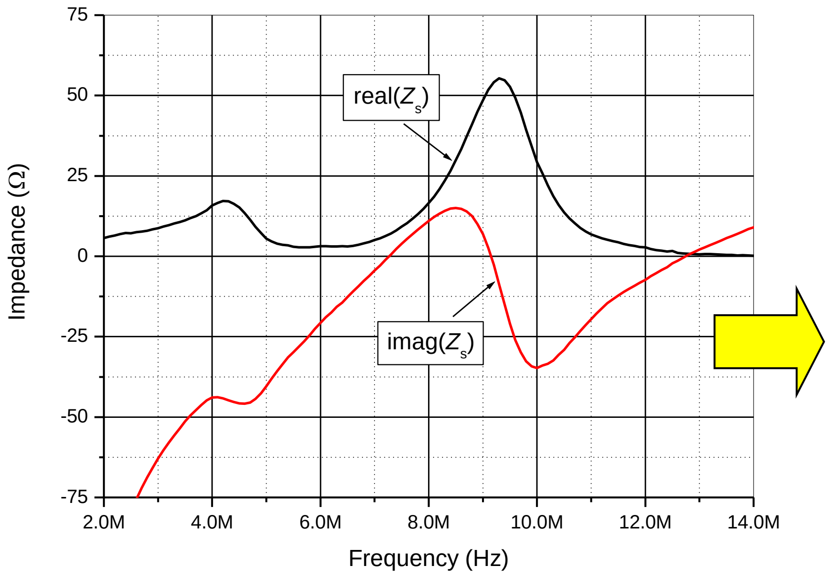


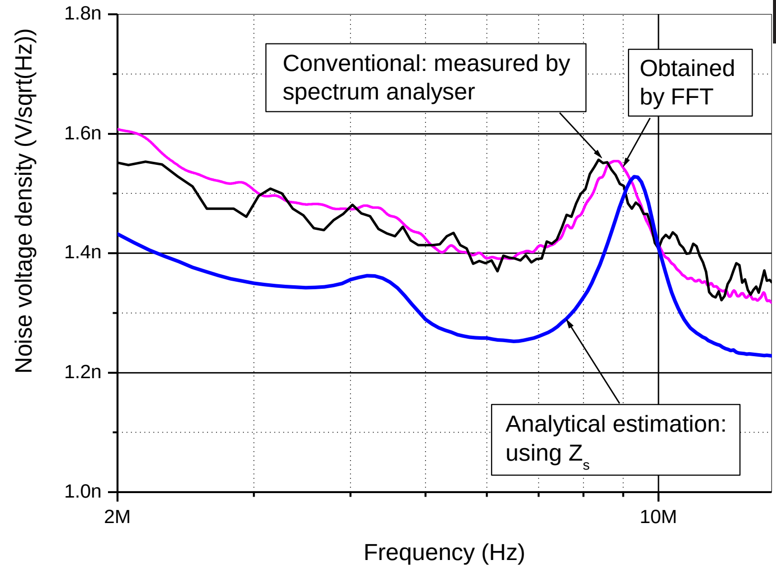


$$v_n = \sqrt{\bar{v_n^2}}\sqrt{\Delta f} = \sqrt{4k_B T R \Delta f}$$



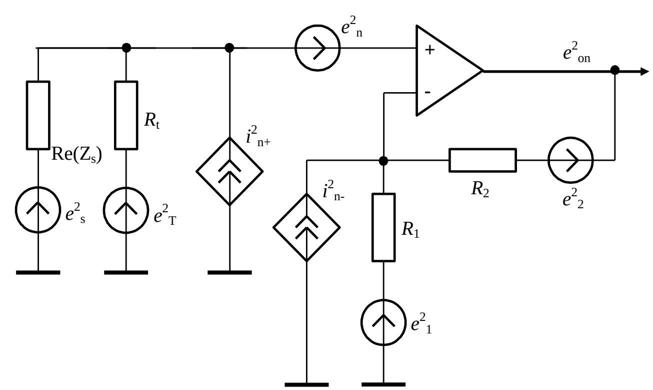
Noise





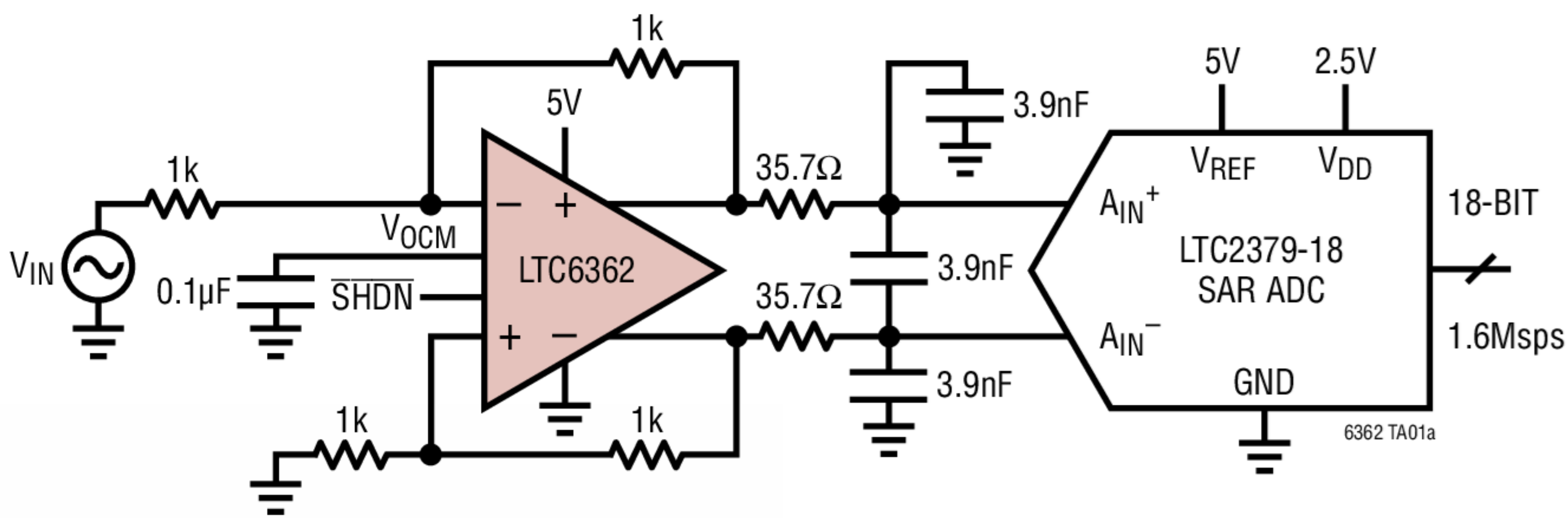
$$e_{ntotA}^2 = \left|\frac{GR_t}{R_t + Z_s}\right|^2 e_s^2 + \left|\frac{GZ_s}{R_t + Z_s}\right|^2 e_T^2 + \left|\frac{GR_t Z_s}{R_t + Z_s}\right|^2 i_{n+}^2 + G^2 e_n^2 + (G-1)^2 e_1^2 + e_2^2 + R_2^2 i_{n-}^2$$

Amplifier adds voltage and current noise sources



Amplitude estimation: ADC

DC-Coupled Interface from a Ground-Referenced Single-Ended Input to an LTC2379-18 SAR ADC



ADC sampling

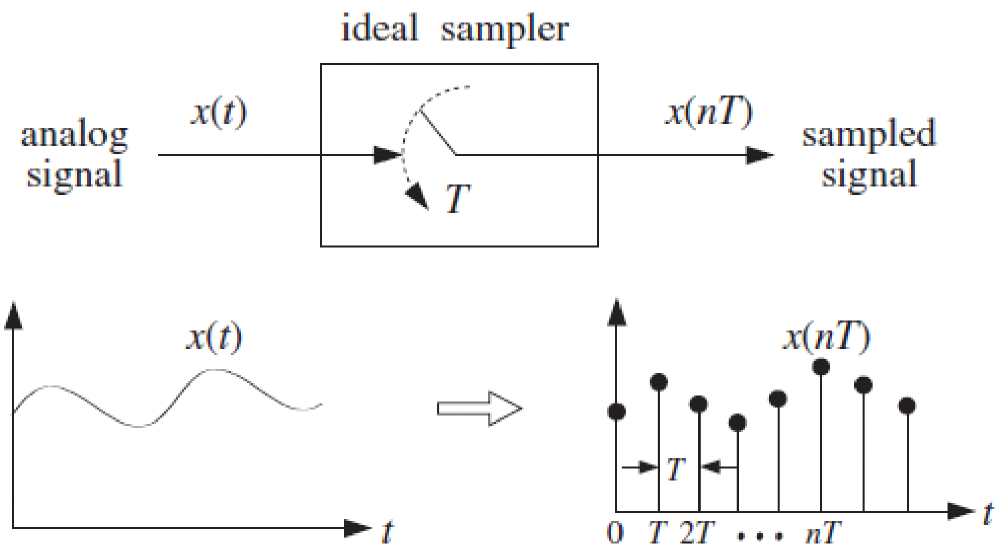
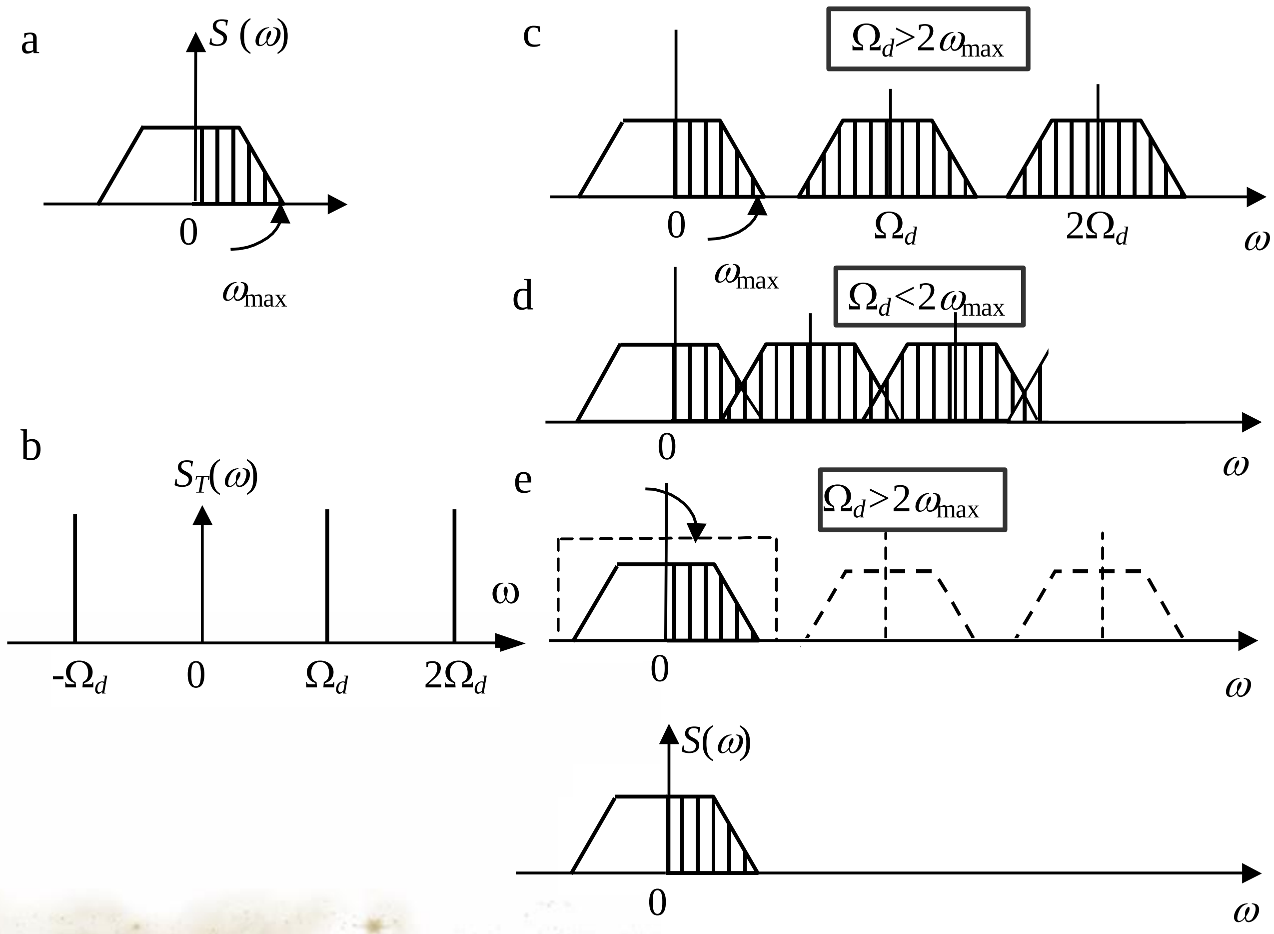
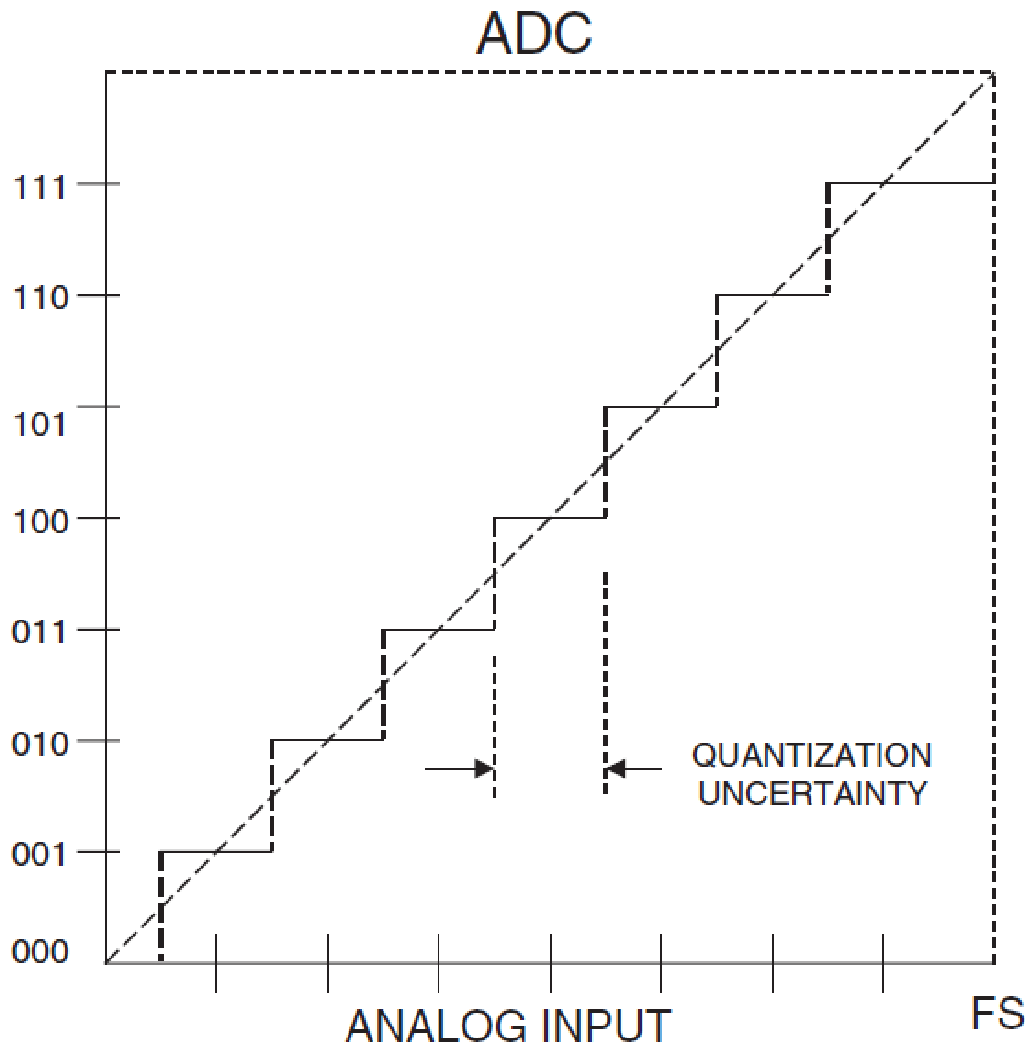


Fig. 1.3.1 Ideal sampler.

ADC aliasing

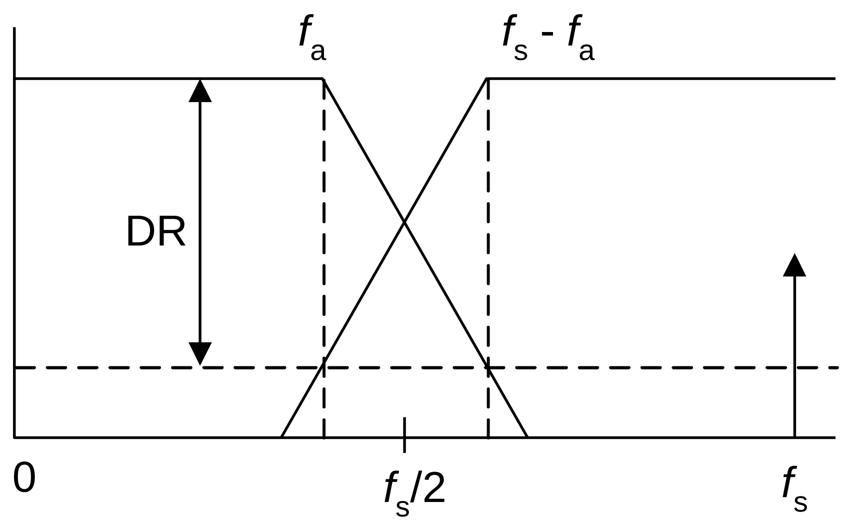


ADC quantization noise

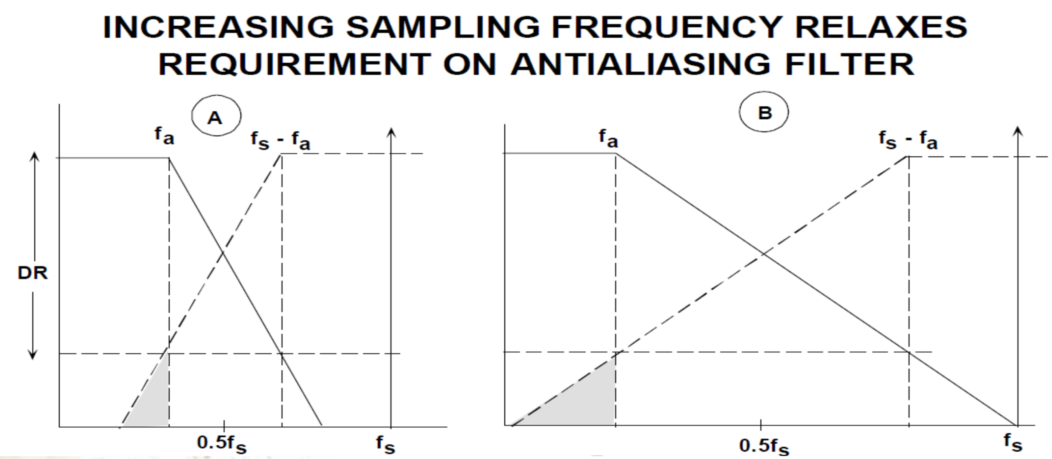


$$e_{nQ}^{\;2} = \frac{V_{ref}^{\;2}}{0.5 f_s \cdot 2^{2 \cdot \mathrm{bits}} 12}$$

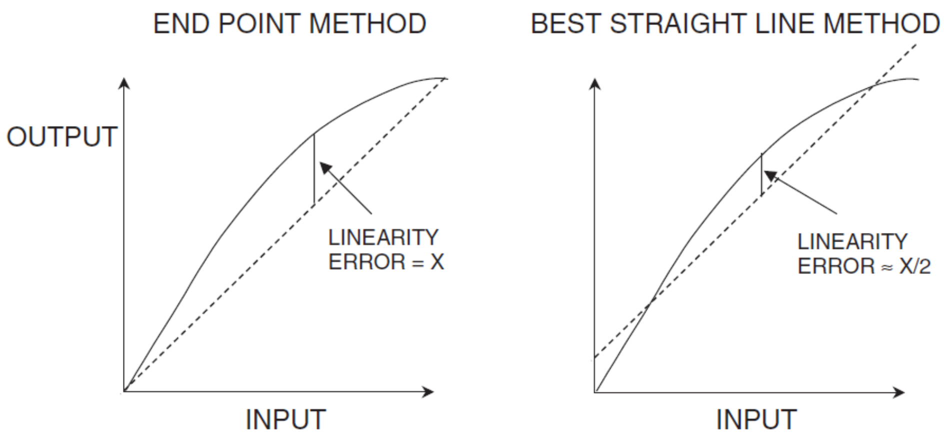
ADC antialiasing filter selection

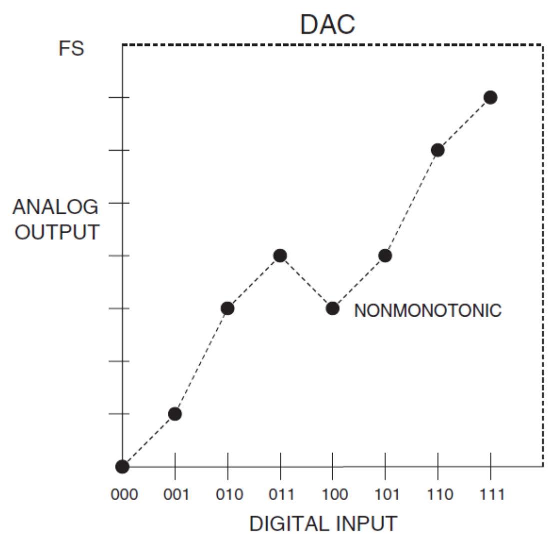


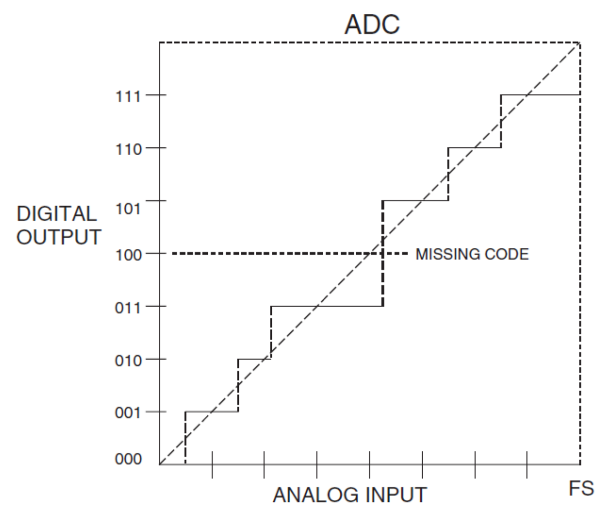
DR=6.021*N + 1.763 [dB]
DR=20lg(VmaxRMS/std(noise))



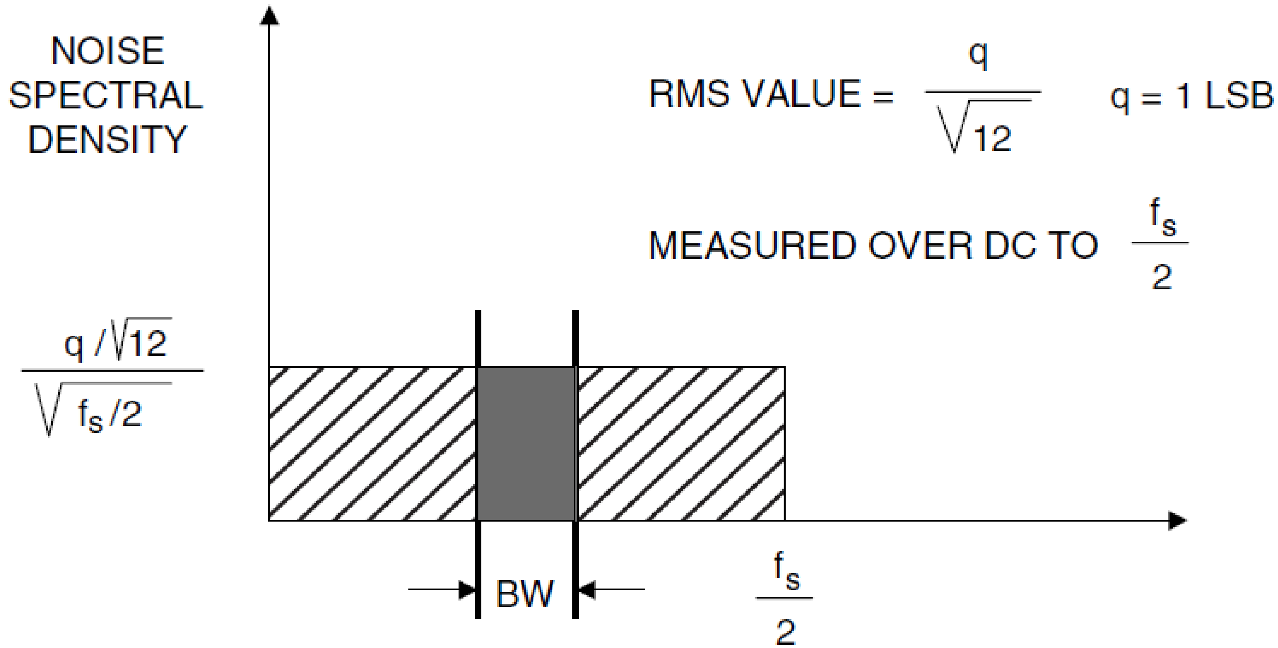
ADC nonlinearity





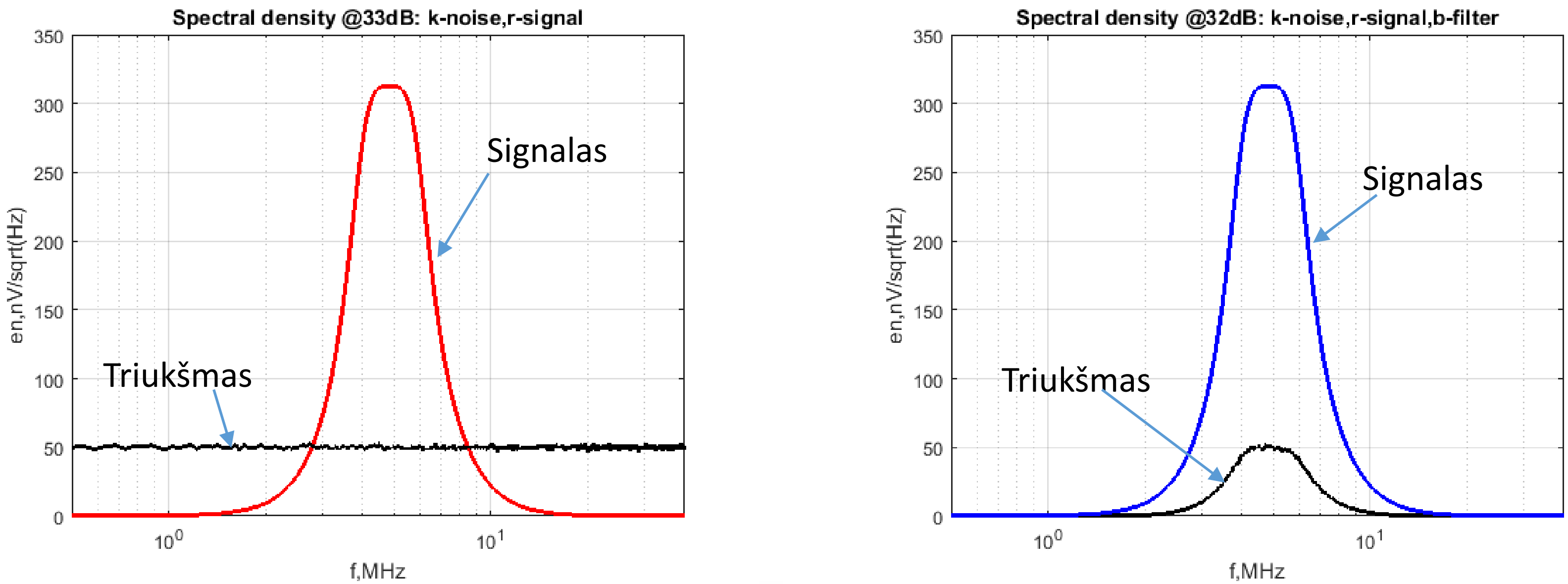


Digital filtering gain / oversampling



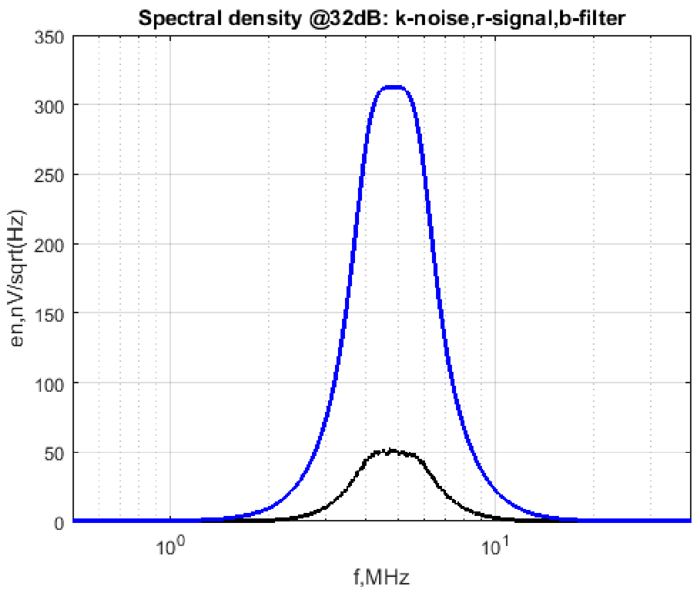
$$SNR = 6.02\ N + 1.76\ dB + \underbrace{10\log_{10}\left[\frac{f_s}{2 \times BW}\right]}_{\text{Process Gain}} \quad \text{FOR FS SINEWAVE}$$

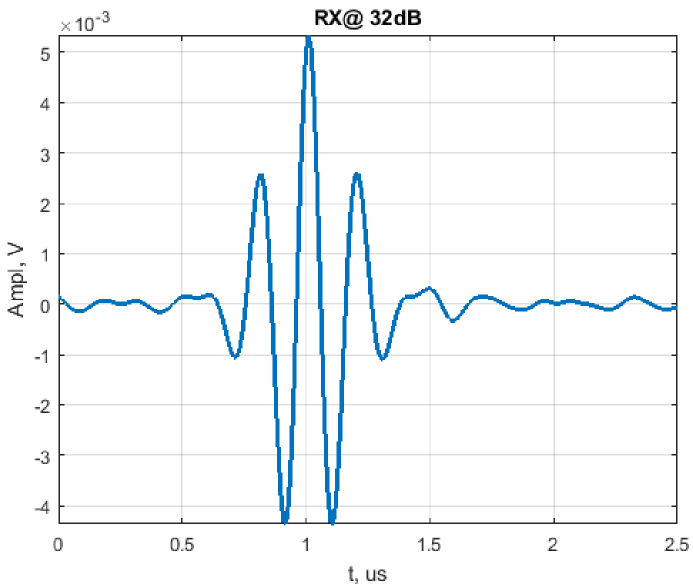
Filtering



Kuriuo iš pavaizduotų atvejų galime taikyti filtrą?

Filtering types





Matched filter

Optimal filter

BP filter

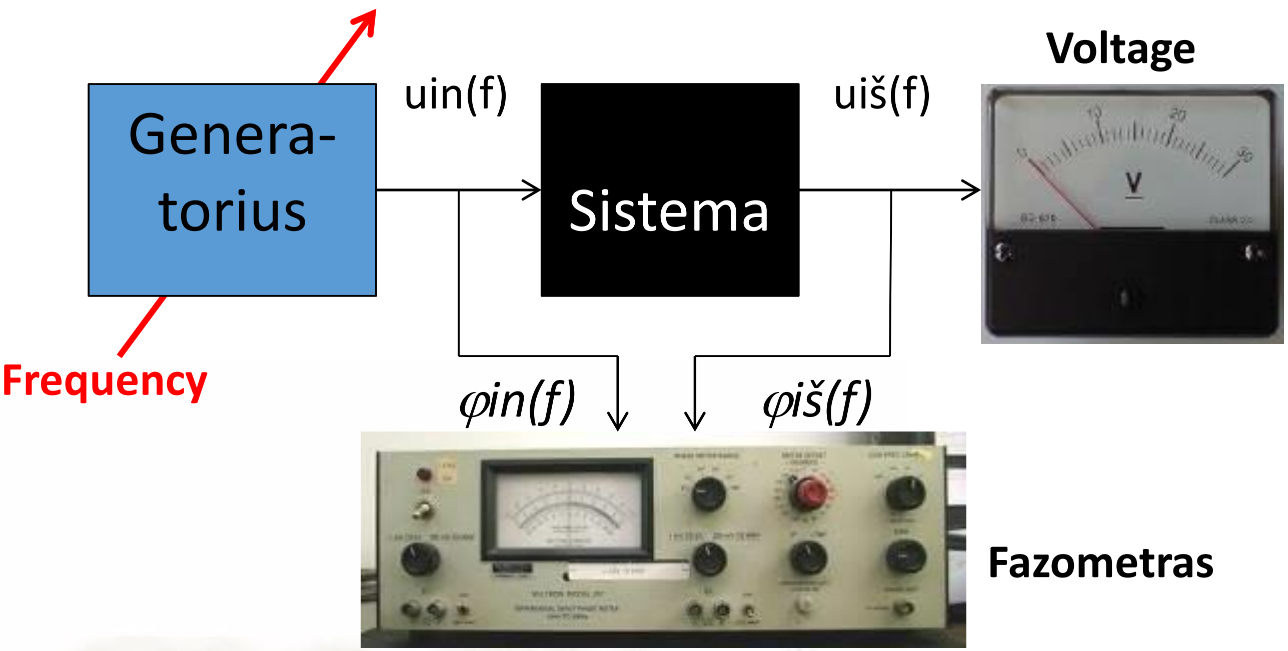
End: DC measurements

Conclusions:

- Bias and random errors exist in measurements
- Source of Bias errors – deviation of the system model from real situation
- Source of random errors – noise in electronics
- Careful consideration of the dividers, attenuators, and amplifier modes is needed in order to avoid bias errors
- Even amplifier biasing requires special attention, as well the gain bandwidth evaluation and settling time
- Use of ADC in measurements could add additional sources of errors
- Required bandwidth and dynamic range define the ADC antialiasing filter requirements which can be relaxed by the increase of the sampling frequency
- Filtering can reduce random errors, one of the simplest examples - oversampling

AC response measurement

Probing signal – variable frequency sinusoid



Magnitude of transmission (gain) AC response

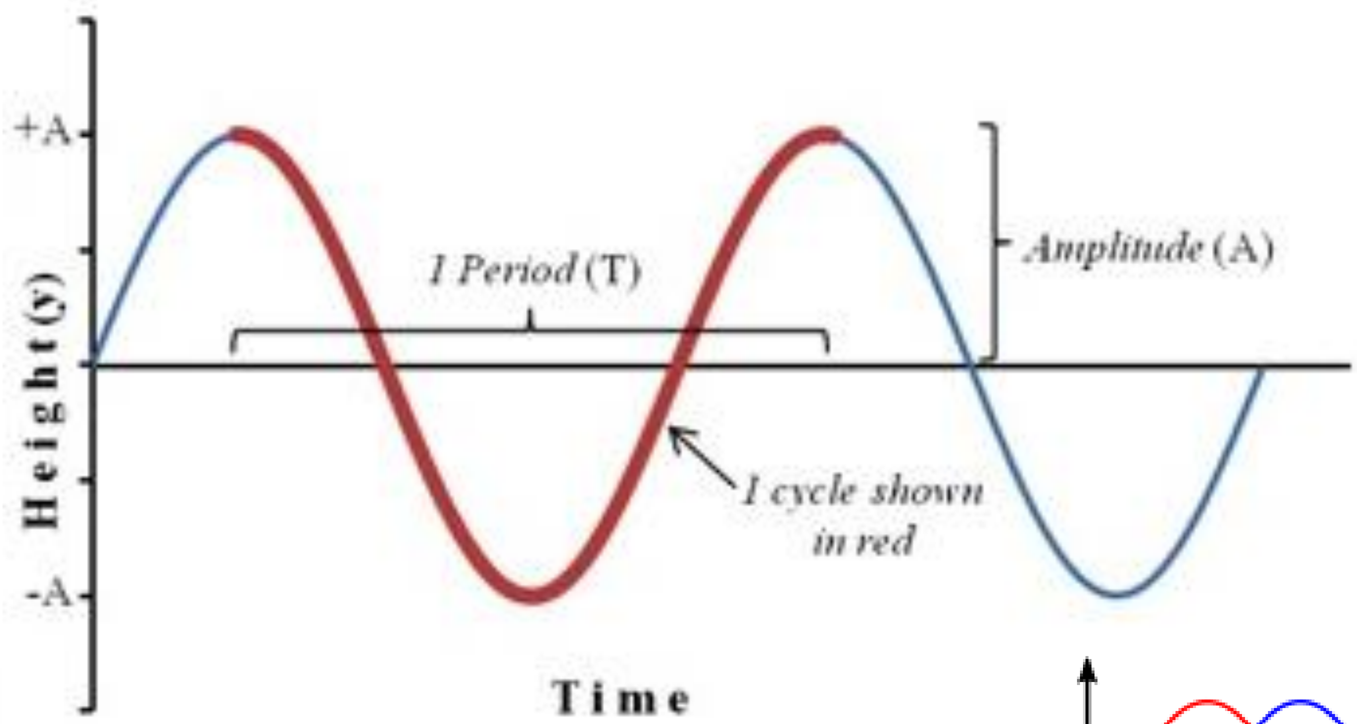
$$\boxed{H(f) = \frac{u_{iš}(f)}{u_{in}(f)}}$$

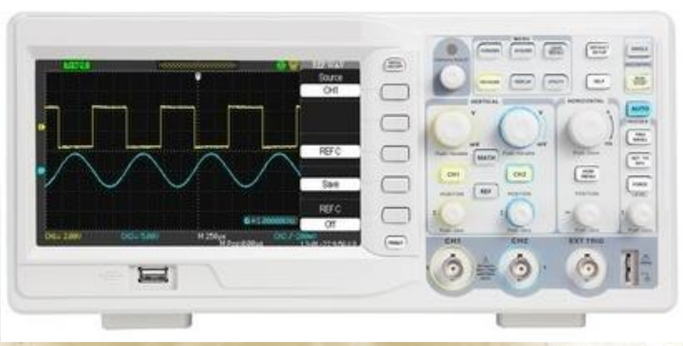
Phase AC response

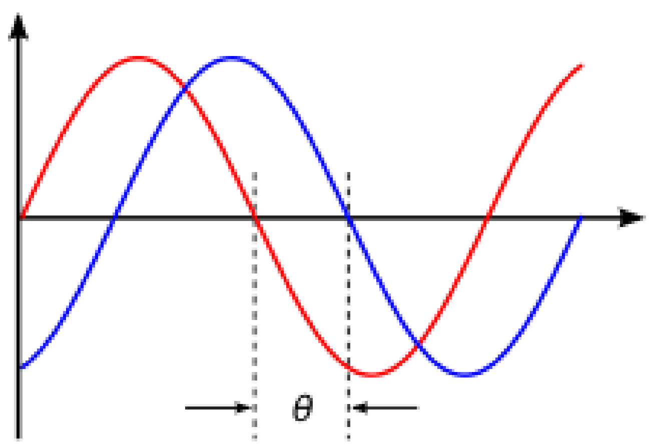
$$\boxed{\begin{aligned} \varphi(f) &= \varphi_{in}(f) - \varphi_{iš}(f) = \\ &= -\varphi_{iš}(f) \end{aligned}}$$

$\varphi_{in}(f) = 0$, then, but can we keep that?

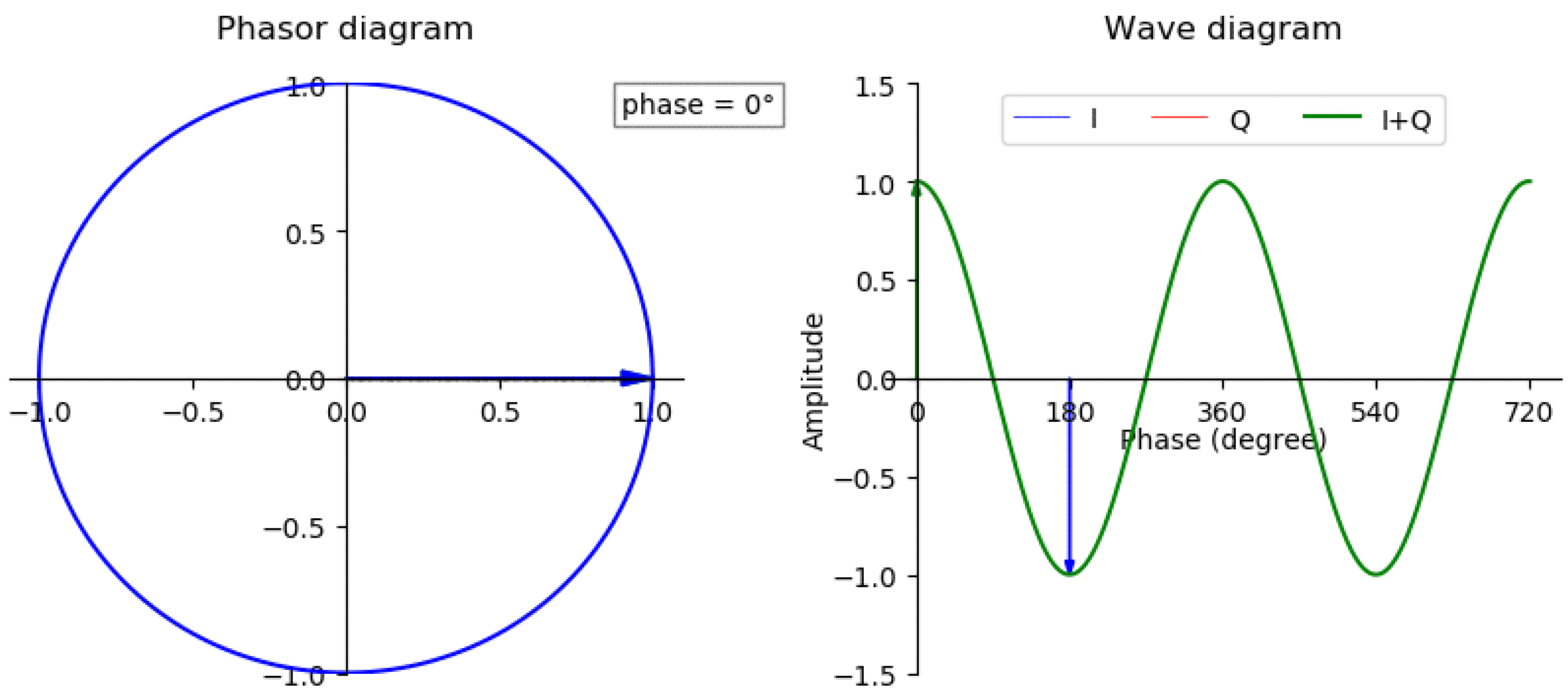
Sinusoid


+A
-A
Height (y)
1 Period (T)
Amplitude (A)
1 cycle shown in red
Time




θ

Sinusoid

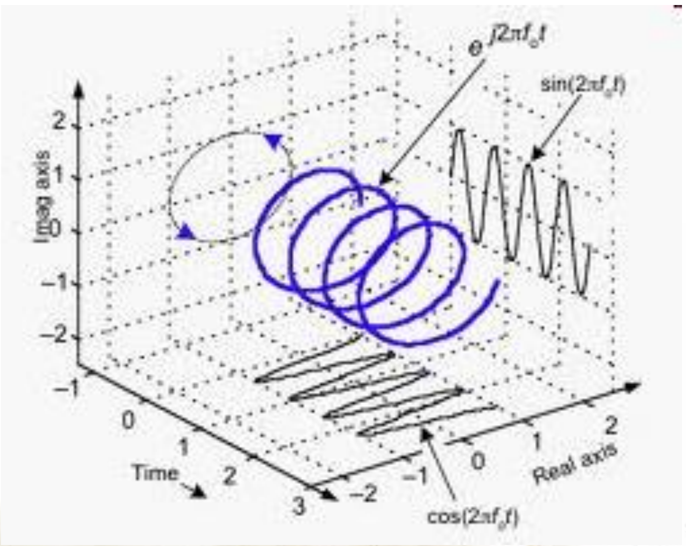


Phasor diagram
phase = 0°
1.0
0.5
0.0
−0.5
−1.0
−1.0
−0.5
0.0
0.5
1.0
Wave diagram
1.5
1.0
0.5
0.0
−0.5
−1.0
−1.5
Amplitude
I
Q
I+Q
0
180
360
540
720
Phase (degree)

Sinusoid

- Complex exponent is periodical and has real and imaginary parts:

$$x[n] = e^{j\omega n} = \cos(\omega n) + j\sin(\omega n)$$

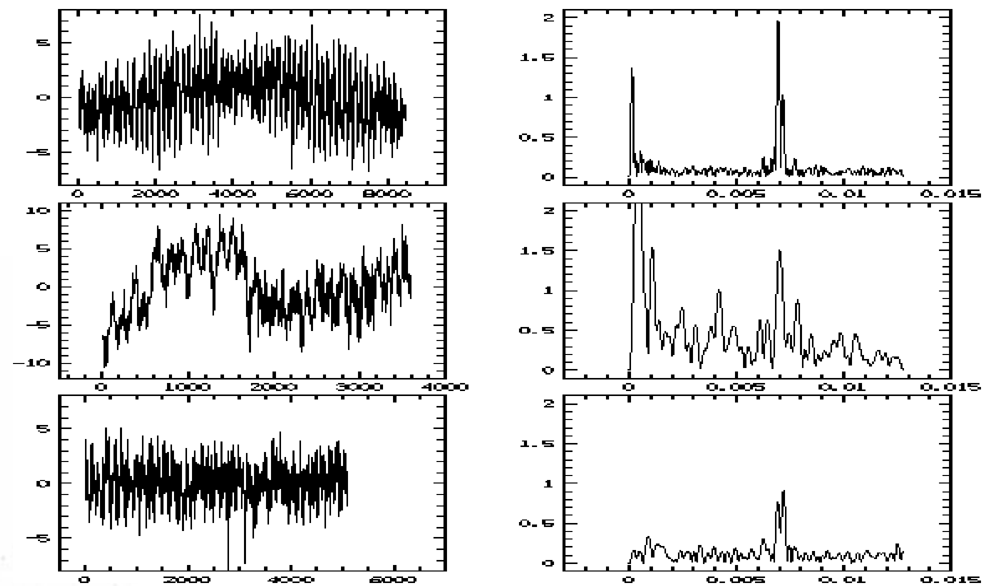


FT

- Continuous time and frequency signal has ALL the values presented (sampled). Such signal is not periodic both in time and frequency domain
- Fourier transform (**FT**) is used to describe continuous, non-periodic (single) signals

$$S(\omega) = \int_{-\infty}^{+\infty} s(t) e^{-j\omega t} dt$$

$$s(t) = \frac{1}{2\pi} \int_{-\infty}^{+\infty} S(\omega) e^{j\omega t} d\omega$$



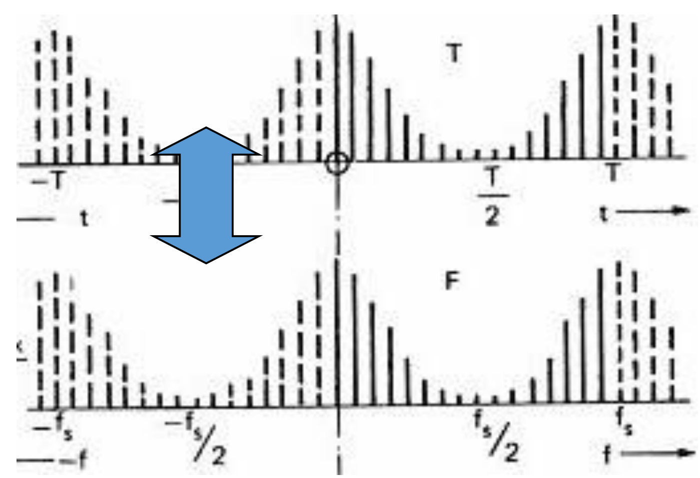
DFT

Discrete time AND frequency signal is presented only at discrete time AND frequency samples. Such signal is periodic both in time AND frequency domain. Briefly: such signal is both discrete and periodic;

DFT-Discrete Fourier transform is used to transforms the siagnal between time and frequency axes. **FFT** is the **separate DFT case** where signal length N is log2 (4,16,32,64,128,256,512,1024 etc.)

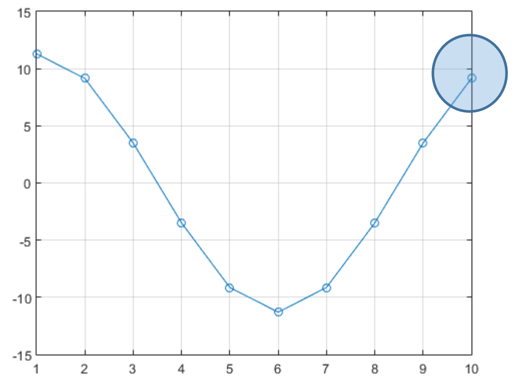
$$S_k = \sum_{n=0}^{N-1} s_n e^{-j2\pi nk/N} = S_{k+N}$$

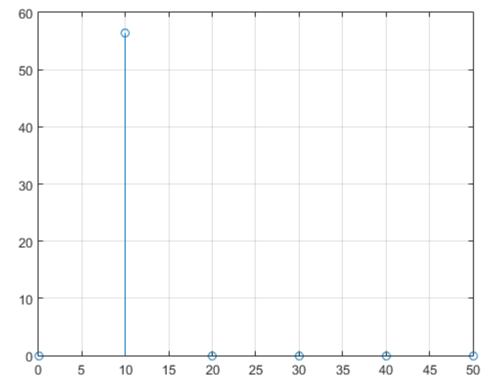
$$s_n = \frac{1}{N}\sum_{k=0}^{N-1} S_k e^{j2\pi nk/N} = s_{n-N}$$



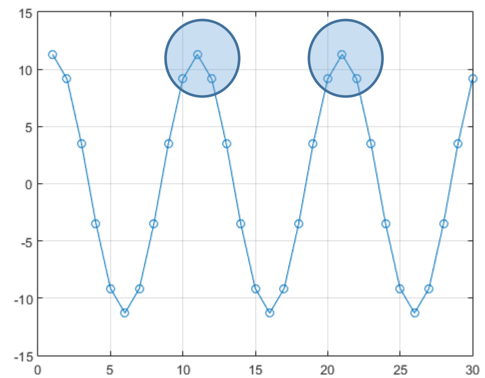
Sinusoid parameters estimation DFT

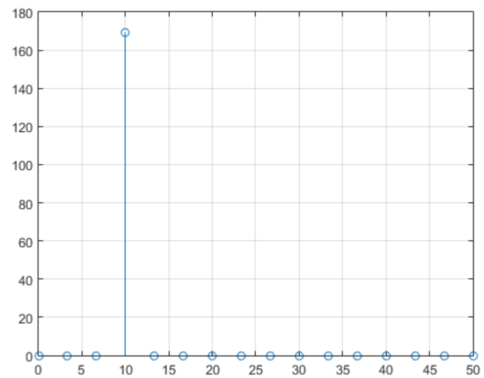
DFT is assuming signal as periodical in time domain
Everything is OK while record length matches period:

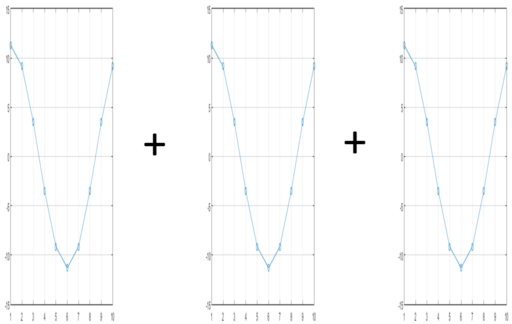




F_s=100MHz, f_0=10MHz, N=10

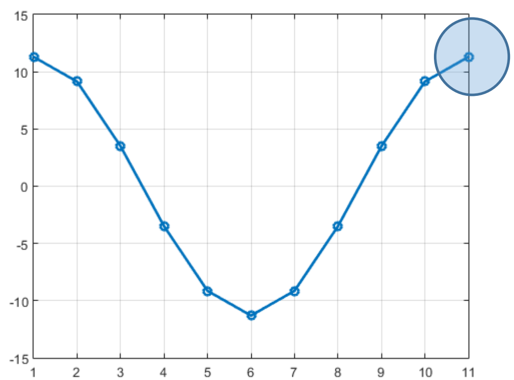


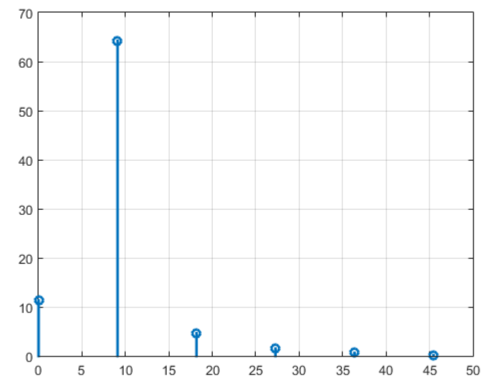




Sinusoid parameters estimation DFT

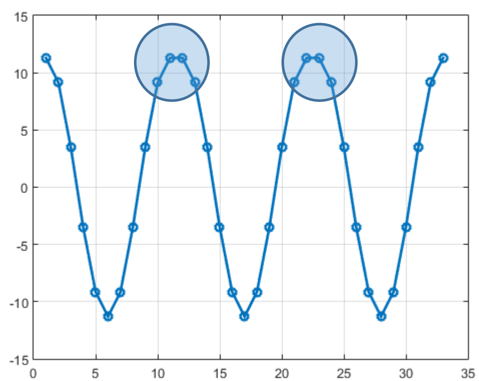
Tačiau vos tik įrašo ilgis nebeatitinka lyginio periodų skaičiaus,
Spektras keičiasi

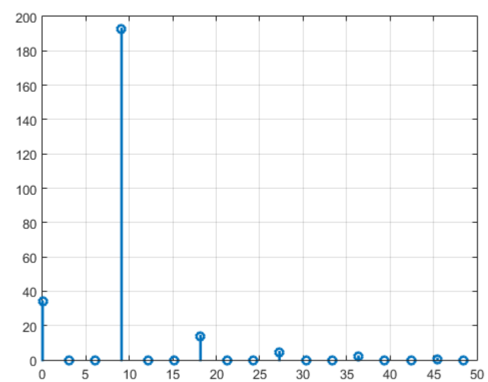


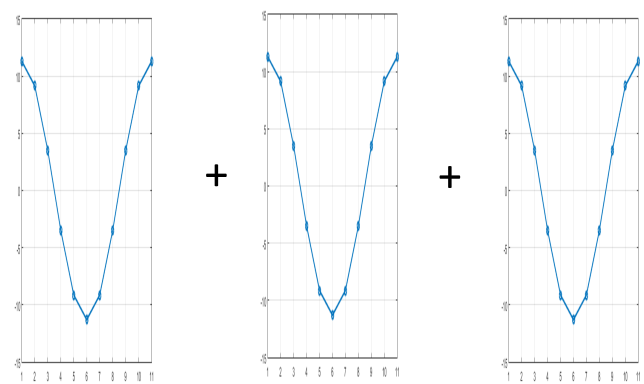


F_s=100MHz, f_0=10MHz, N=11

Kodėl?

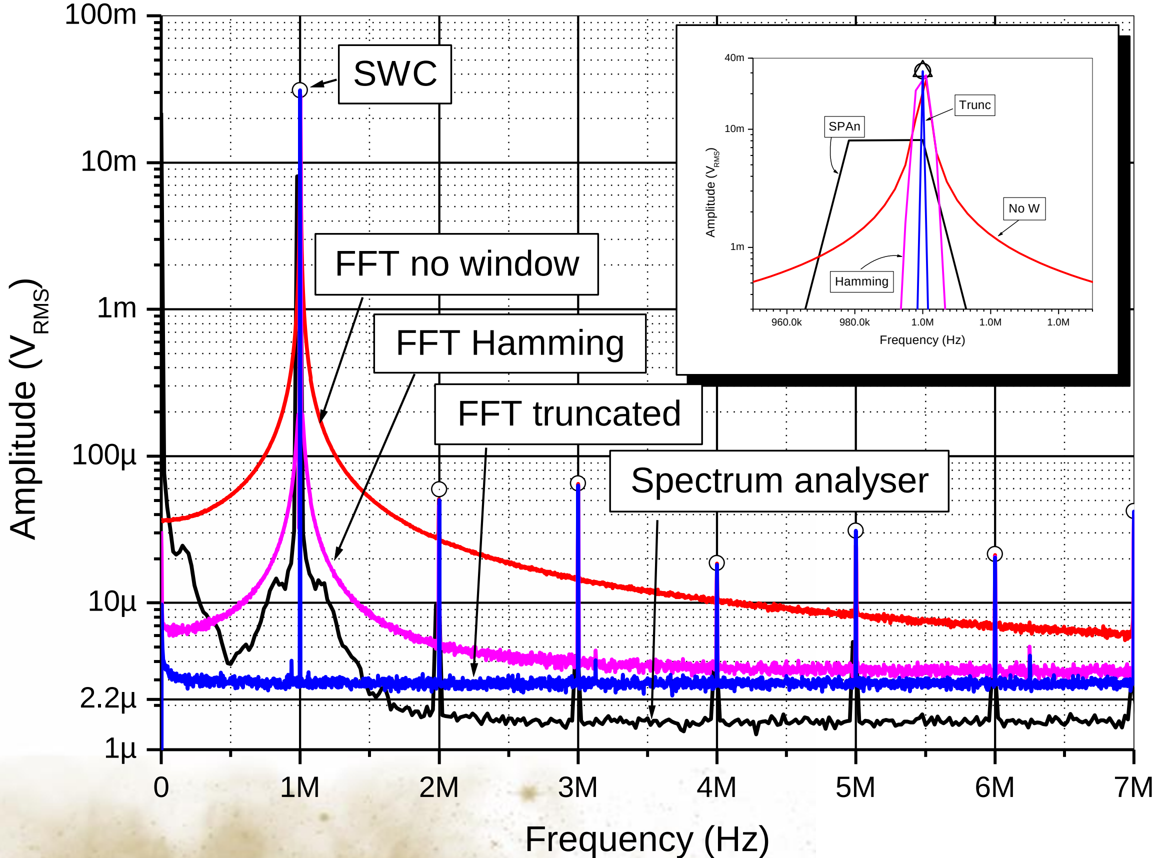






Sinusoid parameters estimation DFT

Problem: not all frequencies’ magnitude can be accurately estimated using DFT/FFT



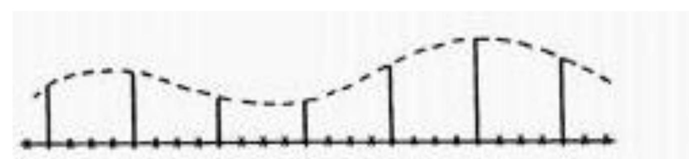
Solution-DTFT

Discrete time signal is described (sampled) at discrete time instances. Spectrum of such signal is periodic

Discrete time Fourie transform **DTFT** is used to analyze sampled single/non-periodic signals

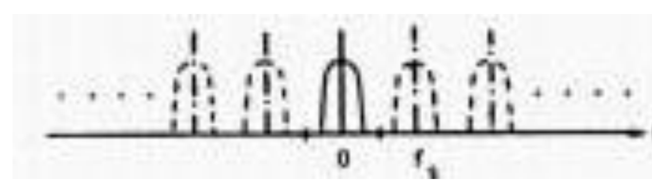
Frequency here is Continuous variable

$$S(\Omega)=\sum_{n=-\infty}^{+\infty} s_n e^{-j\Omega n}=S(\Omega+2\pi)$$



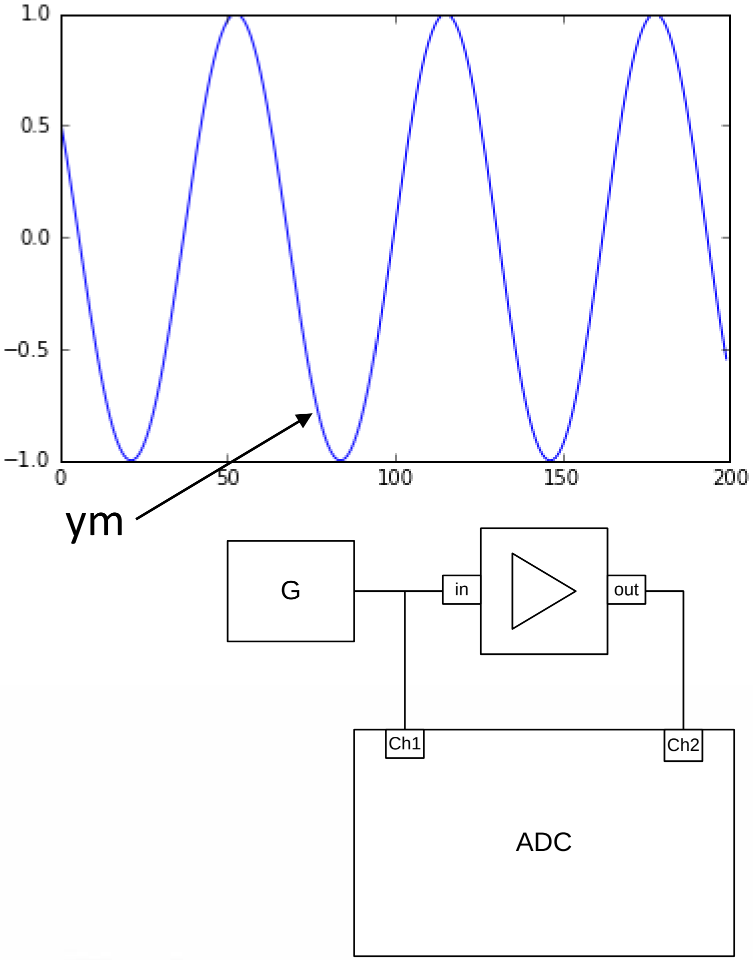
Time is discrete

$$s_n=\frac{1}{2\pi}\int_0^{2\pi} S(\Omega)e^{j\Omega n}\,d\Omega$$



Sinusoid parameters estimation SWC

ktu 1922



SWC=single frequency DTFT

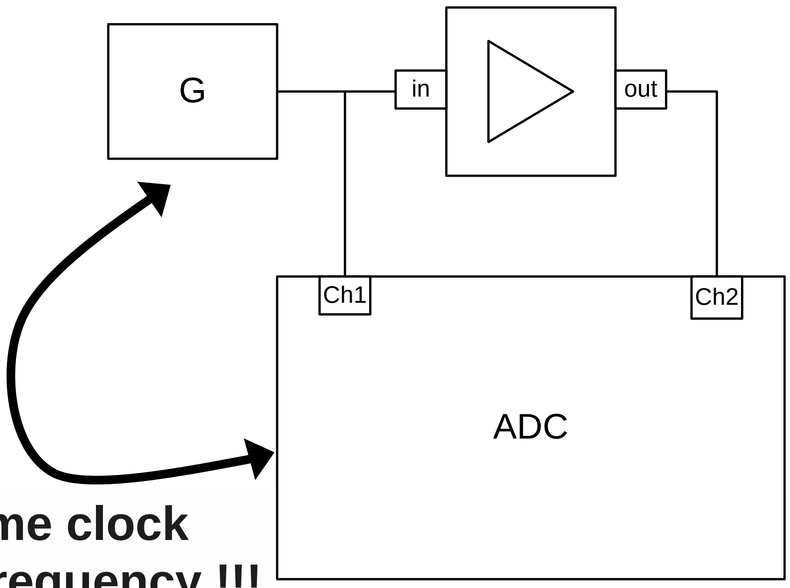
$$u(t) = U_c \cos(2\pi f t) + U_s \sin(2\pi f t) + U_{DC}$$

$$U_c = \frac{\sum_{m=1}^{M} \left[\cos(2\pi f t_m) \cdot y_m\right]}{\sum_{m=1}^{M} \left[\cos(2\pi f t_m)\right]^2}$$

$$U_s = \frac{\sum_{m=1}^{M} \left[\sin(2\pi f t_m) \cdot y_m\right]}{\sum_{m=1}^{M} \left[\sin(2\pi f t_m)\right]^2}$$

https://se.mathworks.com/matlabcentral/fileexchange/68174-swctruncated

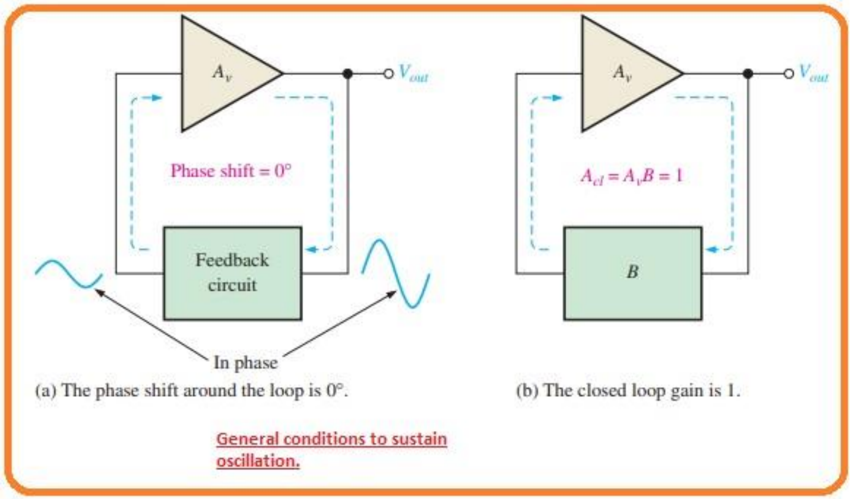
Sinusoid parameters estimation SWC

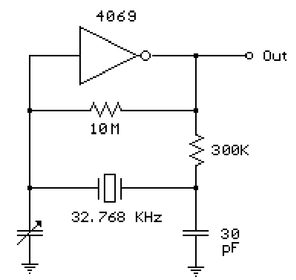


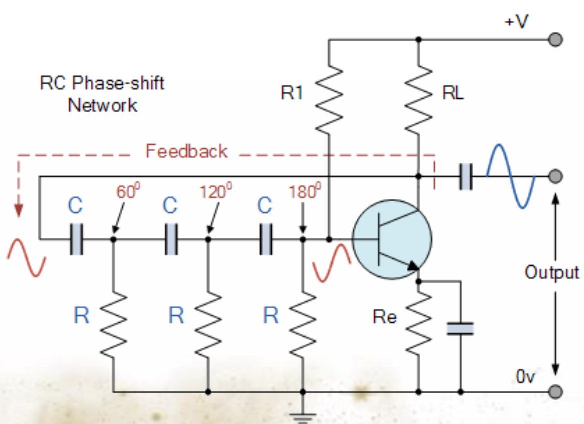
Same clock frequency !!!

https://se.mathworks.com/matlabcentral/fileexchange/68174-swctruncated

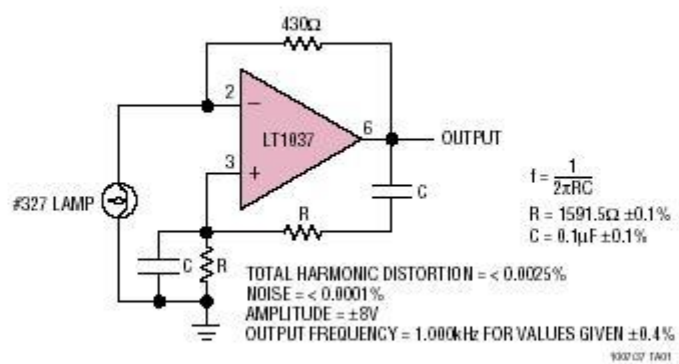
Sinusoid production





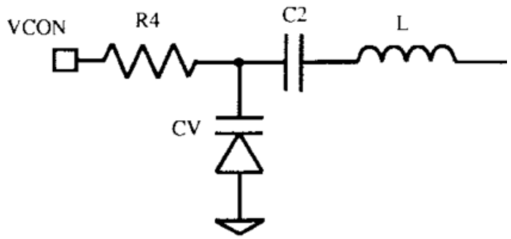


Ultrapure 1kHz Sine Wave Generator

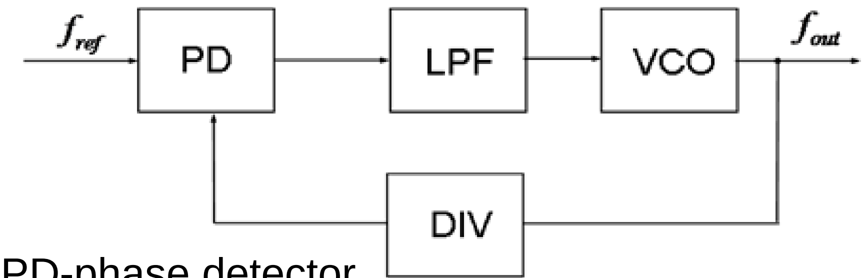


Sinusoid production

VCO-voltage controlled oscillator

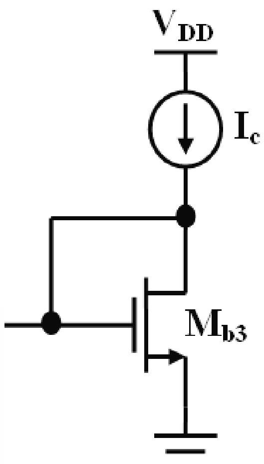


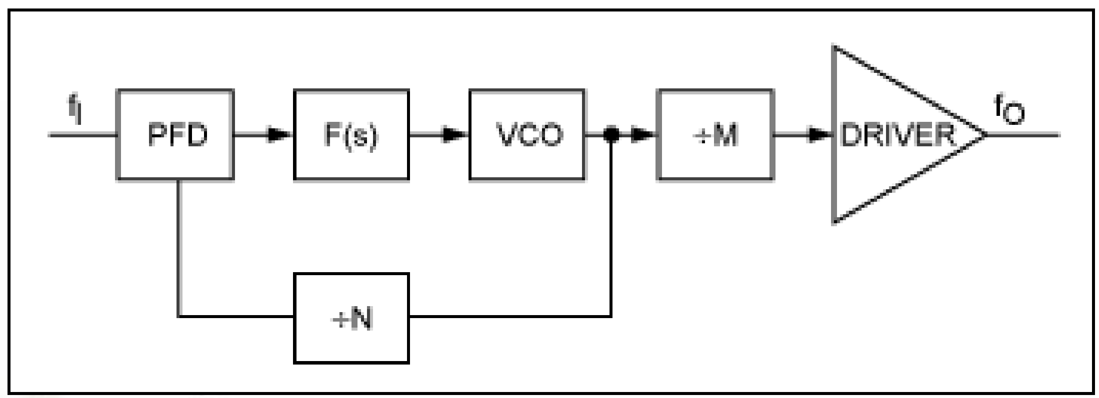
PLL-phase-locked loop



PD-phase detector

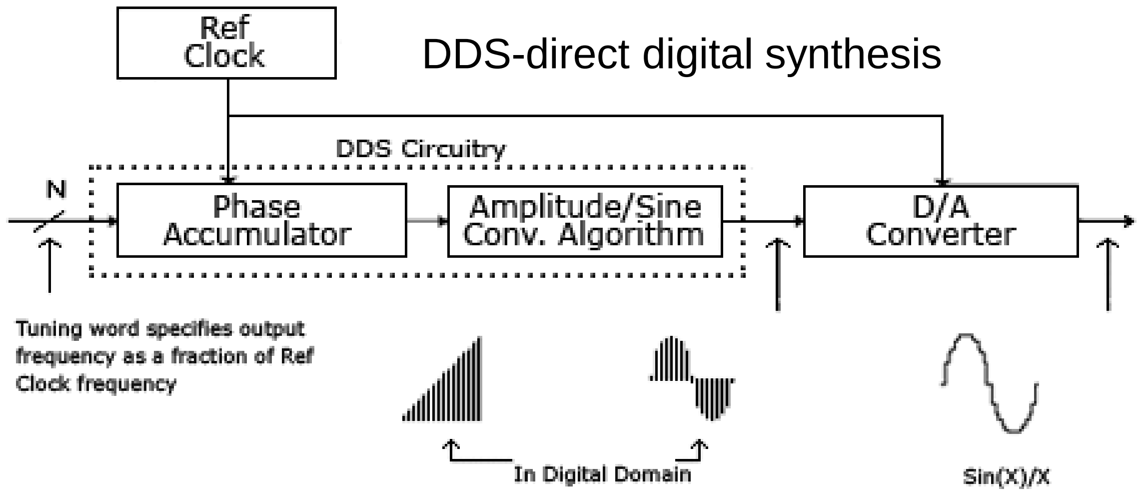
CCO-current controlled oscillator

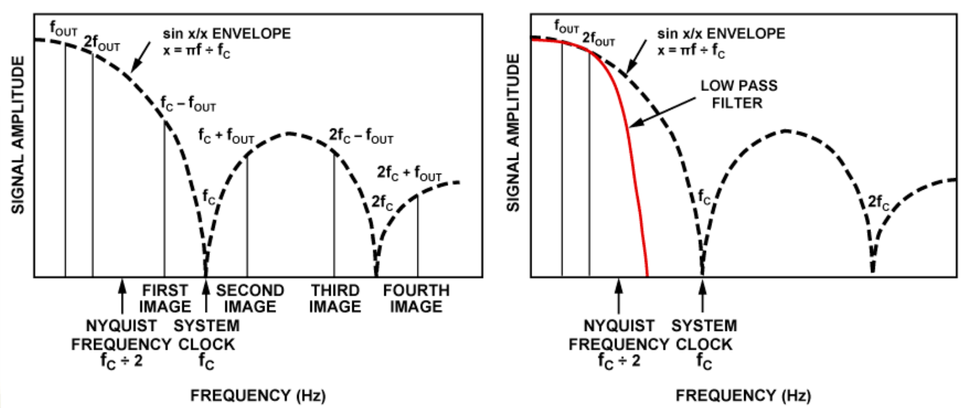




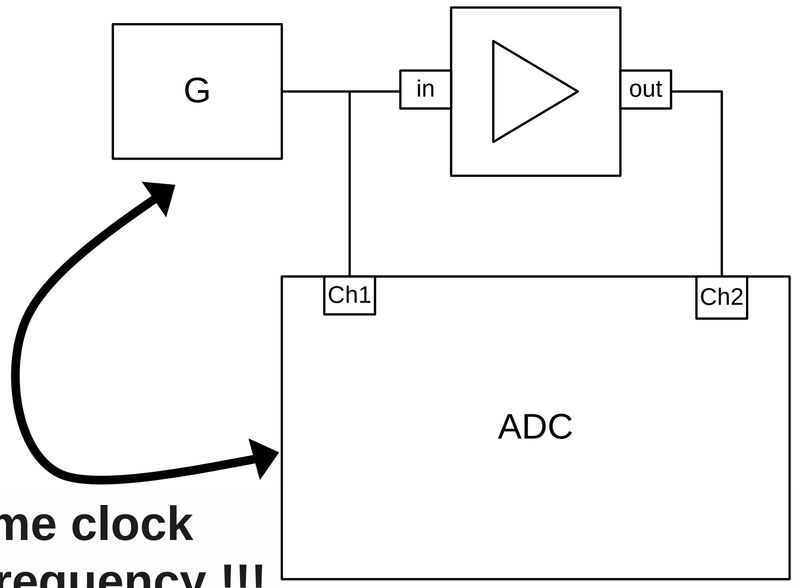
Sinusoid production

DDS-direct digital synthesis

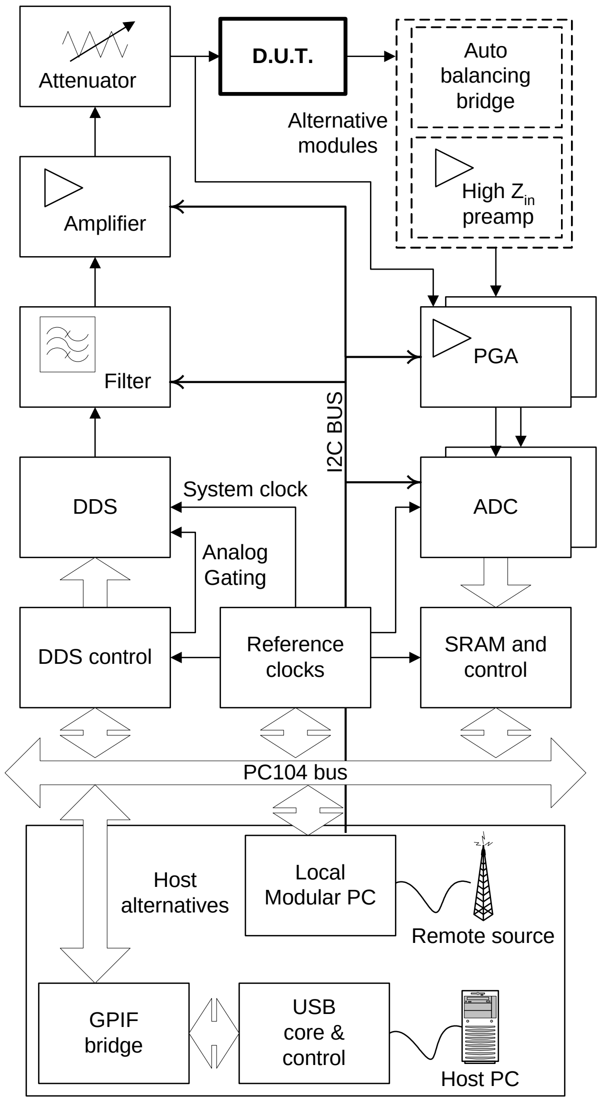




Transmission AC response: SWC



Same clock frequency !!!



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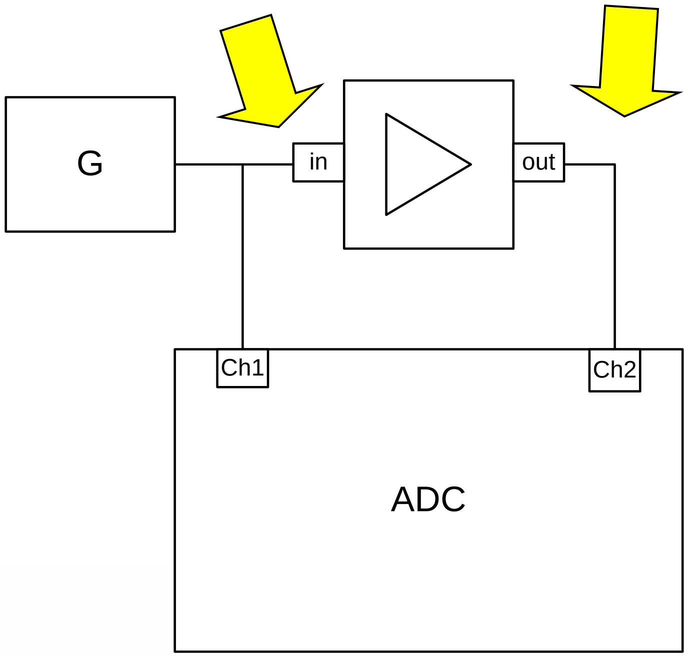
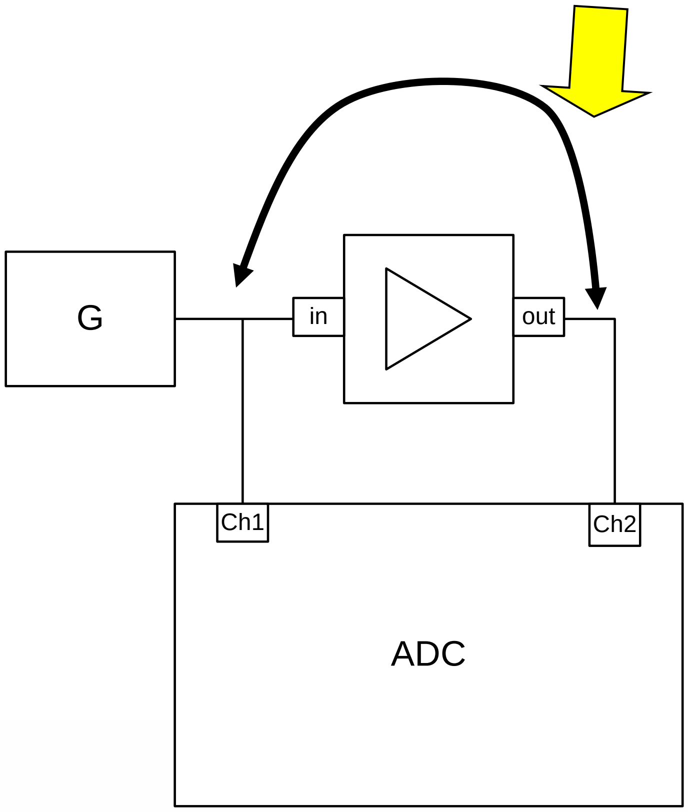
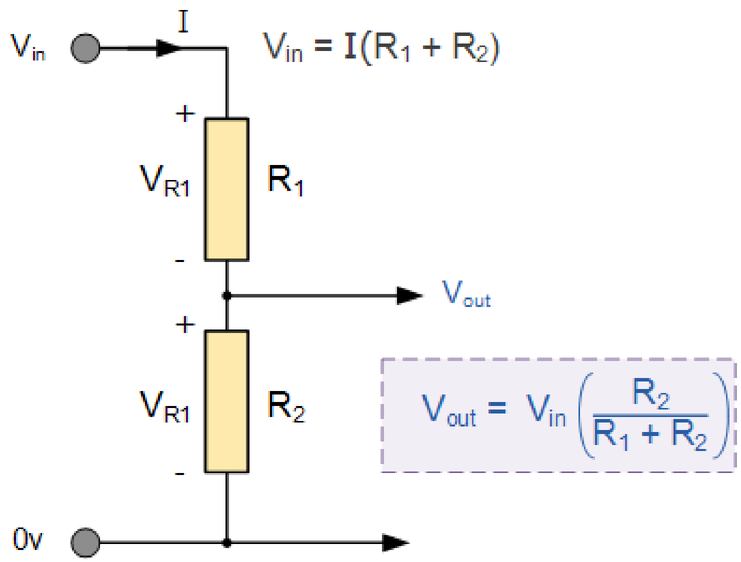
Connections

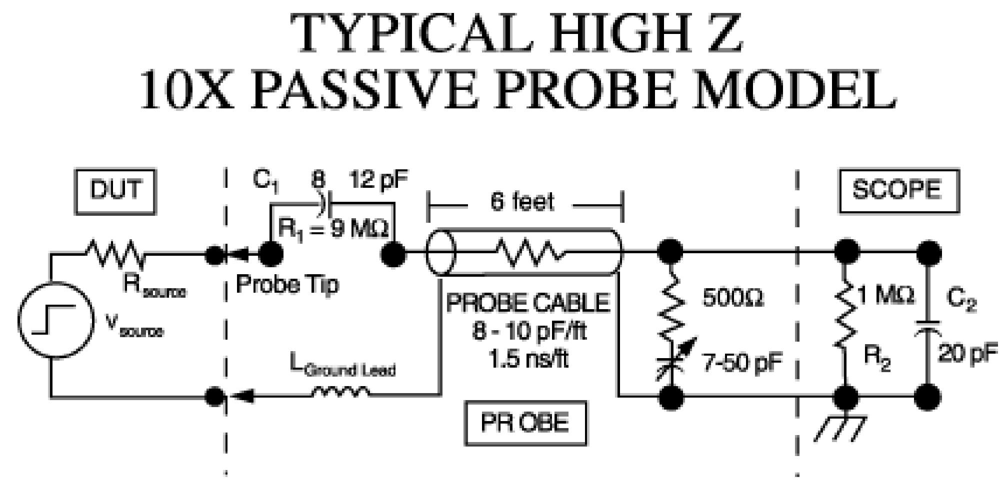


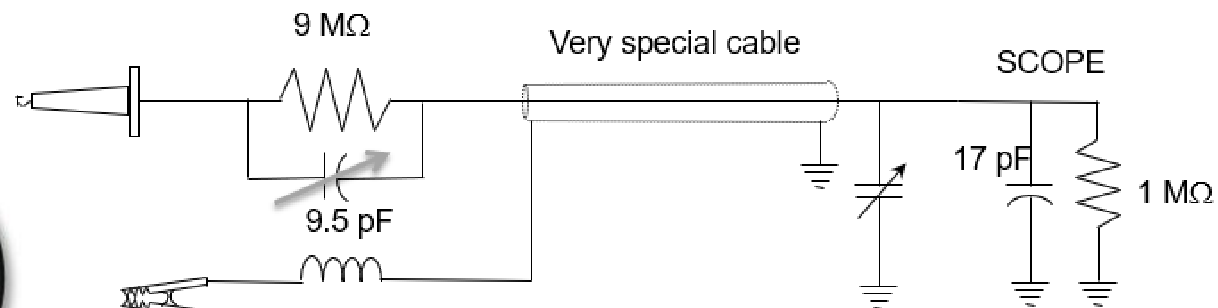
Calibration

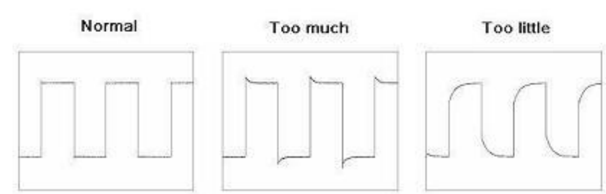


Connections

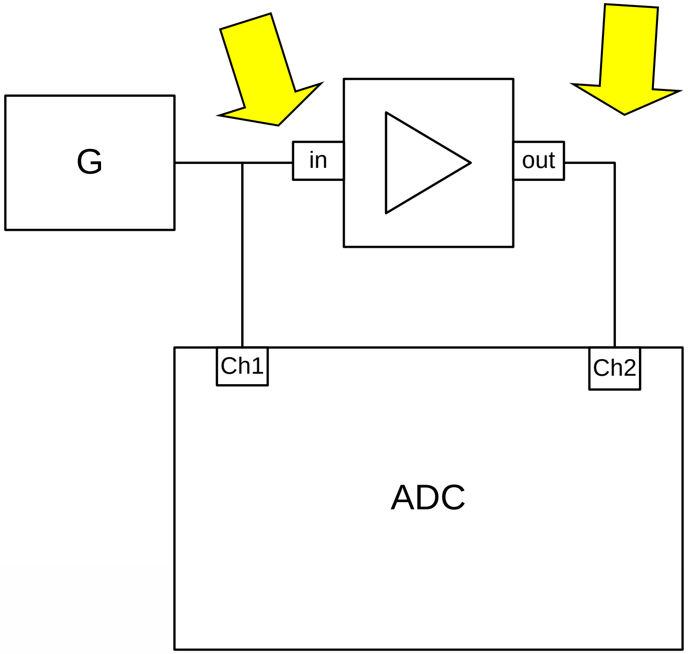






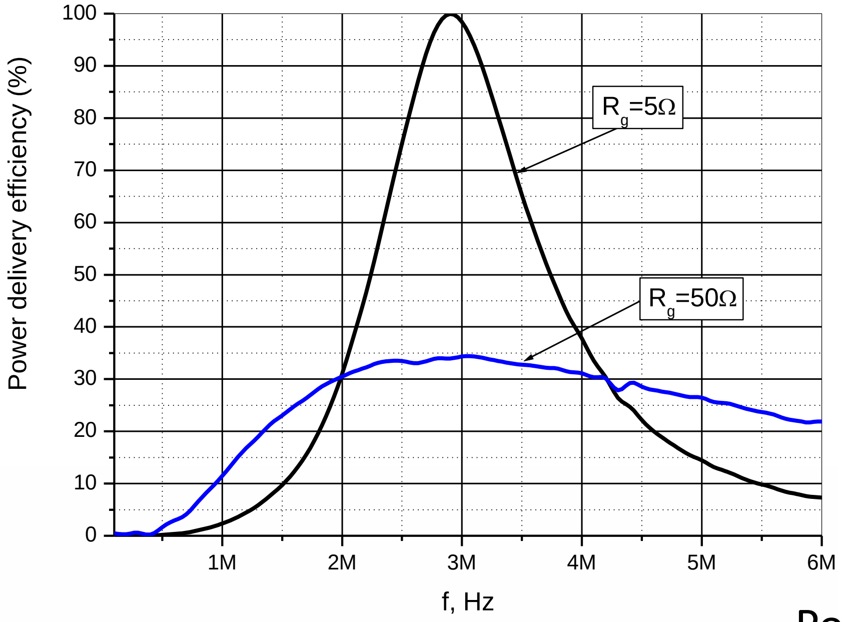


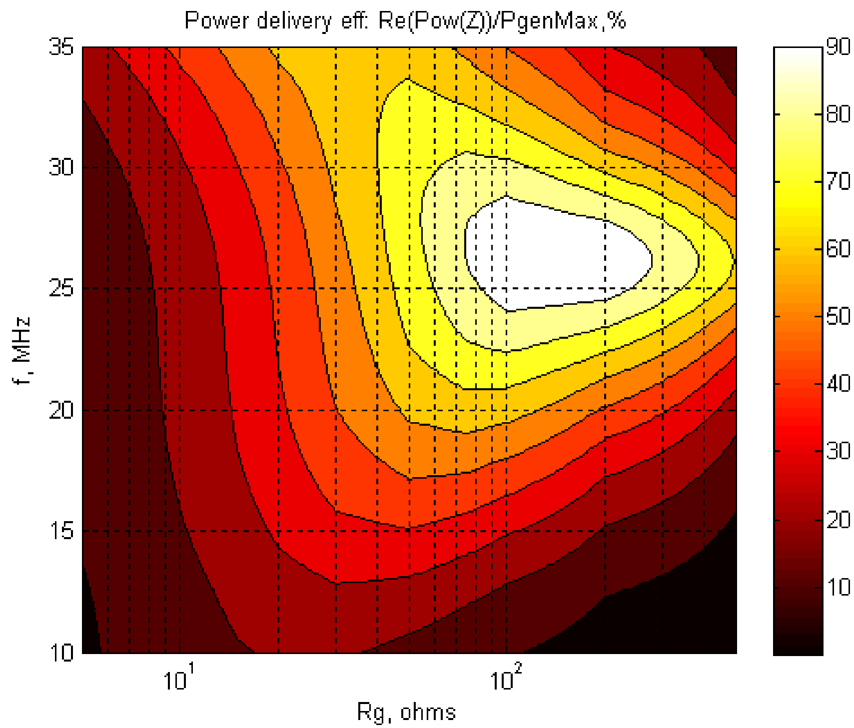
Z matching



Z matching

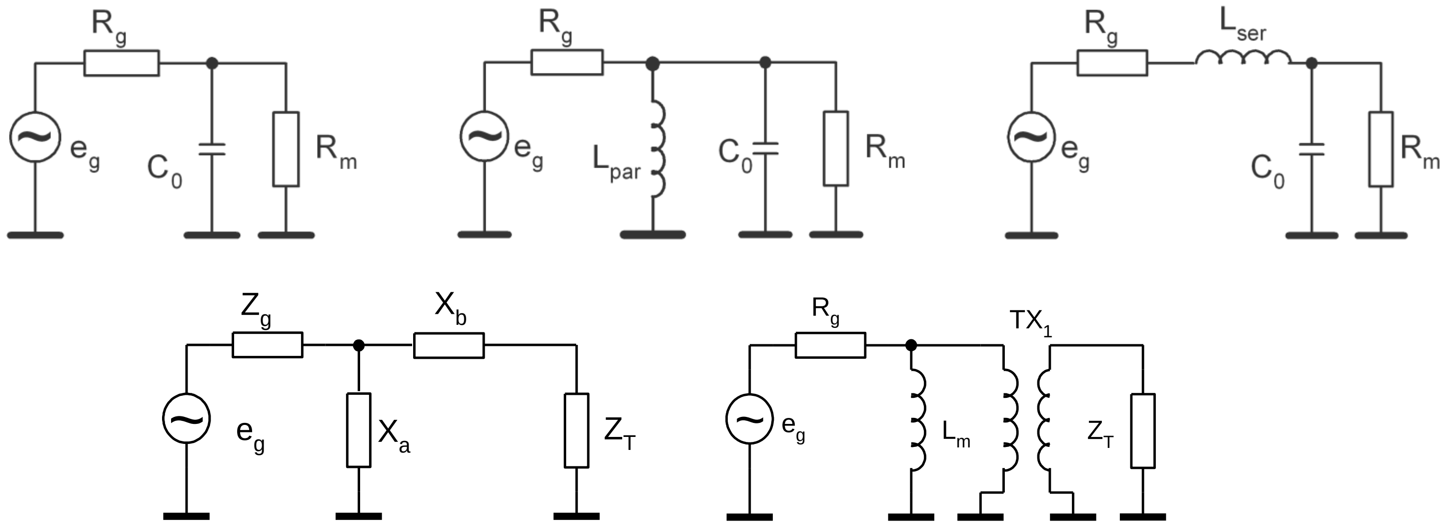
$$\eta = \frac{4R_g \operatorname{Re}(S_T)}{e_g^2} 100\% = \frac{4R_g Z_{in}}{(R_g + Z_{in})(R_g + Z_{in}^*)} 100\%$$





Power delivery efficiency vs. frequency and Zout

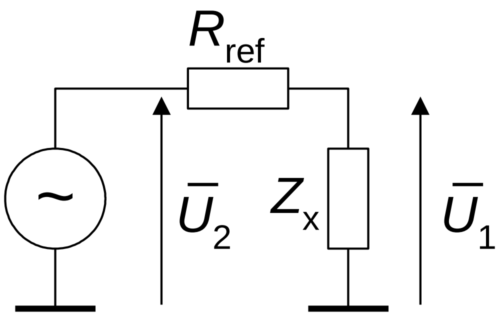
Z matching



$$L_{par} = -\frac{X_{par}\big|_{\omega=\omega_s}}{\omega_s} \qquad L_{ser} = \frac{X_T}{\omega_s} \qquad X_a = -\frac{R_g^2 + X_g^2}{QR_g + X_g} \qquad X_b = QR_T - X_T$$

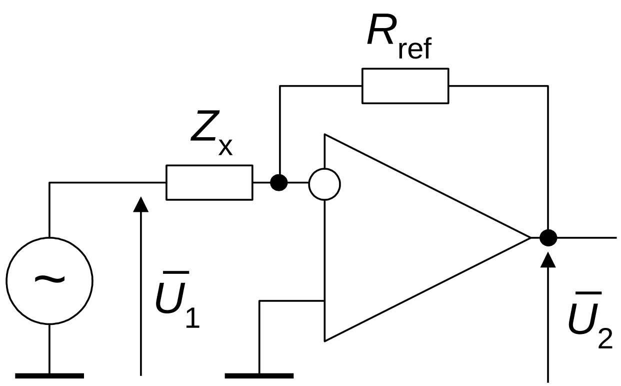
$$Q = \pm\sqrt{\frac{R_g}{\mathrm{Re}(Z_T)}\left[1+\left(\frac{X_g}{R_g}\right)^2\right]-1} \qquad L_m = \frac{10R_g}{\omega_s} \qquad n = \sqrt{\frac{|Z_T|}{R_g}}$$

Z measurement



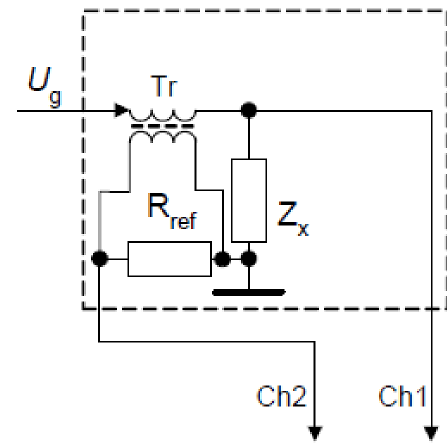
$$Z_x = \frac{\bar{U}_1}{\bar{U}_2 - \bar{U}_1} R_{ref}$$

I-V



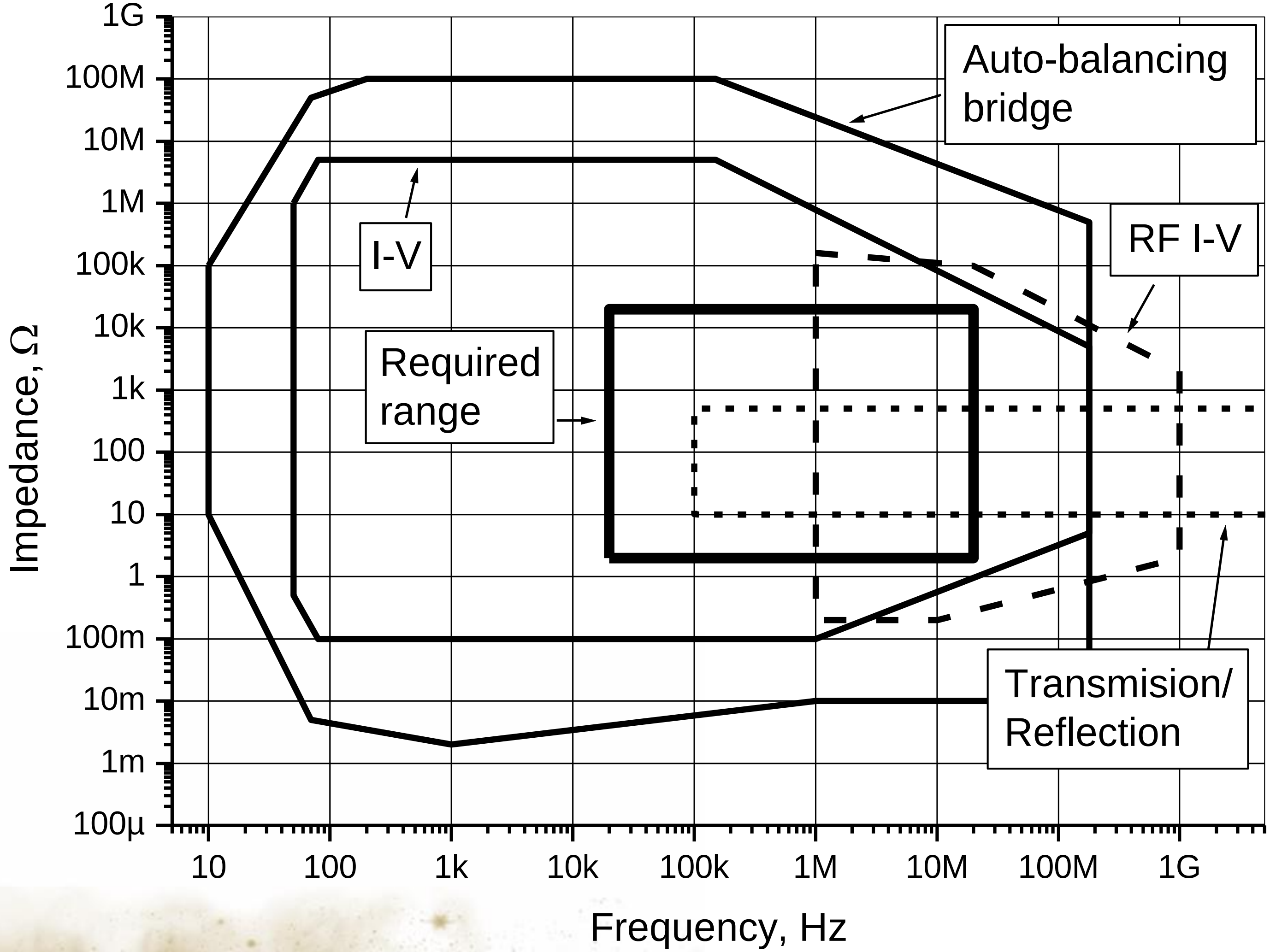
$$Z_x = \frac{\bar{U}_1}{\bar{U}_2} R_{ref}$$

ABB



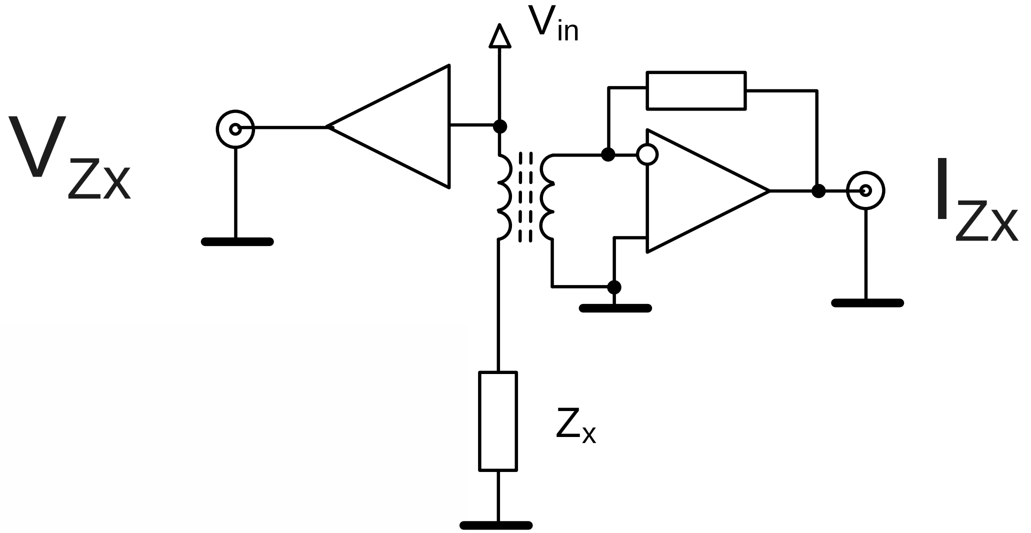
RF I-V

Z measurement



Z measurement: RF+ABB

From gen



Z measurement: OSL

$$Z_{DUT} = \frac{Z_{xm} - Z_s}{1 - \frac{(Z_{xm} - Z_s)}{Z_o}}$$

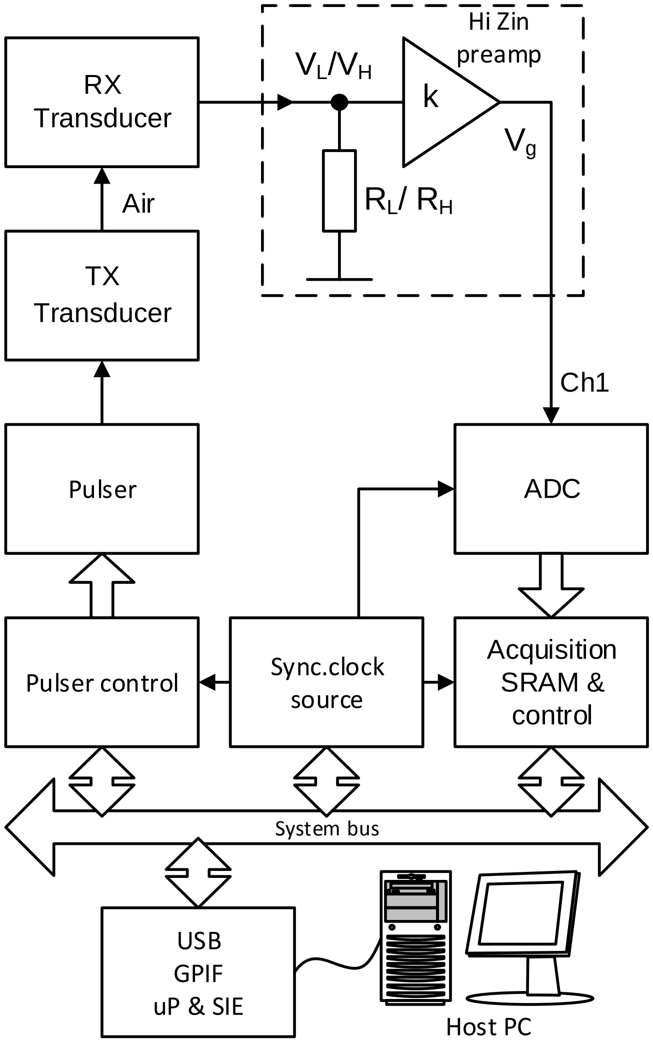
$$Z_{DUT} = Z_{std} \frac{(Z_o - Z_{stdm}) \cdot (Z_{xm} - Z_s)}{(Z_{stdm} - Z_s) \cdot (Z_o - Z_{xm})}$$

Zxm – Z measured before compensation
Zs – Z measured @ short
Z0 – Z measured @ open
Zstd – Z measured @ Zstd (known)

Z_{out} measurement

$$Z_x = \frac{(V_H - V_L)R_L R_H}{R_H V_L - R_L V_H}$$

$$V_i = \frac{(R_H - R_L)V_H V_L}{k(R_H V_L - R_L V_H)}$$



R

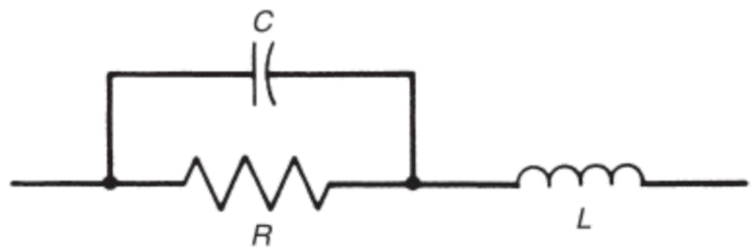
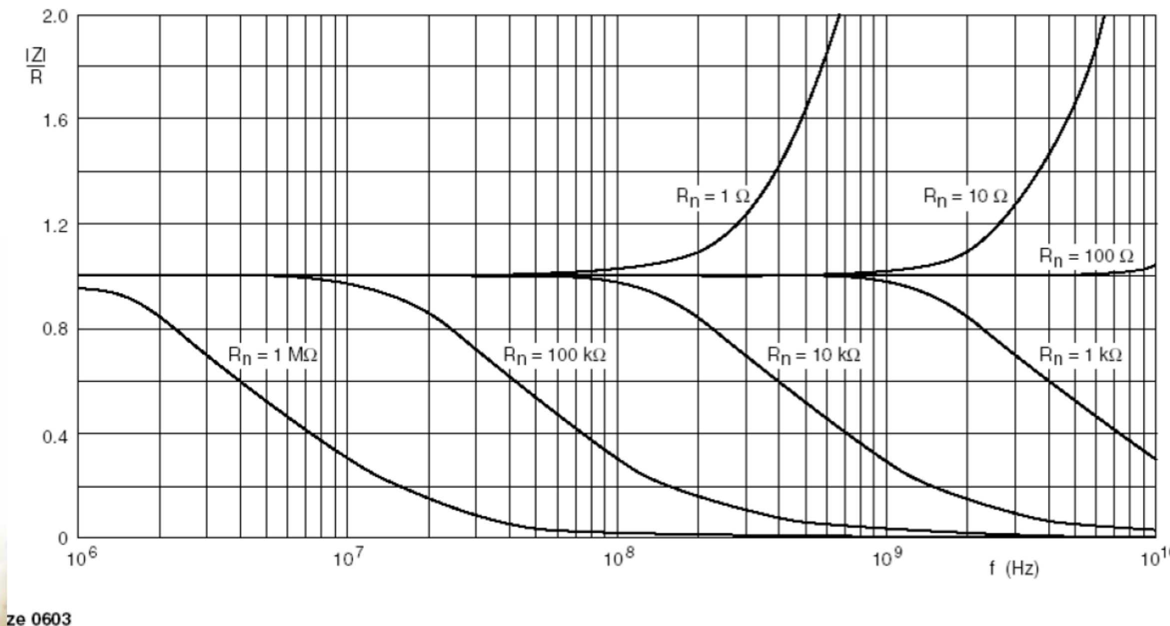


FIGURE 5-14. Equivalent circuit for a resistor.

1206:0.05pF+2nH/ 0805:0.09pF+1nH/ 0603: 0.05pF+0.4nH



R

RESISTORS	CARBON FILM	METAL FILM	THICK FILM	METAL FOIL	CARBON COMPOSITION	WIREWOUND	POWER METAL STRIP
Resistance Value	10 Ω to 22 MΩ	0.22 Ω to 22 MΩ	1 Ω to 100 MΩ	2 mΩ to 1 MΩ	1 Ω to 20 MΩ	0.1 Ω to 300 kΩ	0.1 mΩ to 1.0 Ω
Tolerance [%]	± 2 to ± 10	± 0.1 to ± 2	± 1 to ± 5	± 0.005 to ± 5	± 5 to ± 20	± 0.1 to ± 10	± 0.5 to ± 1
Temperature Coefficient [ppm/K]	- 200 to - 1500	± 5 to ± 50	± 50 to ± 200	± 2 to ± 50	- 200 to - 1500	± 1 to ± 200	± 30 to ± 250
Maximum Operating Temperature [°C]	+ 155	+ 155	+ 155	+ 150	+ 150	+ 400	+ 275
Rated Dissipation P_{70} [W]	0.25 to 2	0.063 to 1	0.063 to 0.25	0.25 to 10	0.25 to 1	0.25 to 100	0.1 to 5
Stability at P_{70} (1000 h) $\Delta R/R$ [%]	± 0.8 to ± 3	± 0.15 to ± 0.5	± 1 to ± 3	± 0.05	+ 4/- 6 (typical - 3)	± 1 to ± 10	± 1 to ± 2
Operating Voltage $U_{max.}$ [V]	200 to 1000	50 to 500	50 to 200	200 to 500	150 to 350	25 to 1000	$\sqrt{P_{70} \times R}$
Current Noise [μV/V]	< 1	< 0.1	< 10	< 0.025	2 to 6	negligible	negligible
Non-linearity A_3 [dB]	> 100	> 110	> 50	negligible	~ 60	negligible	negligible

C

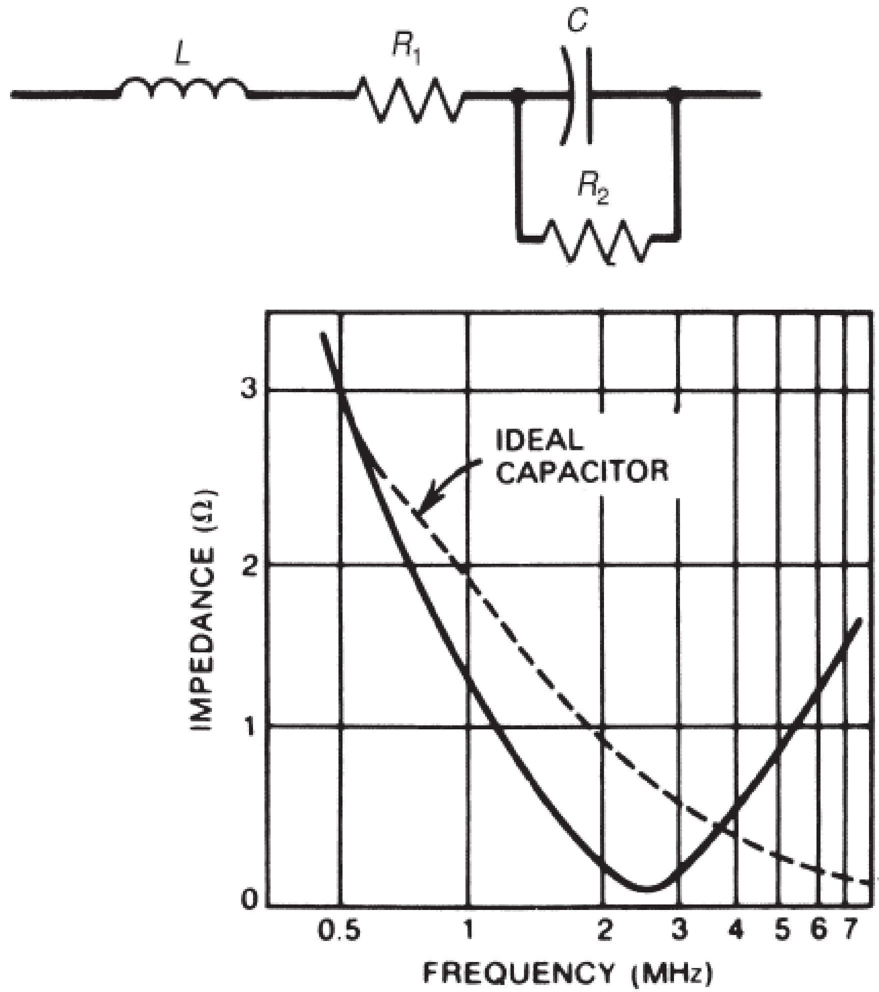


FIGURE 5-2. Effect of frequency on the impedance of a 0.1-μF paper capacitor.

C

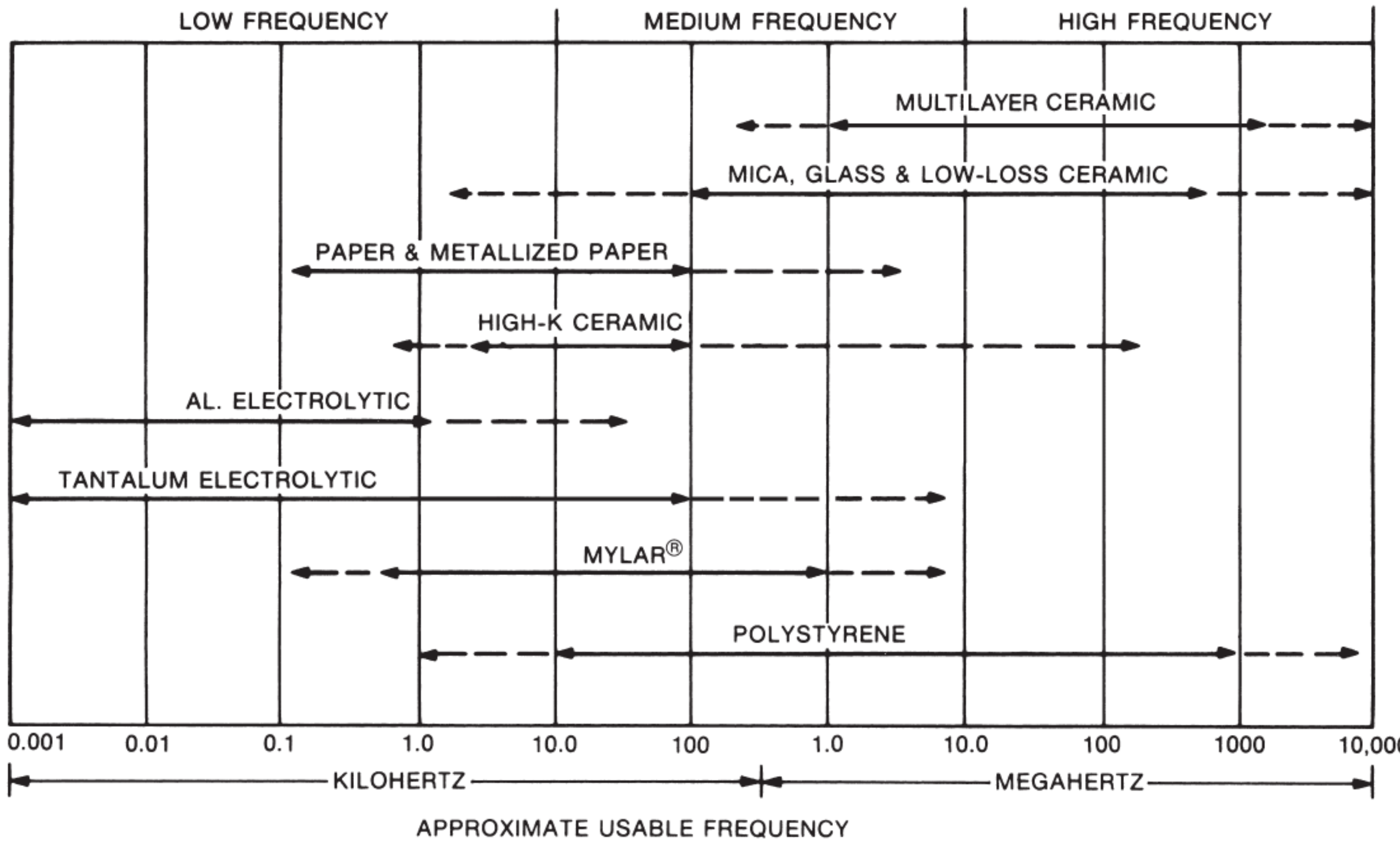


FIGURE 5-3. Approximate usable frequency ranges for various types of capacitors.

L

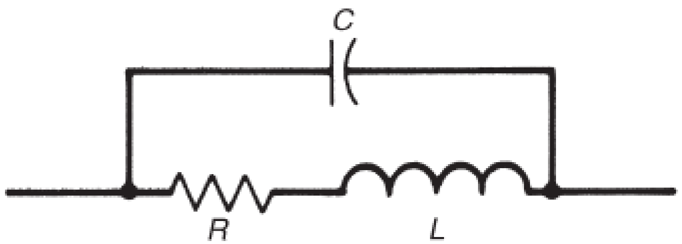


FIGURE 5-9. Equivalent circuit for an inductor.

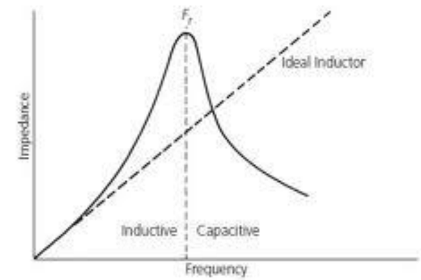
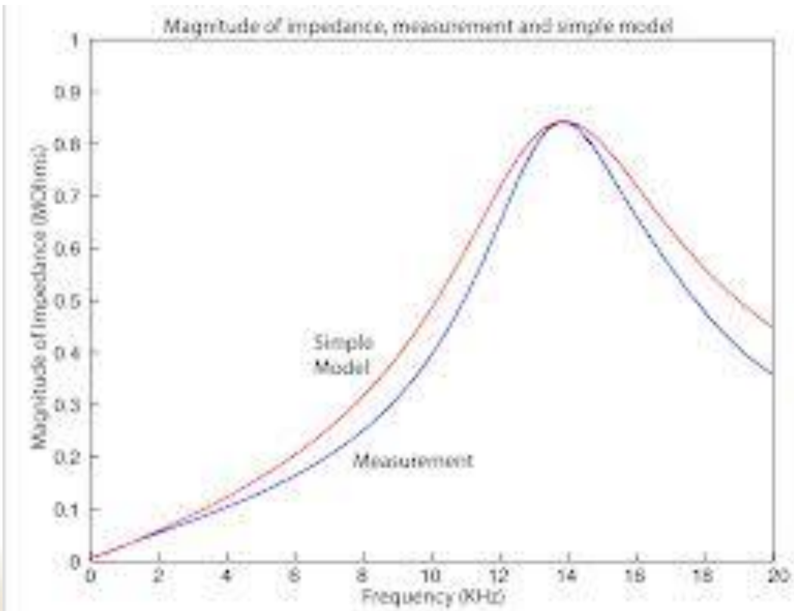
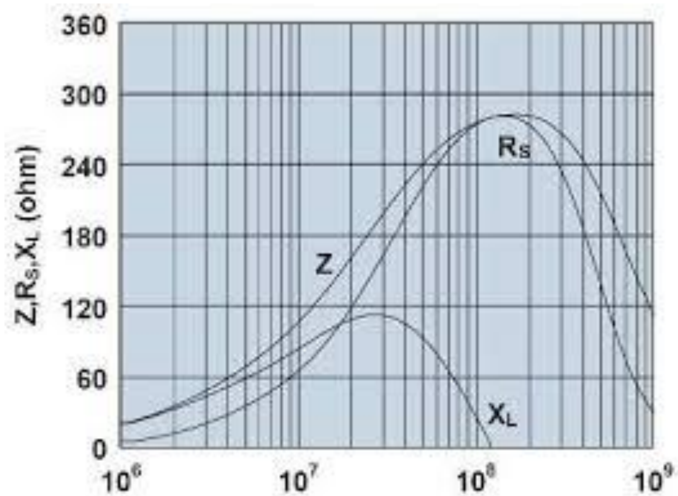
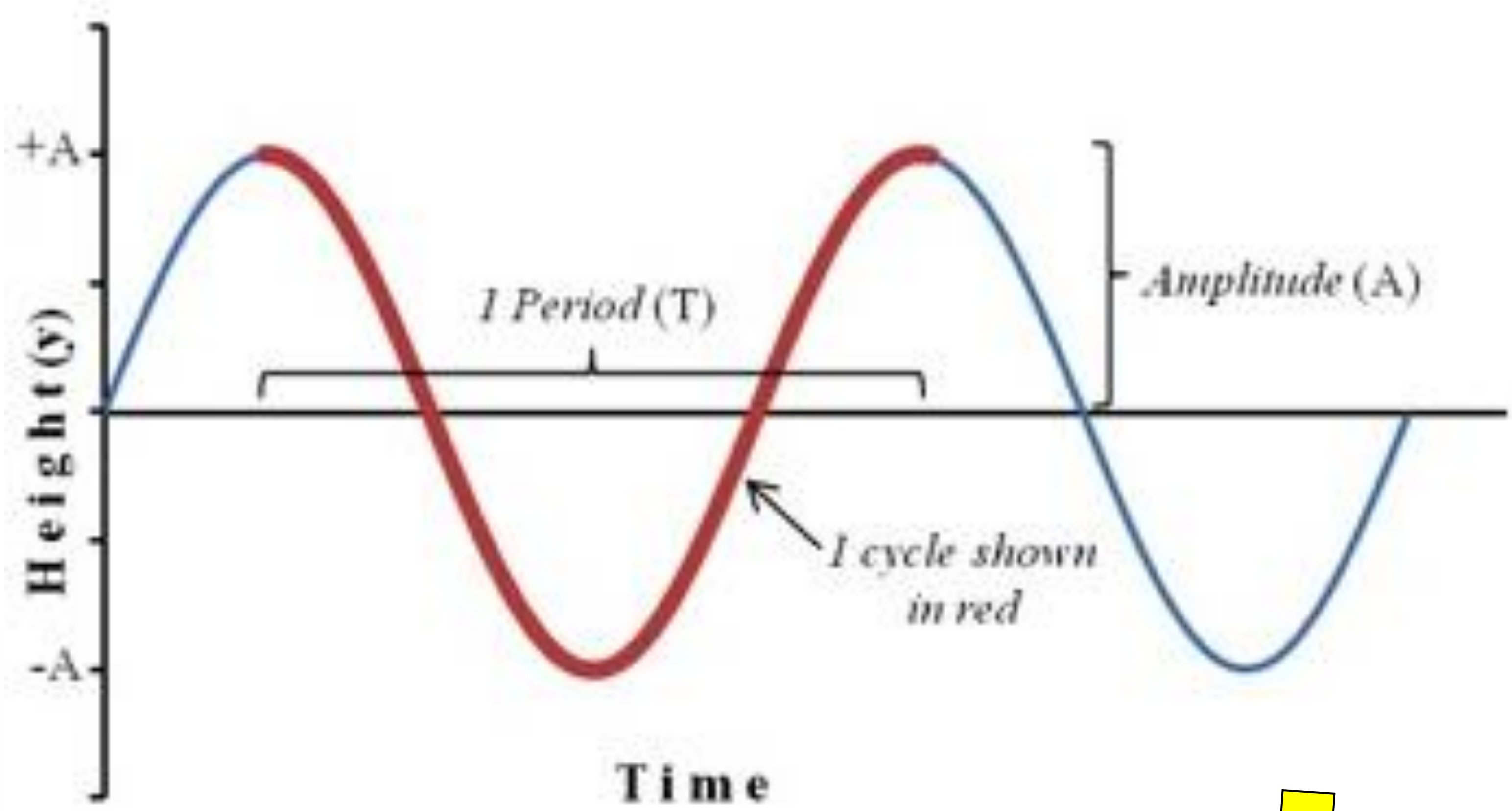


FIG. 1-16. Impedance characteristic vs. frequency for a practical and an ideal inductor.





Frequency estimation



$$x(t) = A\sin(\omega t + \phi)$$

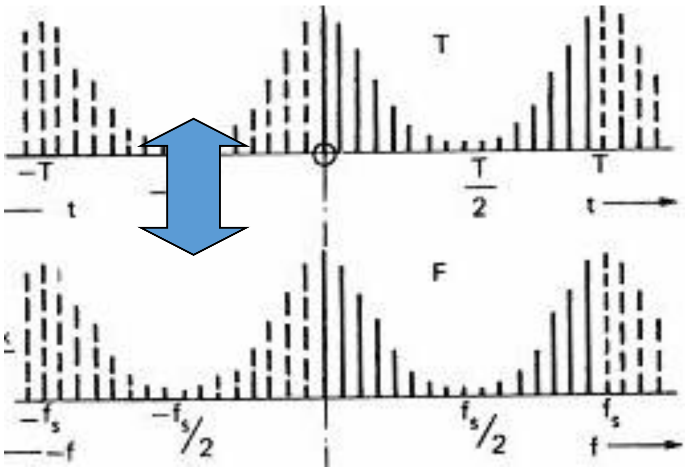
DFT

Discrete time AND frequency signal is presented only at discrete time AND frequency samples. Such signal is periodic both in time AND frequency domain. Briefly: such signal is both discrete and periodic;

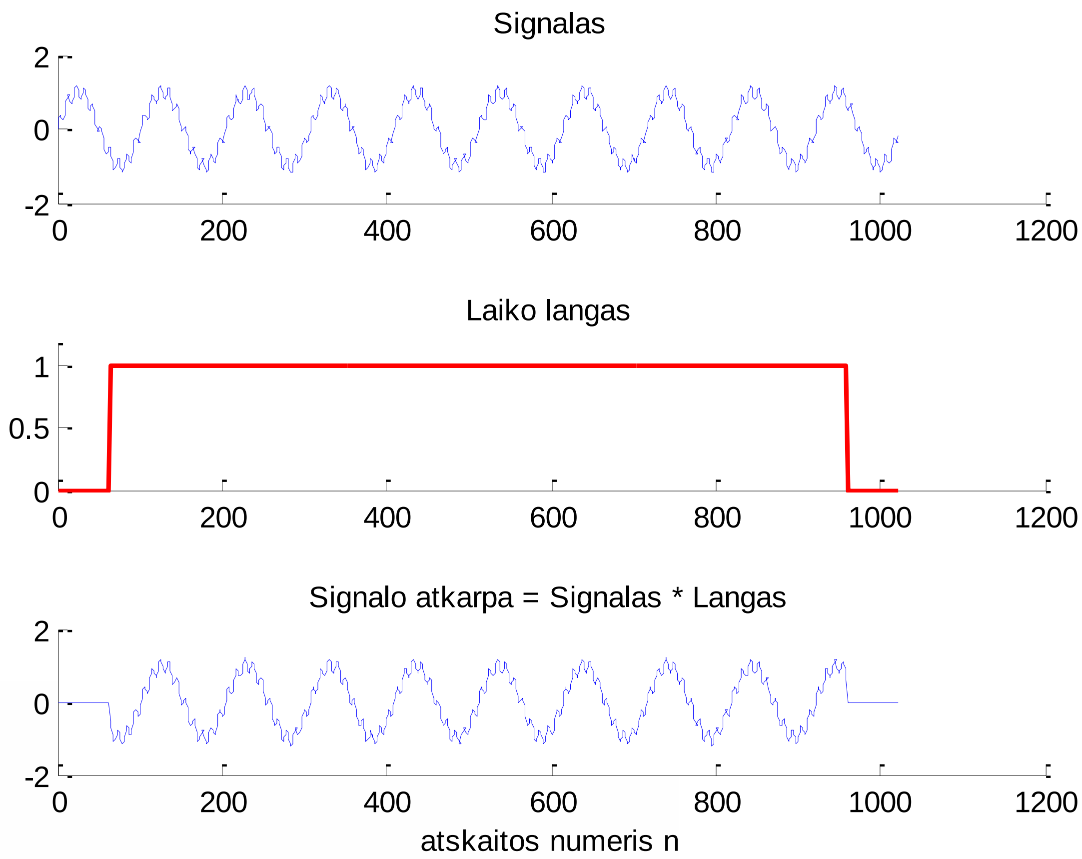
DFT-Discrete Fourier transform is used to transforms the siagnal between time and frequency axes. **FFT** is the **separate DFT case** where signal length N is log2 (4,16,32,64,128,256,512,1024 etc.)

$$S_k = \sum_{n=0}^{N-1} s_n e^{-j2\pi nk/N} = S_{k+N}$$

$$s_n = \frac{1}{N}\sum_{k=0}^{N-1} S_k e^{j2\pi nk/N} = s_{n-N}$$

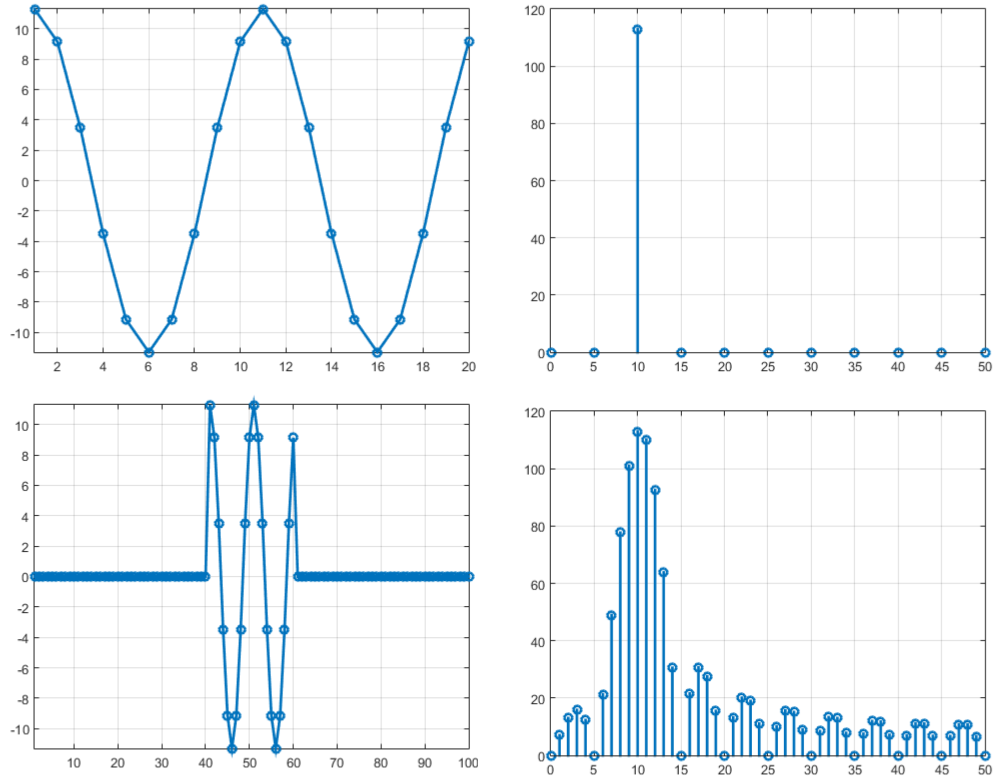


DFT frequency estimation



Windowing/gating can not be avoided

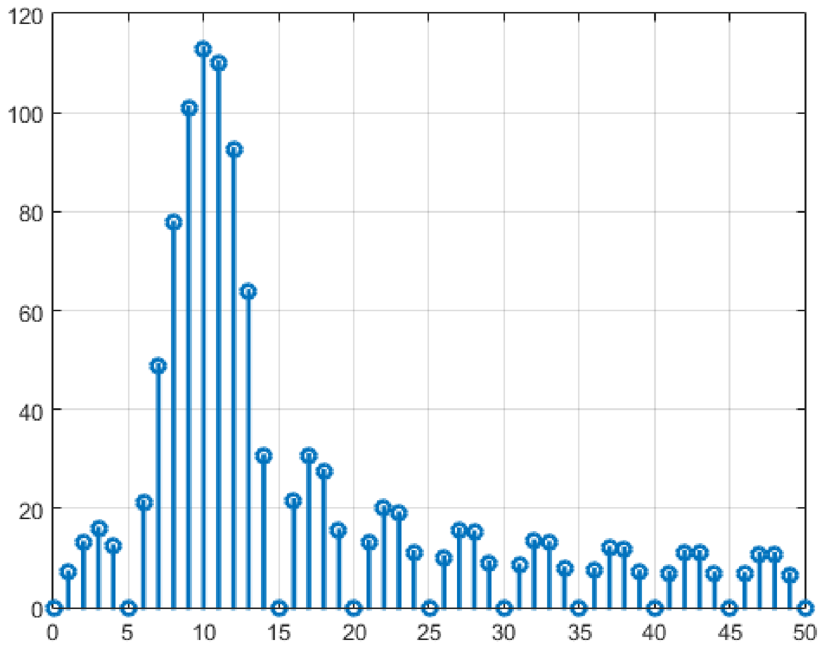
DFT frequency estimation



Record length matches period only in rare occasions

Result of windowing + wrong record length

DFT frequency estimation



DFT frequencies:

$$f = \begin{cases} -f_N, n=1 \\ -2f_N \ldots -\Delta f, n = 2\ldots N/2 \\ 0, n = N/2+1 \\ \Delta f \ldots f_N, n = (N/2+2)\ldots N \end{cases}$$

Actual frequency may be not on grid:

$$\Delta f = f_s/N$$

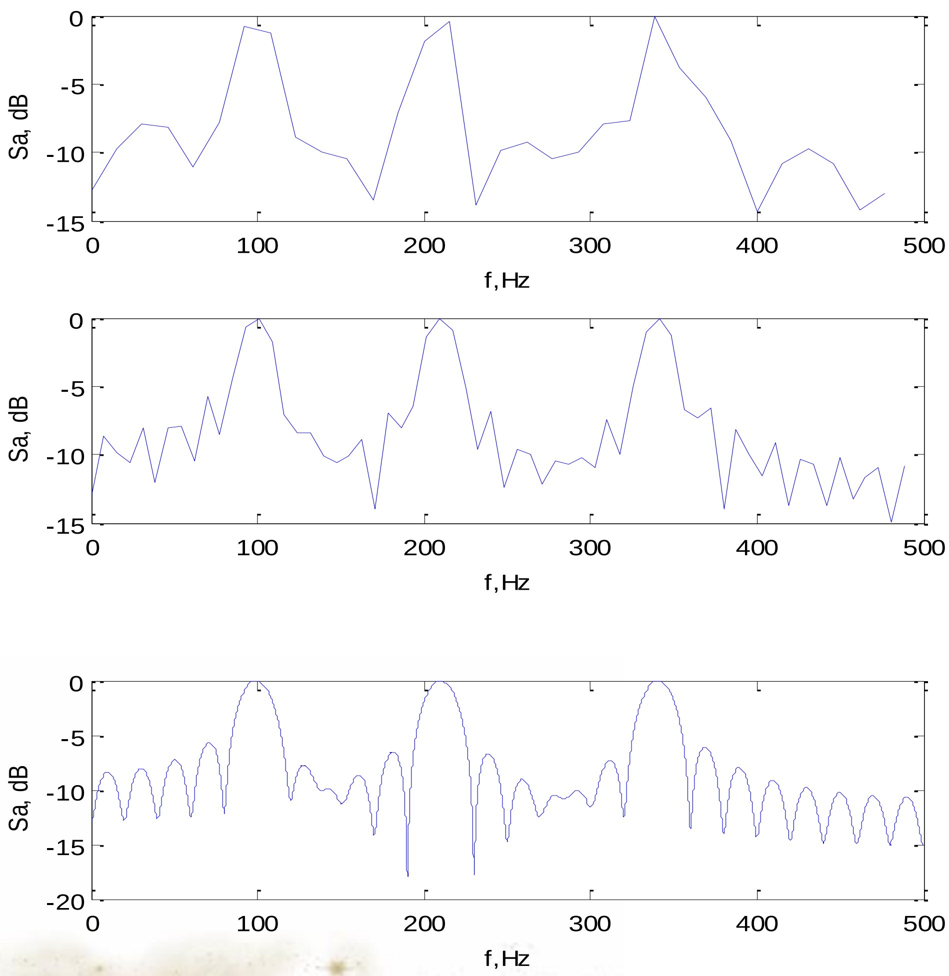
Zero-padding does reduce Δf, because N is increased

DFT frequency estimation

N=64

N=128

N= 16384



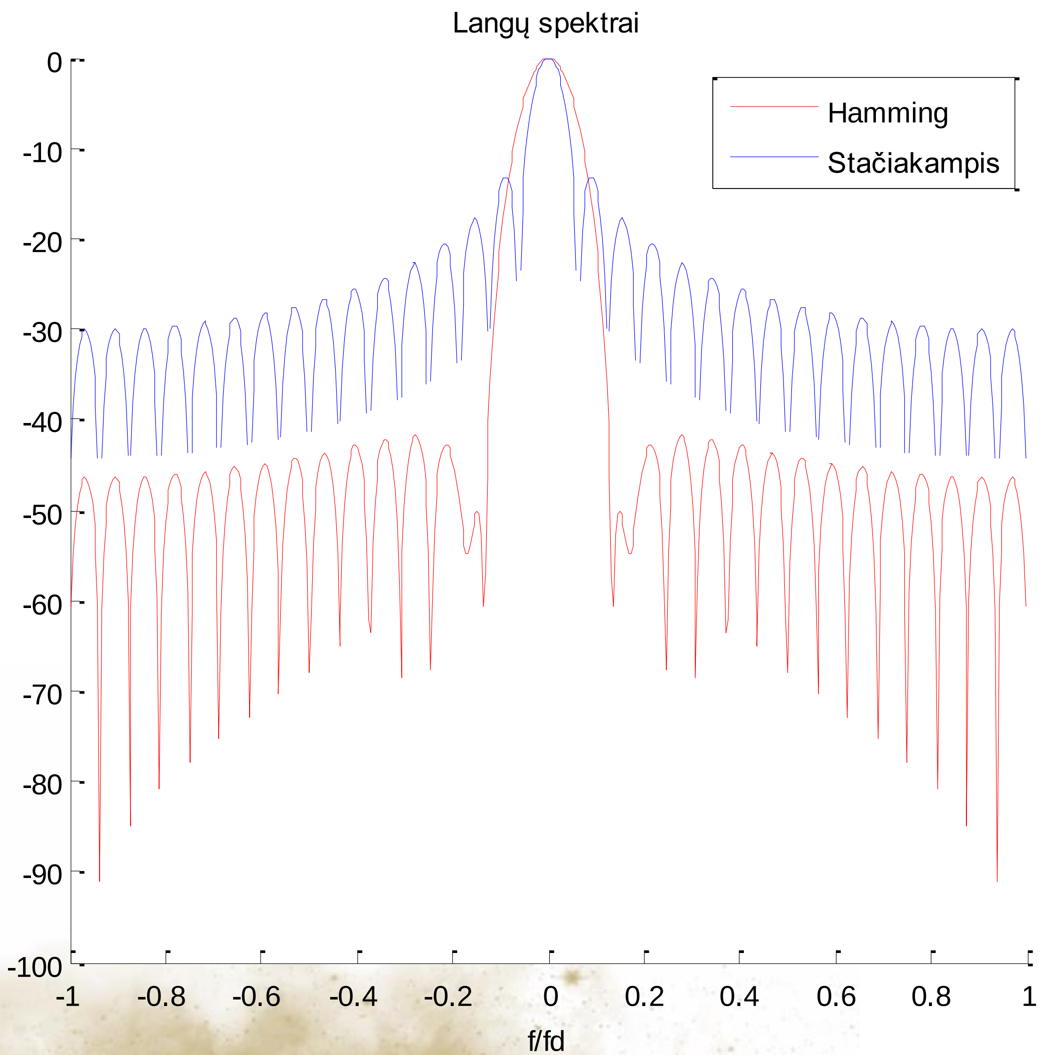
Course frequency grid

Denser grid

Note that shape of the spectral peaks remains the same

Zero-padding does not improve resolution (sharpness of the peak)

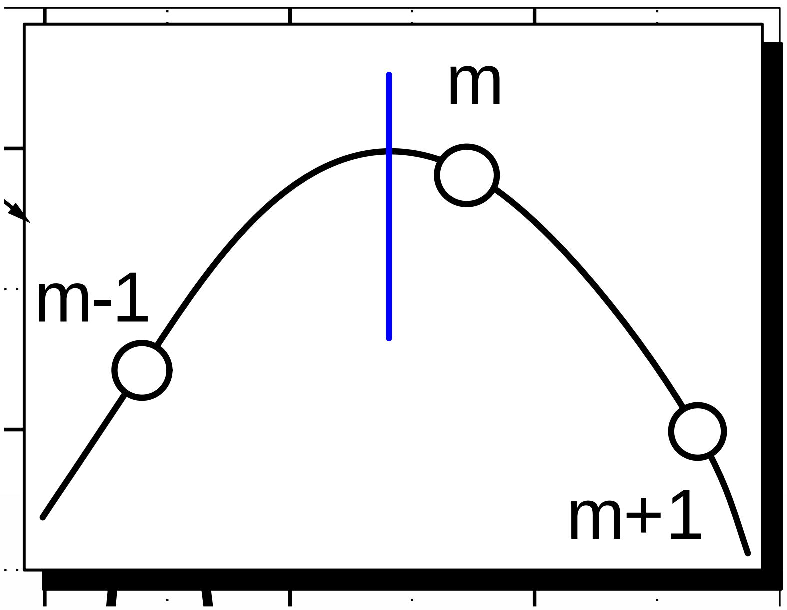
DFT frequency estimation



Other window types only worsen situation

DFT frequency estimation

Interpolation helps for single frequency



$$\Delta fcorr_p = \frac{x_{m-1} - x_{m+1}}{2(x_{m-1} - 2x_m + x_{m+1})} \cdot \Delta f$$

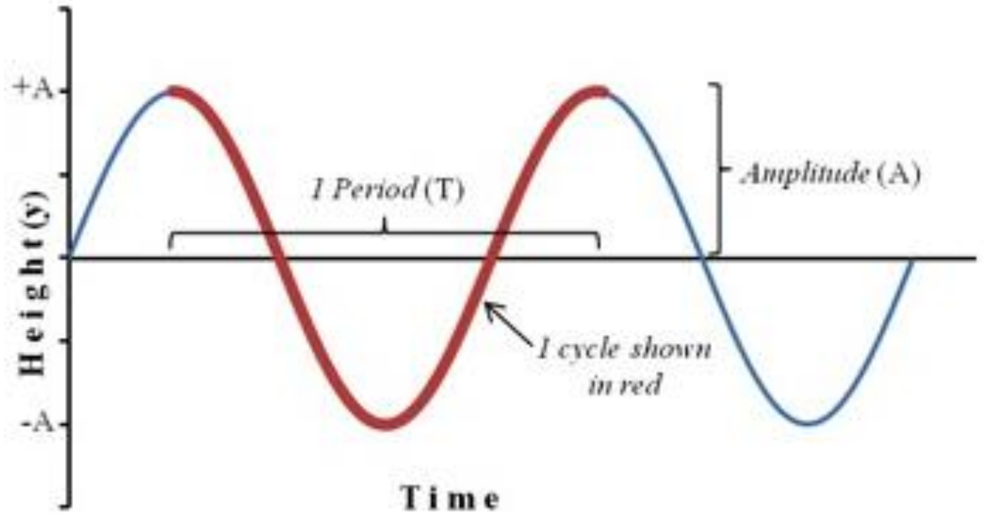
$$\Delta fcorr_G = \frac{\ln(x_{m+1}) - \ln(x_{m-1})}{\left(4\ln(x_m) - 2\ln(x_{m-1}) - 2\ln(x_{m+1})\right)} \cdot \Delta f$$

$$\Delta fcorr_{\cos} = -\frac{\theta}{{}_s\omega_0} \cdot \Delta f \quad \omega_0 = \arccos\left(\frac{x_{m-1} + x_{m+1}}{2x_m}\right) \; \theta = \arctan\left(\frac{x_{m-1} - x_{m+1}}{2x_m \sin\omega_0}\right)$$

Sine parameters estimation:
4 parameters fit

Probing frequency is unknown:
4 parameters fit:
Vm, V0, phase, frequency

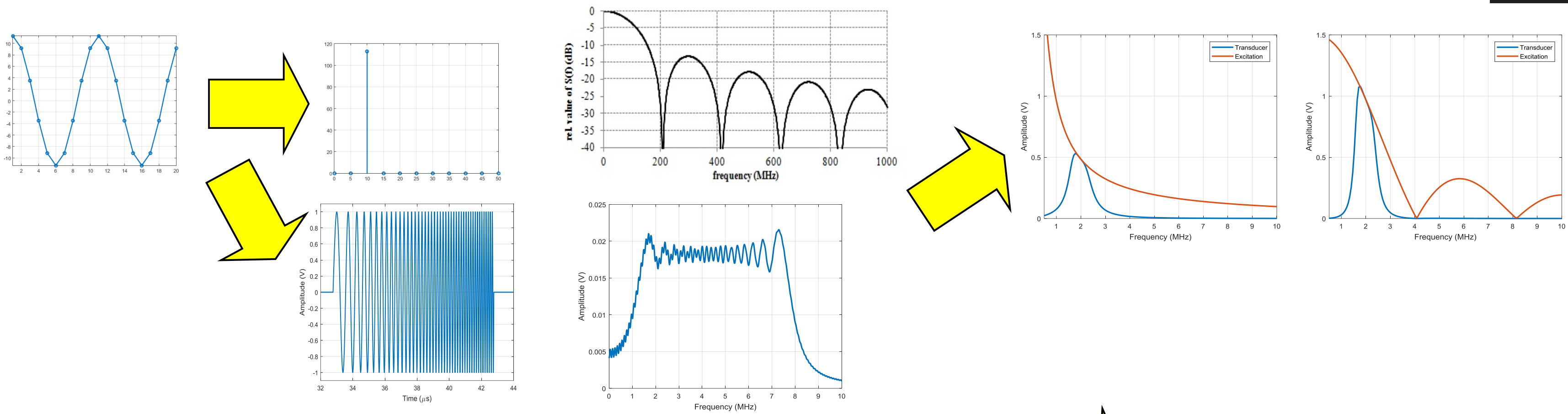
$$x(t) = A\sin(\omega t + \phi)$$

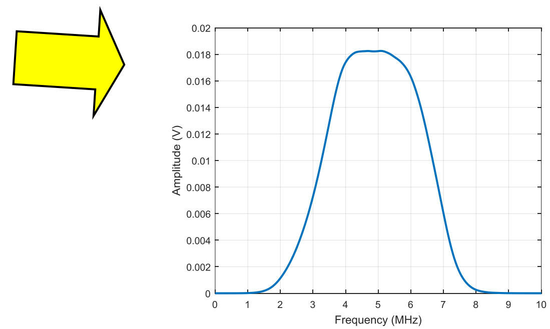


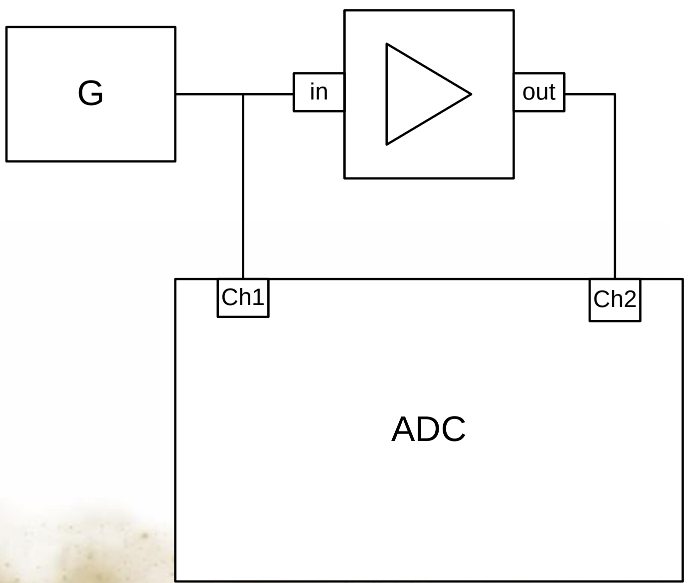
$$r_n = y_n - A\cos(2\pi f t_n) - B\sin(2\pi f t_n) - C$$

https://se.mathworks.com/matlabcentral/fileexchange/23214-four-parameter-sinefit

AC measurement using broadband signal







$$\begin{matrix} s_{in}(t) & \xrightarrow{\text{DFT}} \\ s_{uot}(t) & \xrightarrow[\text{DFT}]{} \end{matrix} \qquad \frac{S_{in}(f)}{S_{out}(f)} = T(f)$$

End: AC measurements

Conclusions:

- Harmonic signal parameters (amplitude, phase, frequency) estimation using DFT will not produce reliable results unless signal closely matches Fourier bin frequency
- Bin frequency depends on sampling frequency and record length: not all frequencies can satisfy bit match
- High accuracy, automated AC measurements can be carried out using SWC (derivative of TDFT)
- Same clock frequency should be used both for probing sine production and ADC (this is why measurement equipment has 10MHz sync signal)
- Measurements using SWC are accurate, but time-consuming. Broadband signals can be used for AC estimation via DFT.
- Proper use of windows is essential in such case

END

Thank you for your attention