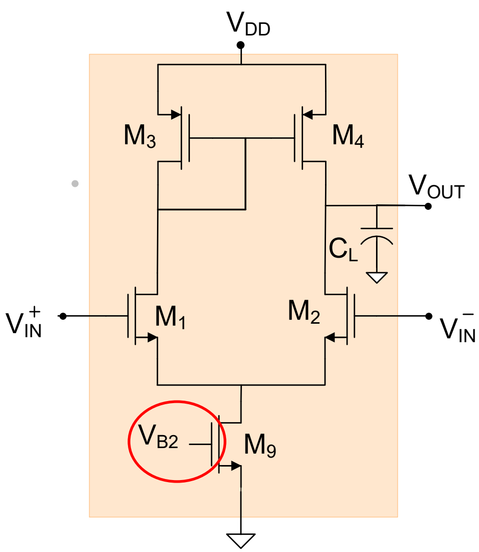
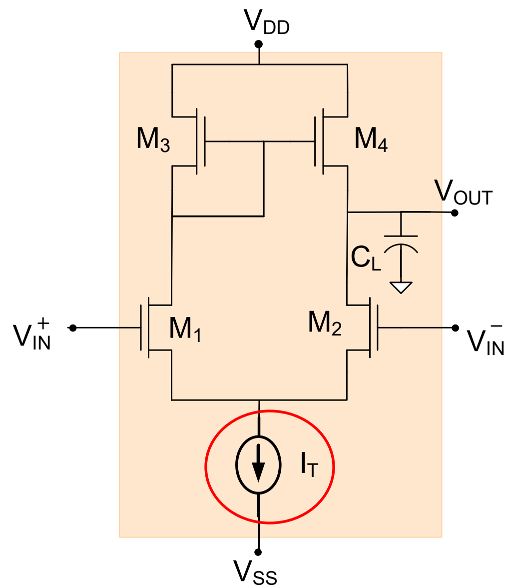
EE 435 Lecture 42

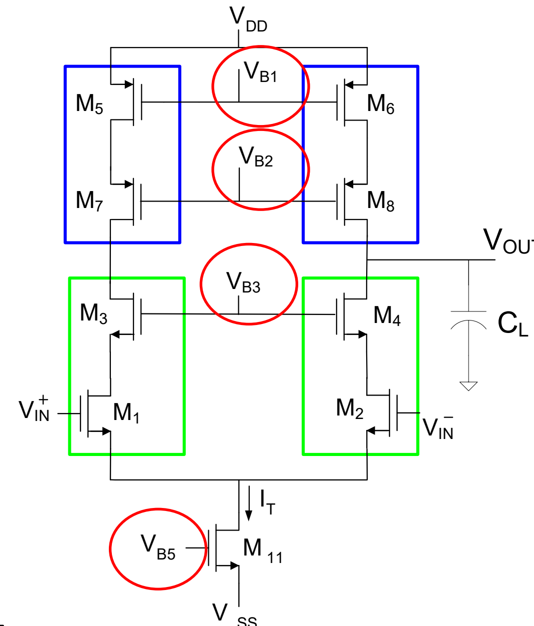
Noise in Analog Circuits
Dynamic Range

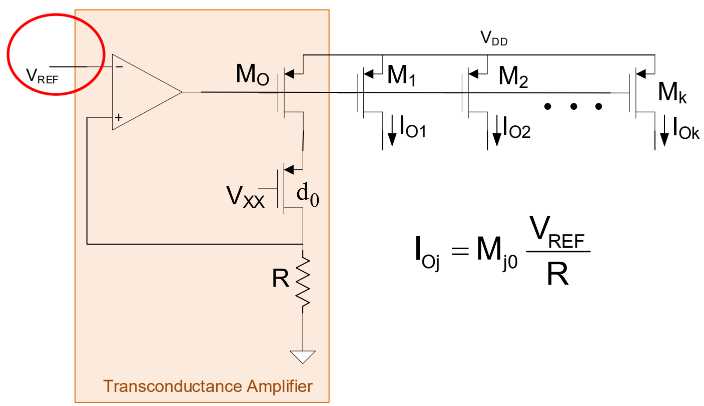
Bias Voltages/Currents and References

How are these voltages and currents generated?

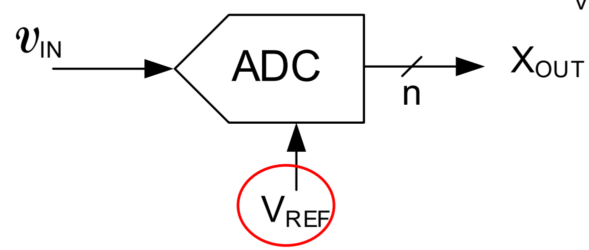


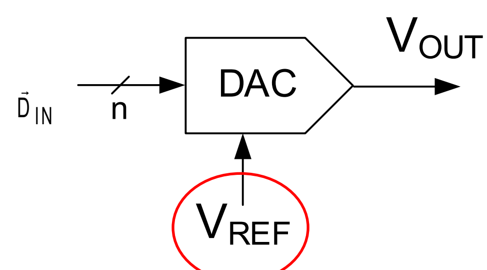




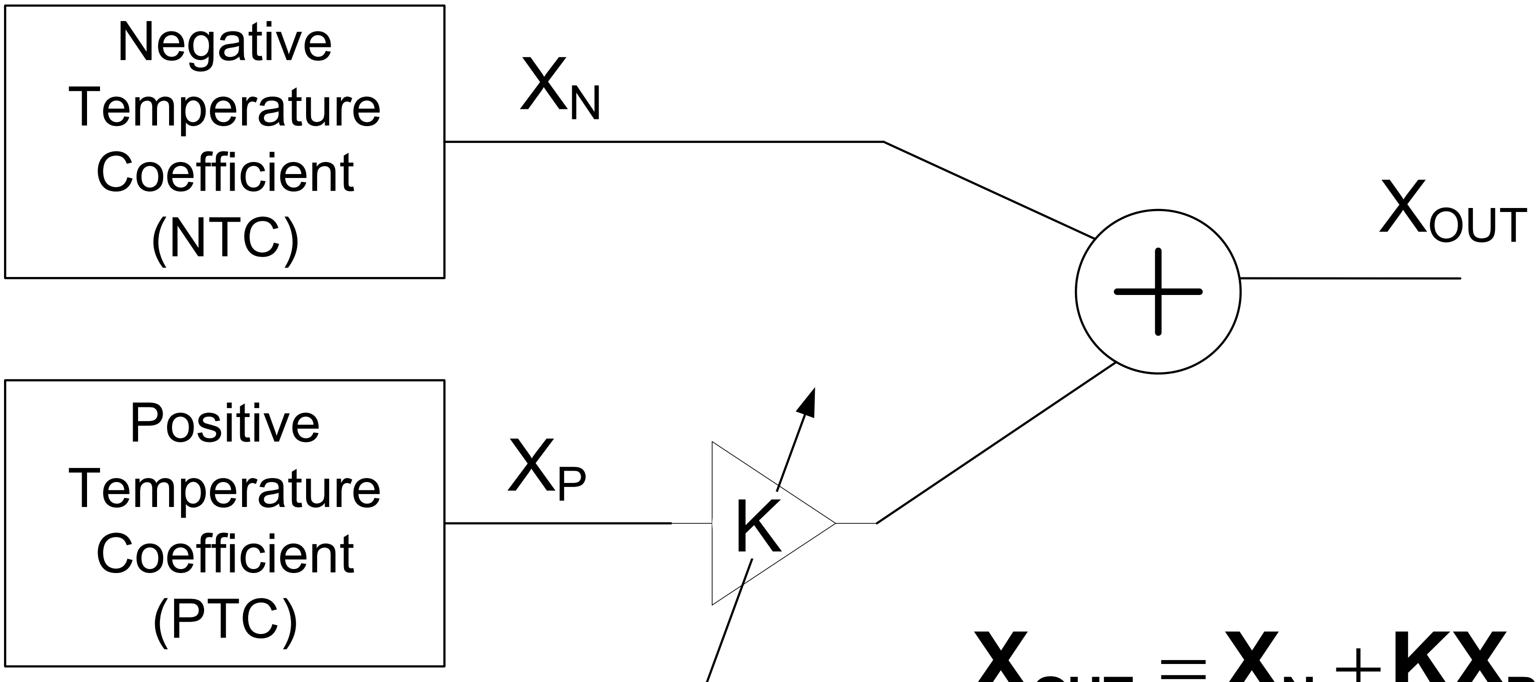


$$I_{Oj} = M_{j0} \frac{V_{REF}}{R}$$





Standard Approach to Building Voltage References



$$\mathbf{X_{OUT} = X_N + KX_P}$$

Pick K so that at some temperature T_0, $\left.\frac{\partial(\mathbf{X_N} + \mathbf{KX_P})}{\partial \mathbf{T}}\right|_{\mathbf{T=T_0}} = \mathbf{0}$

Bandgap Voltage References

Consider two BJTs (or diodes)

Key observation about diodes and diode-connected BJTs

1. If ratio of currents in two devices is constant, ΔV_{BE} is PTAT independent of the temperature dependence of the currents and sensitivity is small
2. VBE has a negative temperature coefficient for a wide range of temperature dependent or temperature independent currents and sensitivity is much larger than that of ΔV_{BE}

Bamba Bandgap Reference

$$I_{R0} = \frac{\Delta V_{BE}}{R_0}$$

$$I_{R1} = \frac{V_{BE1}}{R_1}$$

$$I_{R2} = I_{R1}$$

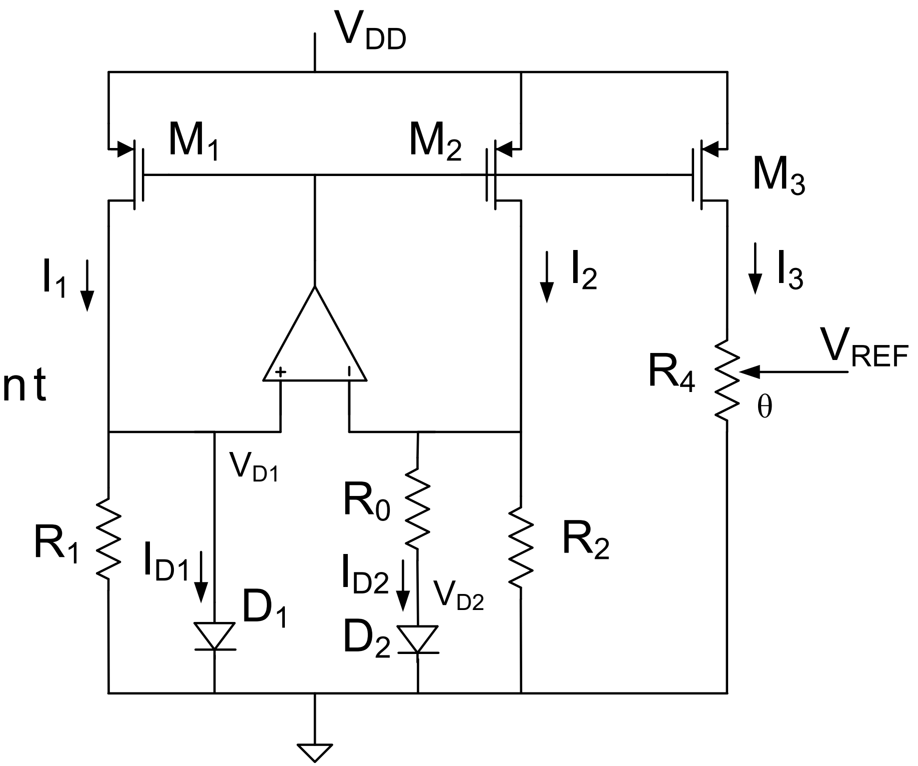
$$I_2 = I_{R0} + I_{R2}$$

$$I_3 = KI_2$$

K is the ratio of I_3 to I_2

$$V_{REF} = \theta I_3 R_4$$

$$\frac{I_{D2}}{I_{D1}} = \text{constant}$$



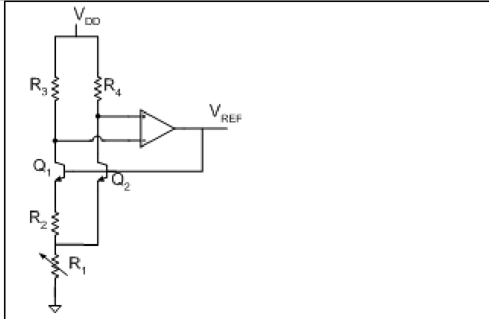
Substituting, we obtain

$$V_{REF} = \theta K R_4 \left(\frac{V_{BE}}{R_1} + \frac{\Delta V_{BE}}{R_0} \right)$$

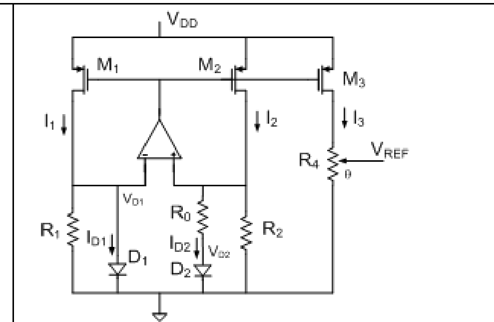
$$V_{REF} = \theta K \frac{R_4}{R_1} \left(V_{BE} + \frac{R_1}{R_0} \Delta V_{BE} \right)$$

$$V_{REF} = a_{11} + b_{11} T + c_{11} T \ln T$$

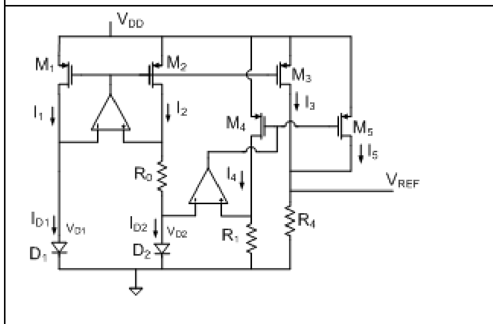
Review from Last Lecture



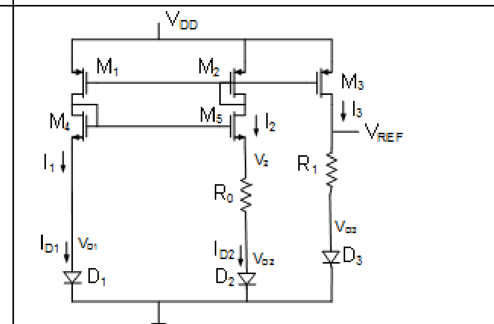
Brokow



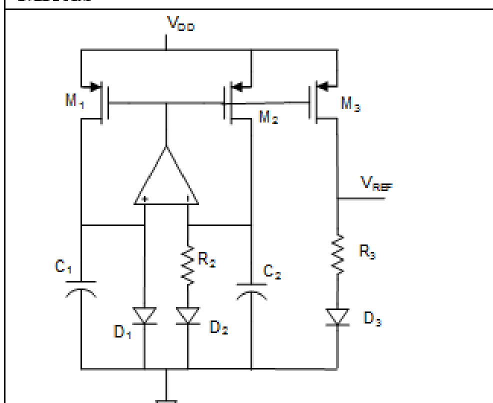
Banba



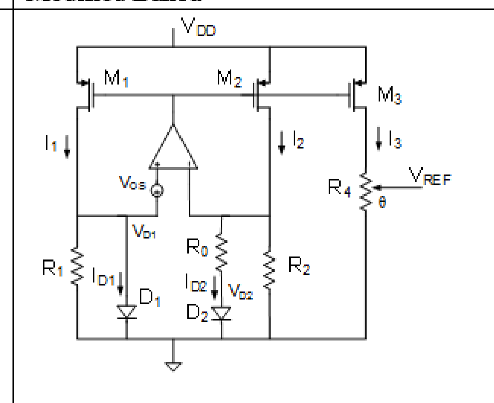
Mietus



Modified Banba

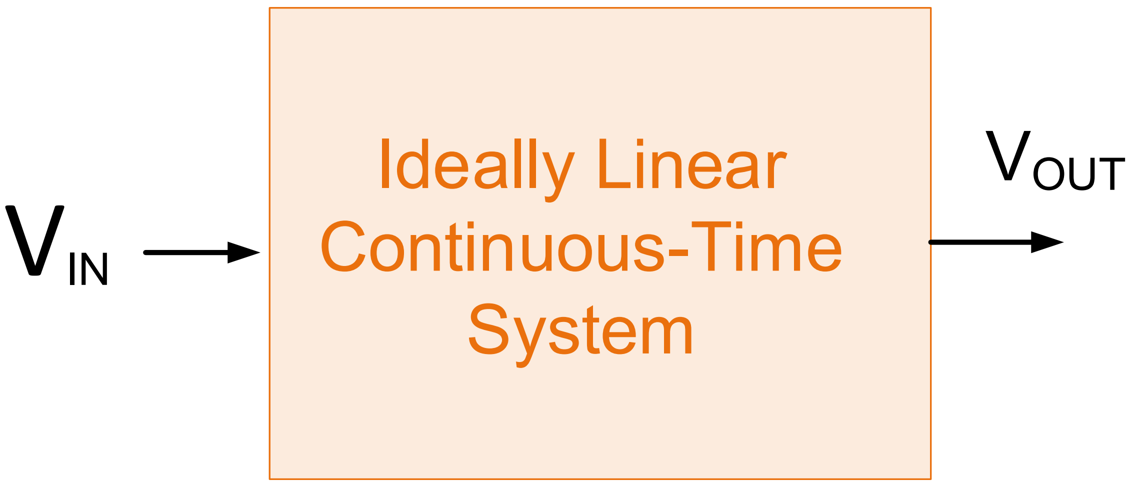


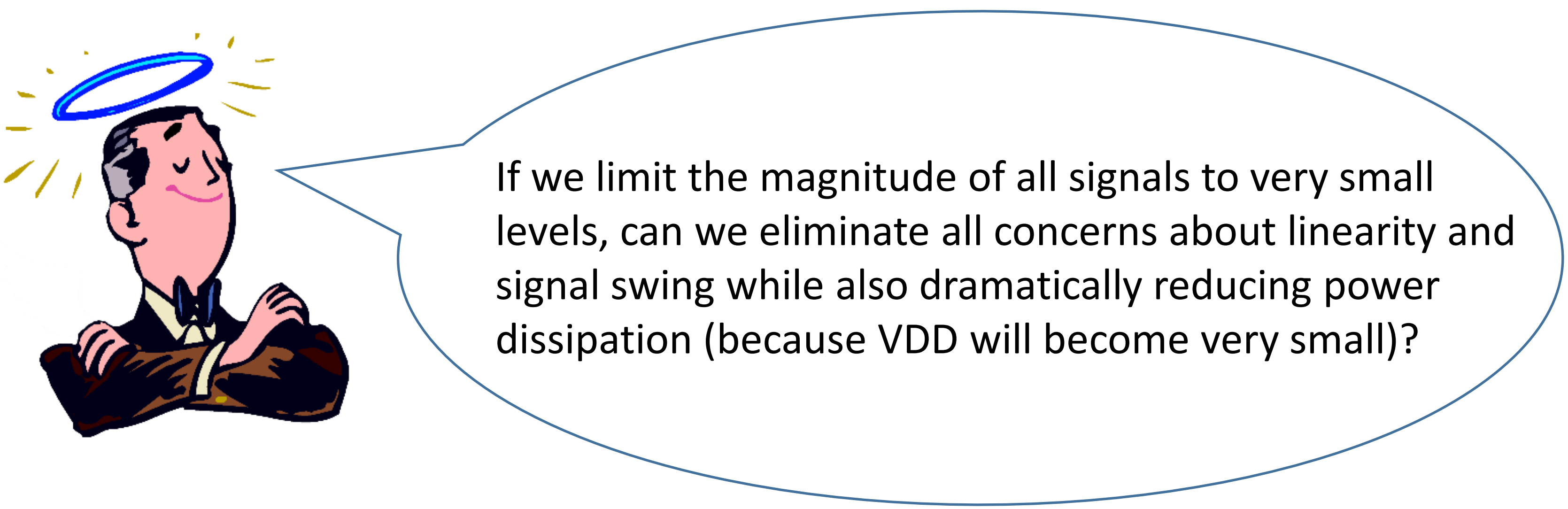
Modified Mietus



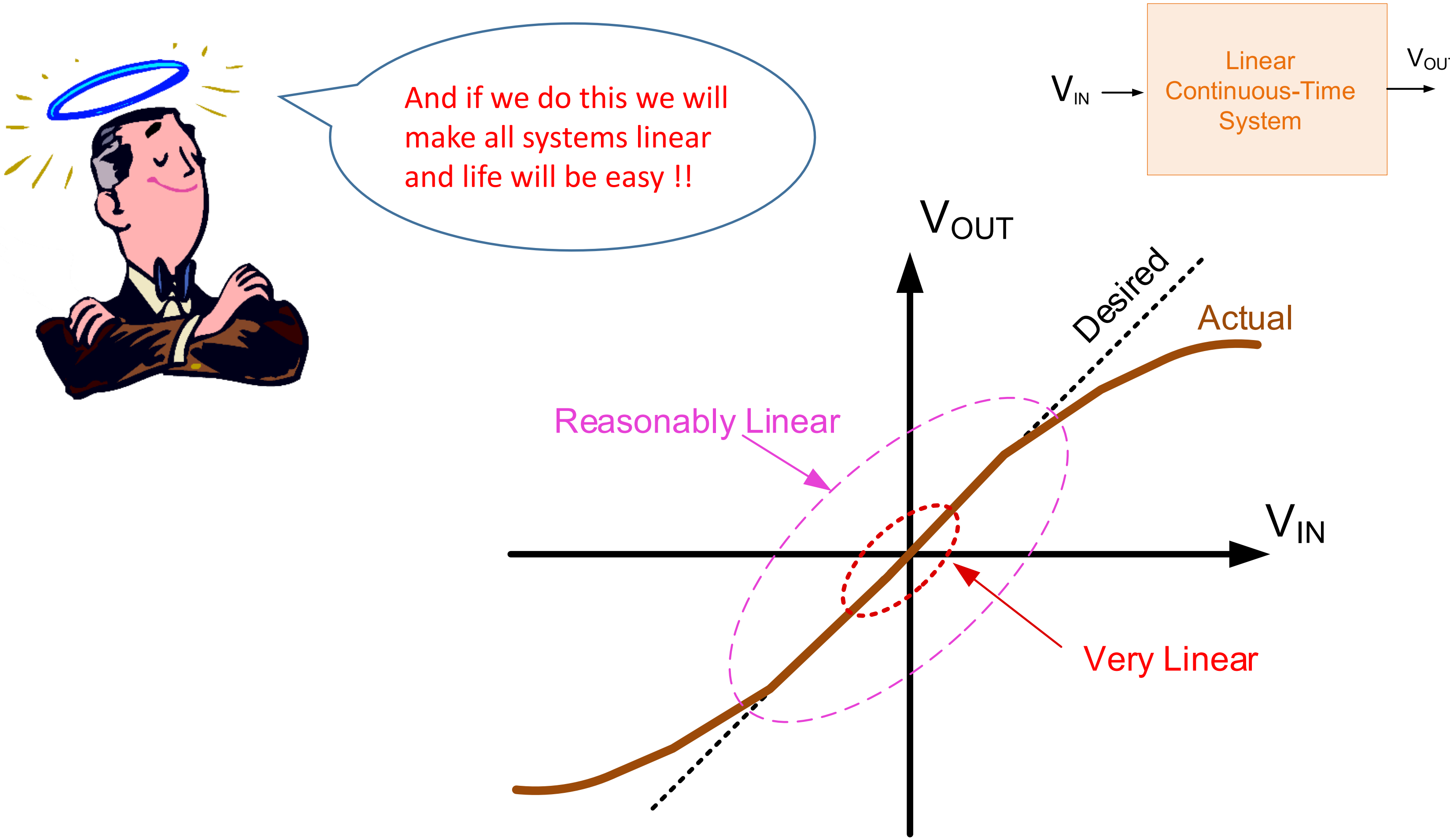
Modified Banba

Signal to Noise Ratios (SNR) and Signal to Noise and Distortion (SNDR), Signal Swing, and Linearity



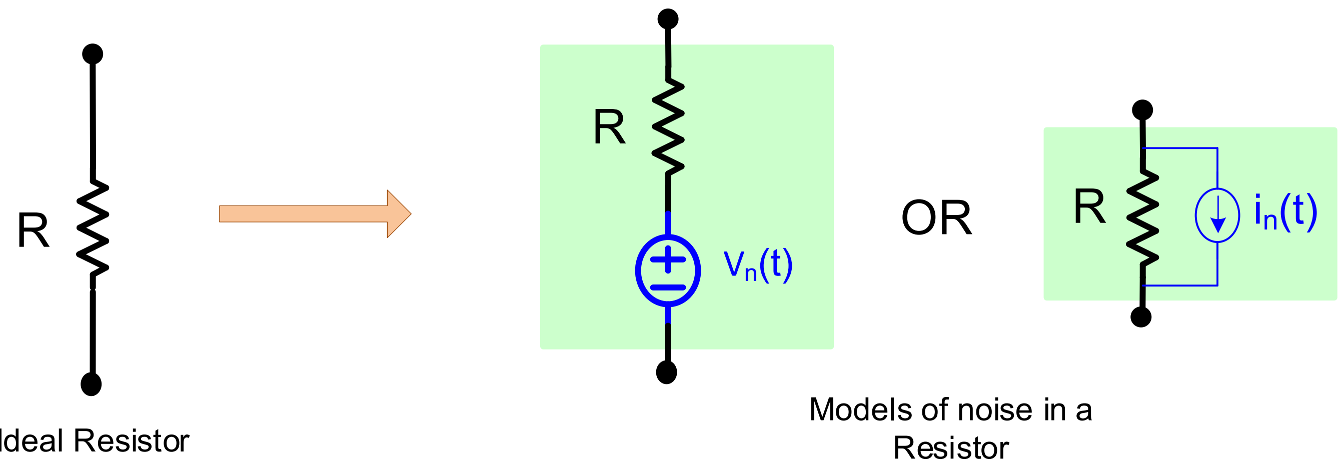


Signal to Noise Ratios (SNR) and Signal to Noise and Distortion (SNDR), Signal Swing, and Linearity



Noise is a random time-domain signal that characterizes movement of electrons in devices

Noise in Resistors

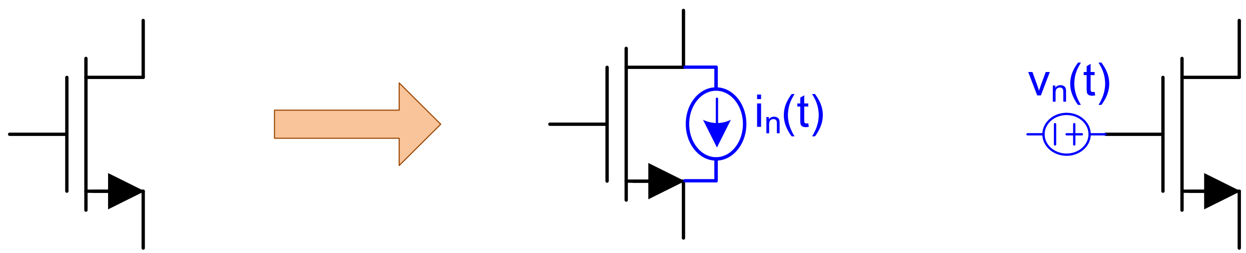


Either model can be used

Noise Analysis will give same results for either model

Noise is a random time-domain signal that characterizes movement of electrons in devices

Noise in MOS Transistors



Noiseless MOSFET

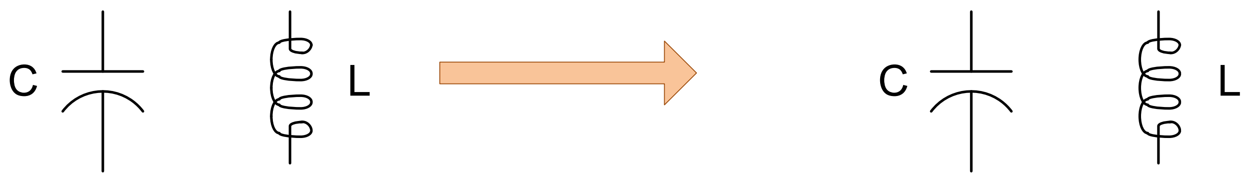
Models of noise in a MOS transistor

Either model can be used

Noise Analysis will give same results for either model

Noise is a random time-domain signal that characterizes movement of electrons in devices

Noise in Capacitors and Inductors (ideal)

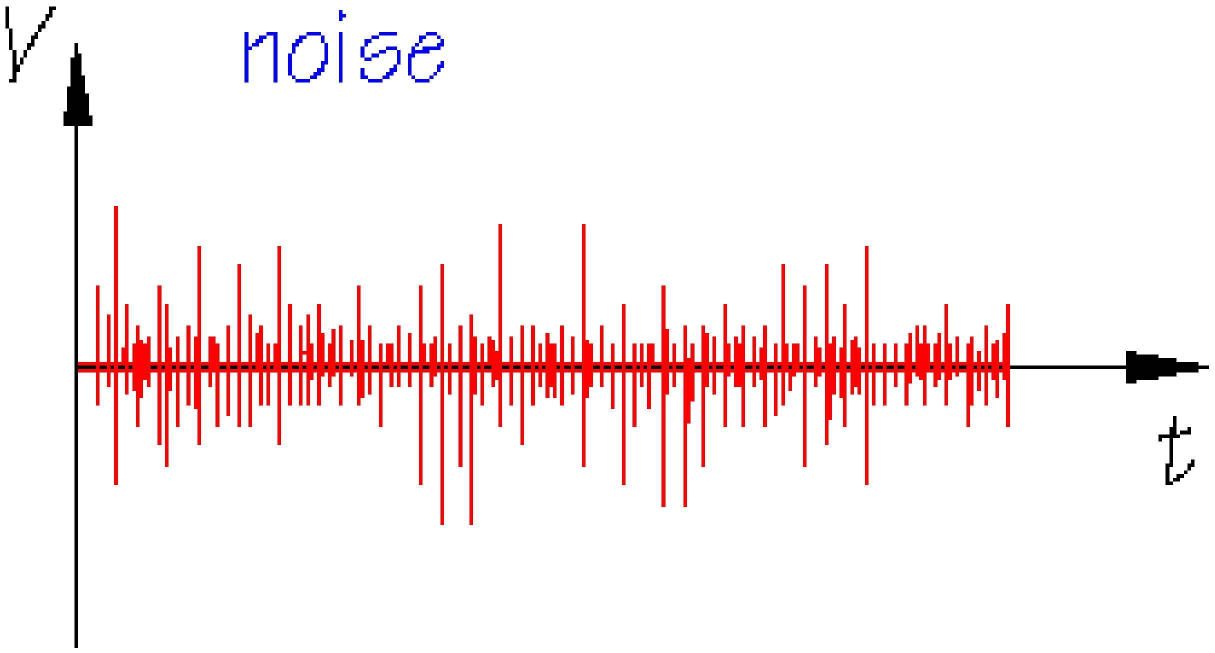


Ideal Inductor
or Capacitor

Models of noise in
Inductor or Capacitor

Capacitors and Inductors (ideal) are noiseless !

Typical noise waveform for a resistor

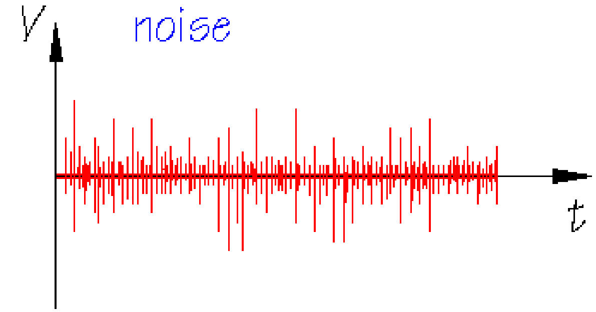


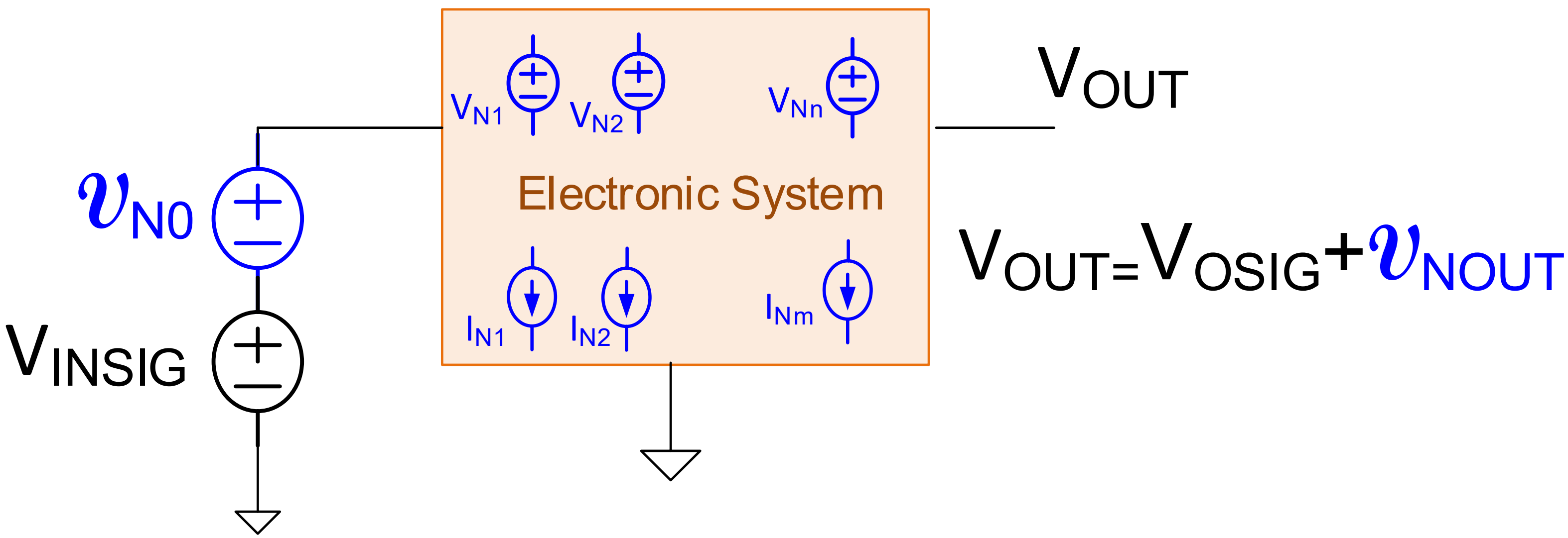
Noise sources in electronic devices are time-domain sources and can be modeled with independent voltage and current sources

Noise sources have a polarity though the statistical characteristics are independent of how the polarity is assigned

Noise is often quantified by the corresponding RMS value of the noise voltage or current at a node or branch in a circuit

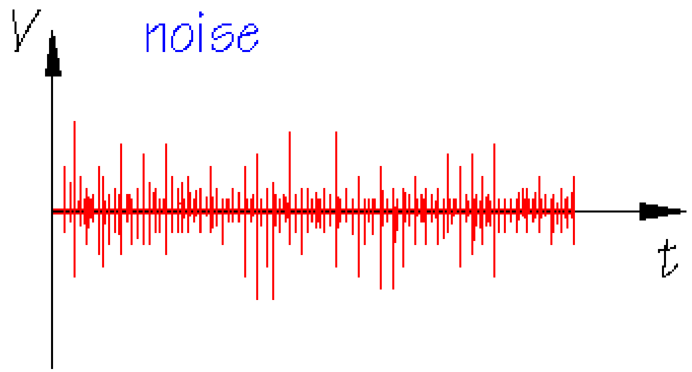
Noise in a System





- Often many noises sources present
- One can be corrupting the input and others are internal to the system
- Noises sources often sufficiently small that superposition can be applied to determine the combined effects of all noise sources on v_{NOUT}

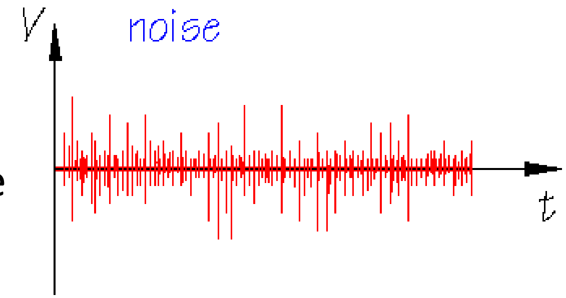
Characterization of a Noise Signal



Noise naturally characterized by its RMS value

$$v_{RMS} = \lim_{T \to \infty} \int_{t_1}^{t_1+T} v_{noise}^2(t)\,dt$$

Noise sources in electronic circuits



Resistors, Transistors, and Diodes all have one or more internal noise sources

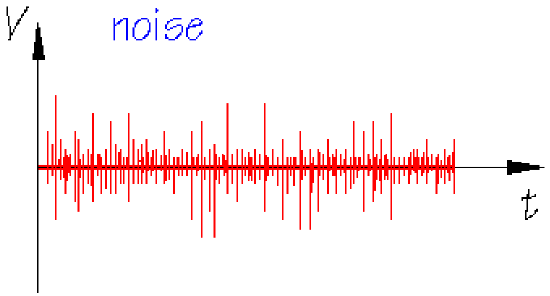
Capacitors and Inductors are noiseless

The presence of noise sources in devices is the only reason that input signals in ideally linear systems are not made arbitrarily small to reduce effects of nonlinearity to arbitrarily small levels

The concept of "Dynamic Range" is used to characterize how small of input signals can be **practically** used in electronic systems

To achieve acceptable linearity in a system, the designer should provide just enough "dynamic range" to satisfy the requirements of an application. Any extra dynamic range will invariably come at the expense of increased design efforts, cost, complexity, and power dissipation

Dynamic Range



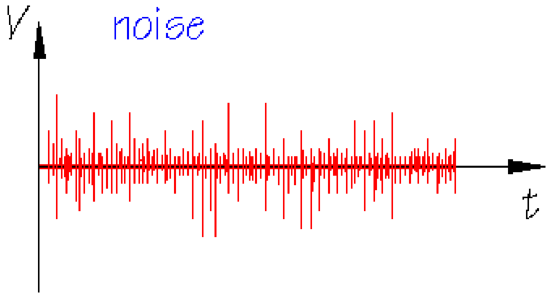
From Wikipedia:

"Dynamic range is the ratio of a specified maximum level of a parameter (e.g. quantity), such as power, current, voltage, or frequency, to the minimum detectable value of that parameter " (from a couple of years ago)

"Dynamic range is the ratio between the largest and smallest values that a certain quantity can assume." (April 30, 2023)

- The maximum level of such a quantity is strongly dependent upon the distortion acceptable in a particular application
- This value may be dependent upon frequency
- The minimum detectable value of a quantity may be dependent upon application
  - Some authors interpret the minimum detectable value to be the RMS value of the quantity when the input signal is zero
- The use of a single value for the DR for a system without knowing the specific applications is of questionable use

Dynamic Range



From Allen and Holberg:

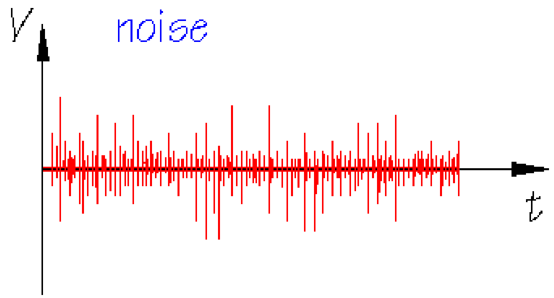
"whereas noise imposes a lower limit on the range of signal amplitudes that can be meaningfully processed by a circuit, linearity often imposes the upper limit. The difference between them is the dynamic range"

From Gregorian and Temes: (in the context of op amp circuits)

"Due to the limited linear range of the op-amp, there is a maximum input signal amplitude, $V_{in,max}$ which the device can handle without generating an excessive amount of nonlinear distortion. …… Due to spurious signals (noise, clock feedthrough, low-level distortion such as crossover distortion, etc.) there is also a minimum input signal $V_{in,min}$ which still does now drown in noise and distortion. The dynamic range of the op amp is then defined as $20\log_{10}\left(V_{in,\max}/V_{in,\min}\right)$ measured in decibels."

Numerous definitions for DR include some "qualitative" terms in the definition making it difficult to identify a universally accepted definition of the DR though the concept is useful

Dynamic Range

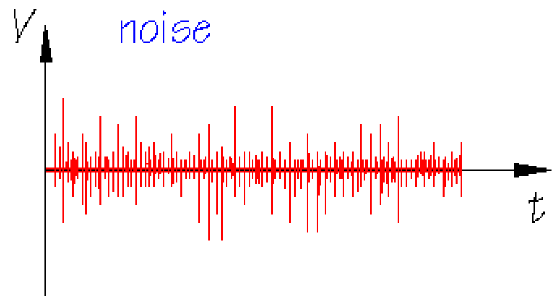


Numerous definitions for DR include some "qualitative" terms in the definition making it difficult to identify a universally accepted definition of the DR though the concept is useful

SNDR is a metric that is rigorously defined that captures some of the DR properties

Though the concept of DR is often not discussed rigorously and though there are various definitions of DR, Dynamic Range should be the primary driver of signal swing, power dissipation, and architecture selection in analog circuit design !

Dynamic Range

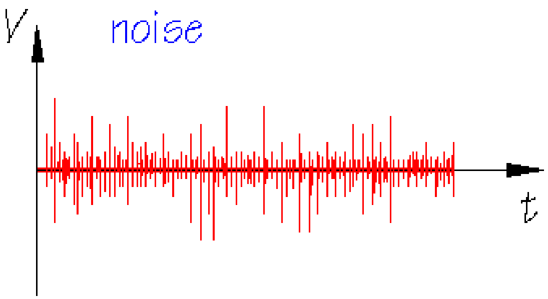


- **Though the concept of DR is often not discussed rigorously and though there are various definitions of DR, Dynamic Range should be the primary driver of signal swing, power dissipation, and architecture selection in analog circuit design !**

- **To achieve acceptable linearity in a system, the designer should provide just enough "dynamic range" to satisfy the requirements of an application. Any extra DR will invariably come at the expense of increased design efforts, cost, complexity, and power dissipation**

So if DR is such an important characteristic, how should it be defined and why is it often ignored?

Dynamic Range



So if DR is such an important characteristic, how should it be defined and why is it often ignored?

Our definition of DR

Assume the maximum acceptable distortion level of an output signal is specified

$$DR = \frac{\text{Output Signal at Maximum Acceptable Distortion}}{\text{Output Noise Over Weighted Freqency Band}}$$

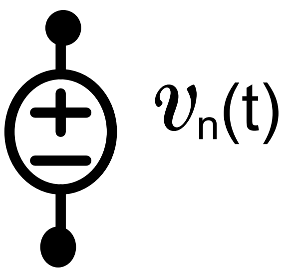
Use of DR in design

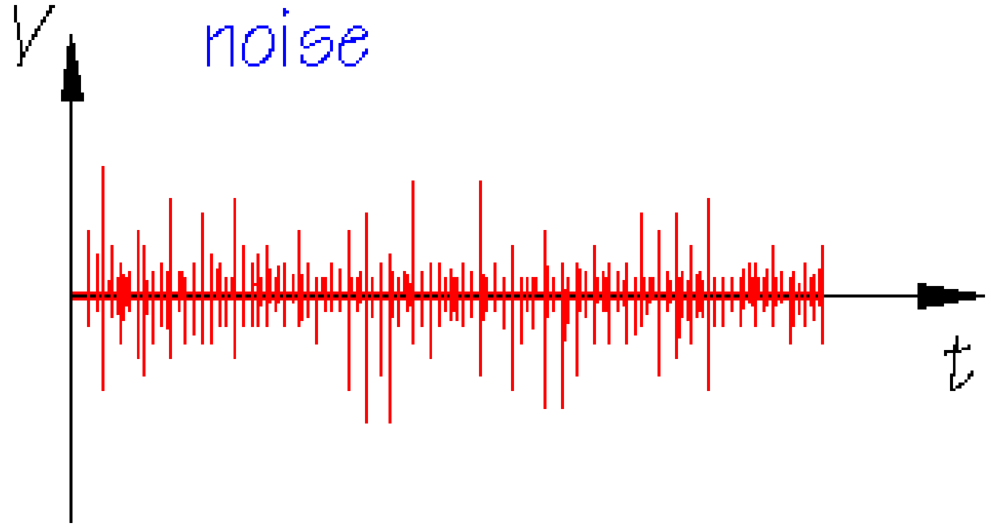
1. Specify the minimum acceptable ratio of the output signal at acceptable distortion to the input noise over the weighted frequency band of interest

2. Calculate DR by above definition

3. Reduce signal swing and/or increase noise (relax noise restrictions) to optimally meet DR requirements

Statistical Characterization of Noise





If $v_n(t_1)$ is a sample of $v_n(t)$, then $v_n(t_1)$ is a random variable

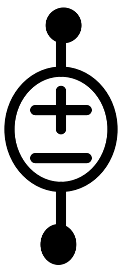
For almost all noise sources, the distribution of $v_n(t_1)$ is zero mean and often Gaussian

For many noise sources, if $v_n(t_1)$ and $v_n(t_2)$ are two distinct samples with $t_1 \neq t_2$, these random variables are identically distributed and uncorrelated (iid)

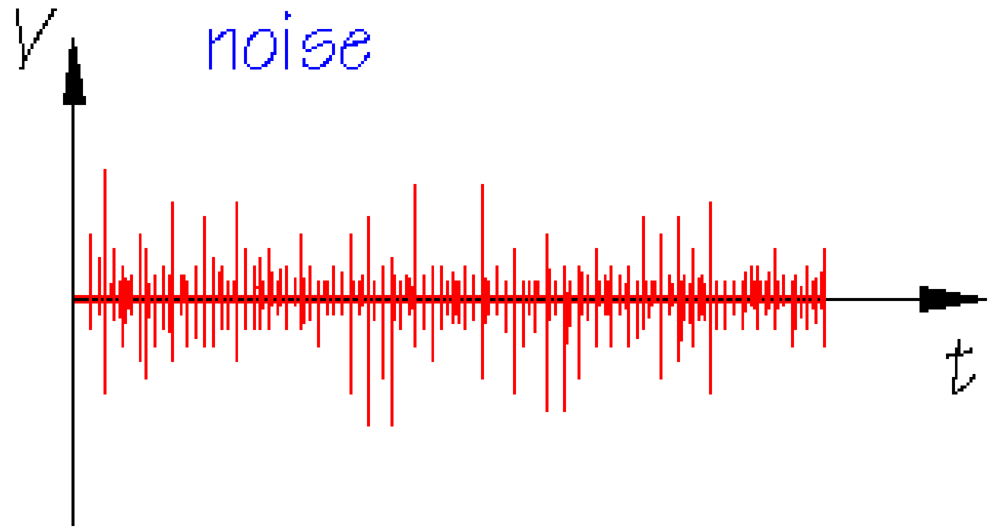
Noise (voltage) is also characterized by how it is distributed throughout the frequency spectrum by its power spectral density, S, or voltage spectral density S_V

Noise is characterized by both S and the amplitude distribution function

Statistical Characterization of Noise



$\mathcal{v}_n(t)$



The RMS noise voltage in the frequency band [f_1,f_2] is given by the expression

$$\mathcal{v}_{RMS}\left(f_1,f_2\right)=\sqrt{\left(\int_{f_1}^{f_2}\mathcal{S}df\right)}$$

$$S=S_V^2$$

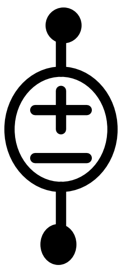
or

$$S=S_I^2$$

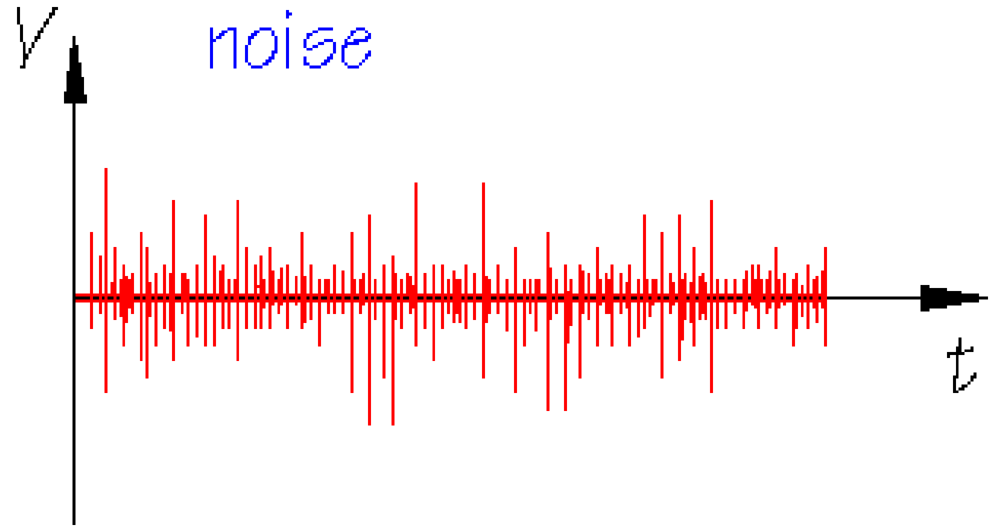
And the total RMS noise voltage is given by the expression

$$\mathcal{v}_{RMS}=\sqrt{\left(\int_{0}^{\infty}\mathcal{S}df\right)}$$

Statistical Characterization of Noise



$v_n(t)$

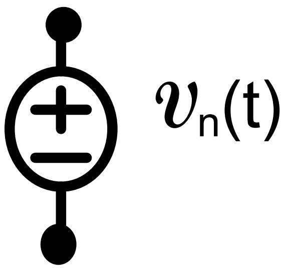


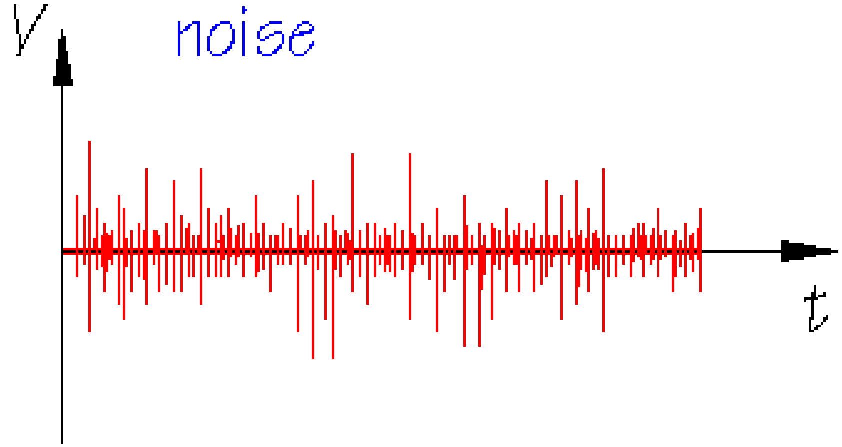
$$v_{RMS} = \sqrt{\left(\int_0^\infty \mathcal{S}\,df\right)} \quad \overset{?}{\Longrightarrow} \quad v_{RMS} = \lim_{T\to\infty} \int_{t_1}^{t_1+T} v^2(t)\,dt$$

Parseval's Theorem

$$\sqrt{\int_{f=0}^{\infty} S df} = \lim_{T\to\infty} \int_{t_1}^{t_1+T} v^2(t)\,dt$$

Statistical Characterization of Noise



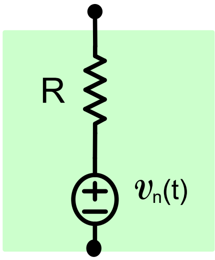


If the spectrum is flat, then the noise is termed "white" noise

White noise can have an amplitude distribution that is Gaussian or non-Gaussian

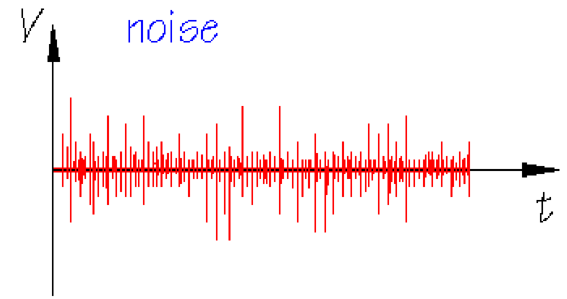
For a resistor, the noise spectrum is white (over a very wide frequency range), the amplitude distribution is Gaussian, and any two distinct samples are iid.

(iid: independent (uncorrelated) and identically distributed)



$$S = 4kTR \quad (V^2\ /\ Hz\ \text{or}\ V^2\ \text{sec})$$

Dynamic Range



Often for audio filters, the DR is defined to be the ratio at the output between that due to a signal at 1% THD to the RMS noise voltage with the actual output noise spectrum multiplied by that of a C-Message bandpass filter

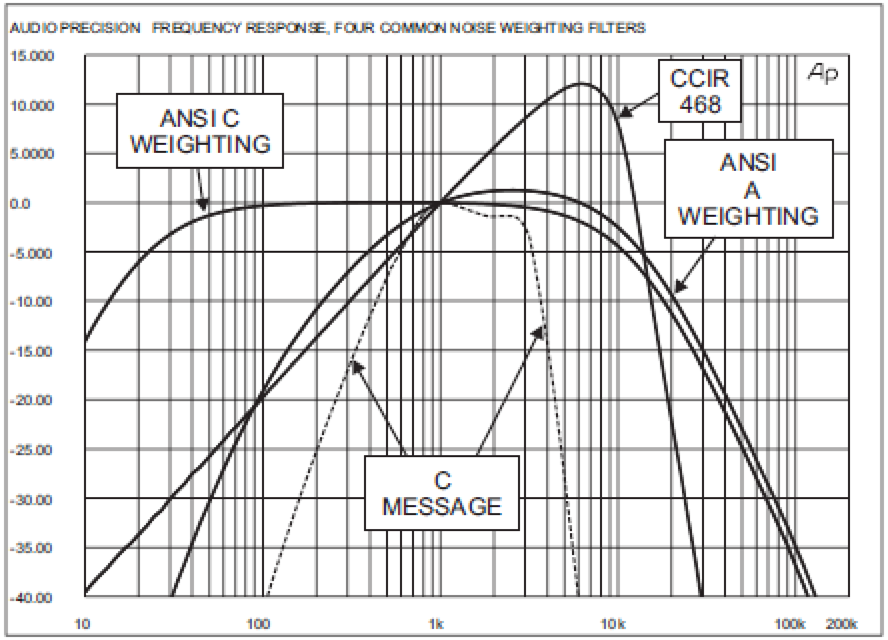
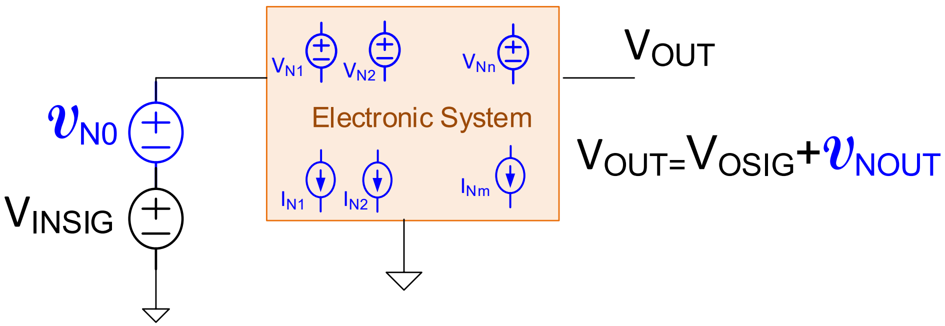


Figure 5. Weighting filter responses, actual measurements. Note that ANSI and C weighting filters are undefined above 20 kHz.

From “Audio Measurement Handbook” by Bob Metzlar

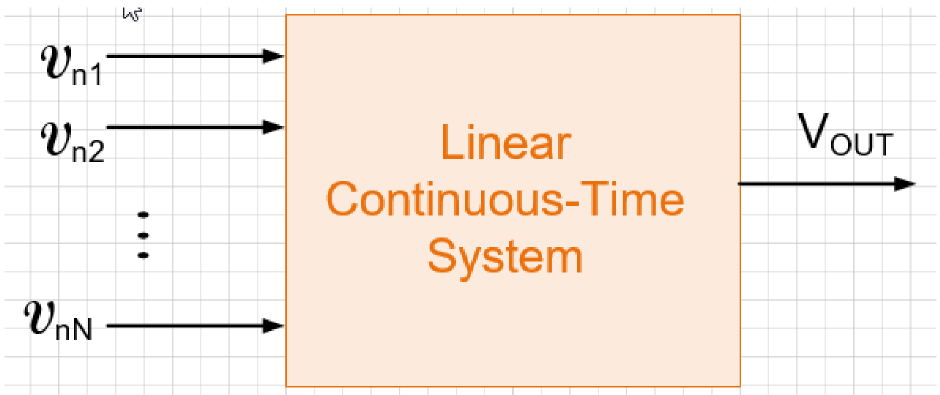
Analysis of Noise in Analog Circuits

Consider an analog circuit with N noise voltage sources (can be easily modified to include both noise voltage and current sources)



N=n+1

The noise sources can be represented by the block diagram shown below

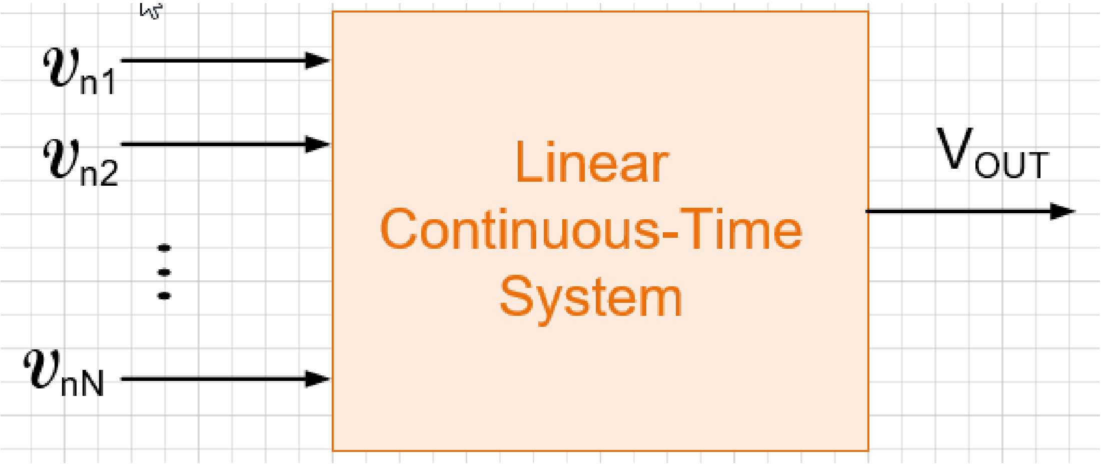


Assume $T_k(s)$ is the transfer function from the kth source to the output

By superposition

$$V_{OUT}(s)=\sum_{i=1}^{N} T_i(s)V_i(s)$$

Analysis of Noise in Analog Circuits



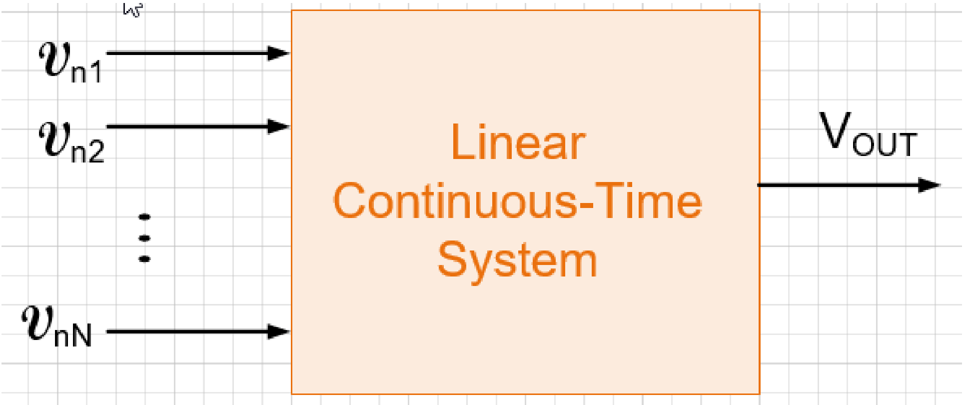
$$V_{OUT}(s) = \sum_{i=1}^{N} T_i(s) V_i(s)$$

If the noise sources are uncorrelated with spectral density S_1, … S_N, the spectral density and the RMS noise voltage at the output are given by the equations:

$$S_{OUT} = \sum_{i=1}^{N} S_i \bullet \left| T_i(j\omega) \right|^2$$

$$v_{OUT\_RMS} = \sqrt{\int_{f=0}^{\infty} S_{OUT} df} = \sqrt{\int_{f=0}^{\infty} \sum_{i=1}^{N} S_i \bullet \left| T_i(j\omega) \right|^2 df}$$

Analysis of Noise in Analog Circuits



$$\mathcal{V}_{OUT\_RMS} = \sqrt{\int_{f=0}^{\infty} S_{OUT} df} = \sqrt{\int_{f=0}^{\infty} \sum_{i=1}^{N} S_i \bullet \left|T_i\left(j\omega\right)\right|^2 df}$$

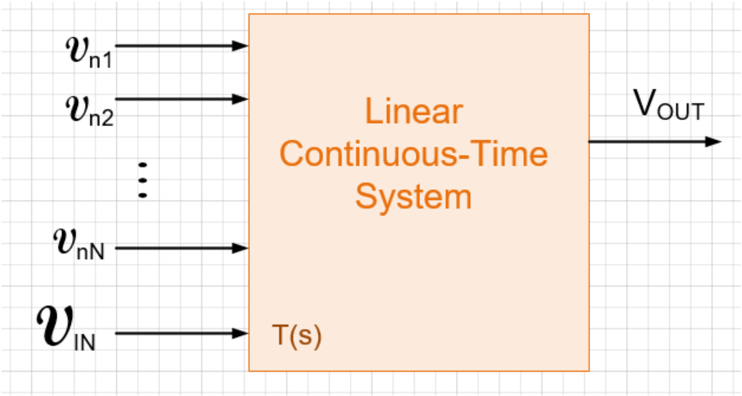
A noise analysis in the frequency domain can be easily run in Spectre to obtain the RMS noise voltage at the output

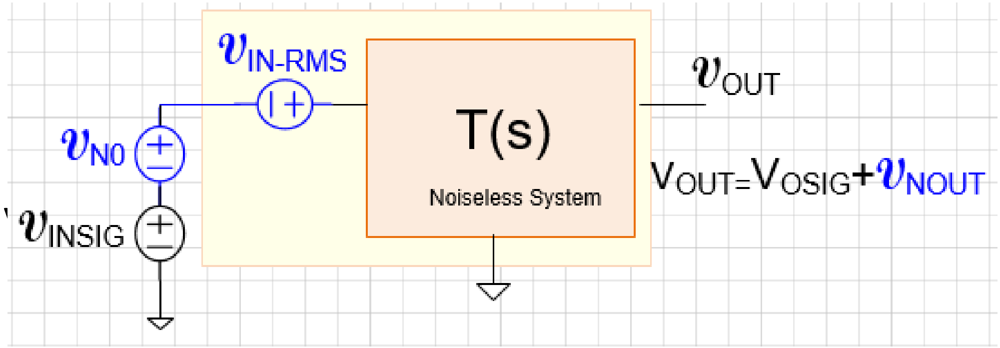
This can be referred back to the input by dividing by the gain from the input to the output to determine the input-referred SNR (see next page)

There is now a time-domain noise analysis capability in Cadence so actual time-domain noise analysis is possible

$\mathcal{V}_{N0}$ usually not part of the analog circuit so it affects system but not the analog circuit

Input-Referred Noise in Analog Circuits





$$v_{OUT\_RMS} = \sqrt{\int_{f=0}^{\infty} S_{OUT} df} = \sqrt{\int_{f=0}^{\infty} \sum_{i=1}^{N} S_i \bullet \left|T_i(j\omega)\right|^2 df}$$

Let T(s) be the transfer function from the input to the output. (usually T(s) will be distinct from each of the noise transfer functions).

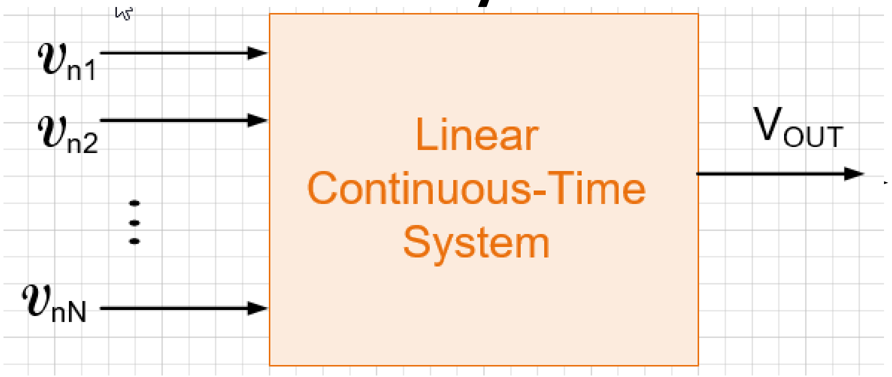
The input-referred noise spectral density is given by the expression

$$S_{IN} = \frac{S_{OUT}}{\left|T(j\omega)^2\right|}$$

The input-referred RMS voltage is thus given by

$$v_{IN\_RMS} = \sqrt{\int_{f=0}^{\infty} \frac{S_{OUT}}{\left|T(j\omega)\right|^2} df} = \sqrt{\int_{f=0}^{\infty} \sum_{i=1}^{N} S_i \bullet \frac{\left|T_i(j\omega)\right|^2}{\left|T(j\omega)\right|^2} df}$$

Relationship between frequency domain and time domain noise analysis



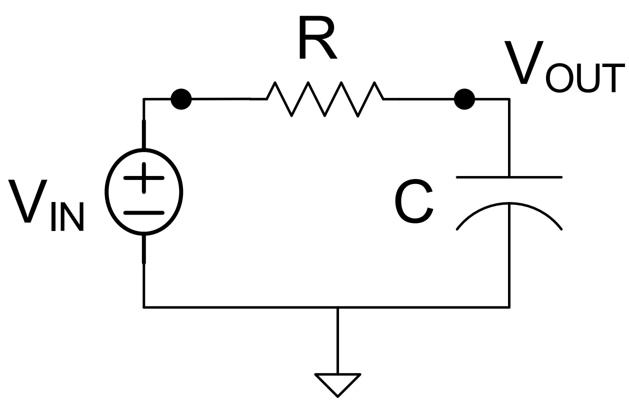
$$v_{OUT\_RMS} = \sqrt{\int_{f=0}^{\infty} S_{OUT} df} = \sqrt{\int_{f=0}^{\infty} \sum_{i=1}^{N} S_i \bullet \left|T_i(j\omega)\right|^2 df}$$

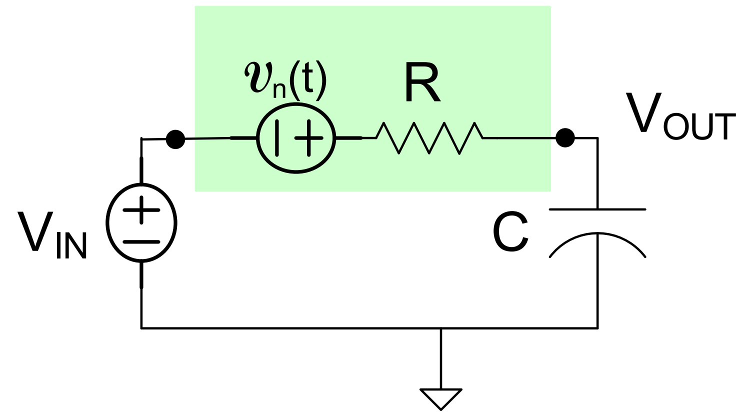
$$V_{RMS\_OUT} = E\left(\sqrt{\lim_{T\to\infty}\left(\frac{1}{T}\int_0^T V_{OUT}^2(t)dt\right)}\right) \simeq \sqrt{\lim_{T\to\infty}\left(\frac{1}{T}\int_0^T V_{OUT}^2(t)dt\right)}$$

Parseval's Theorem

$$V_{RMS\_OUT} = v_{OUT\_RMS}$$

Example: Noise in First-Order RC Network



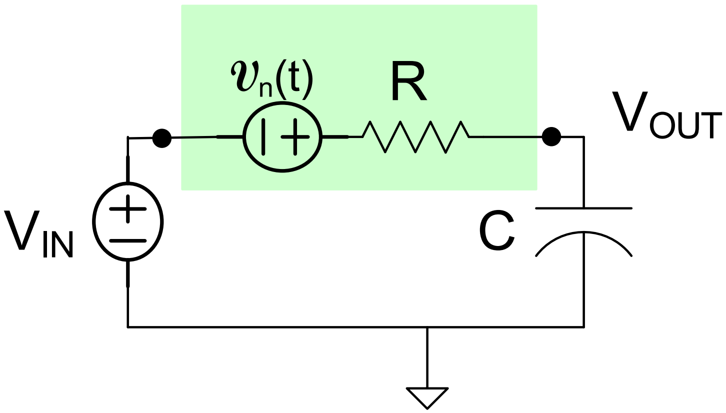


$$T(s) = \frac{1}{1+RCs}$$

$$S_{VOUT} = 4kTR\left(\frac{1}{1+(RC\omega)^2}\right)$$

$$v_{n_{RMS}} = \sqrt{\int_{f=0}^{\infty} S_{V_{OUT}}\, df} = \sqrt{\int_{f=0}^{\infty} \frac{4kTR}{1+\omega^2R^2C^2}\, df}$$

Example: Noise in First-Order RC Network



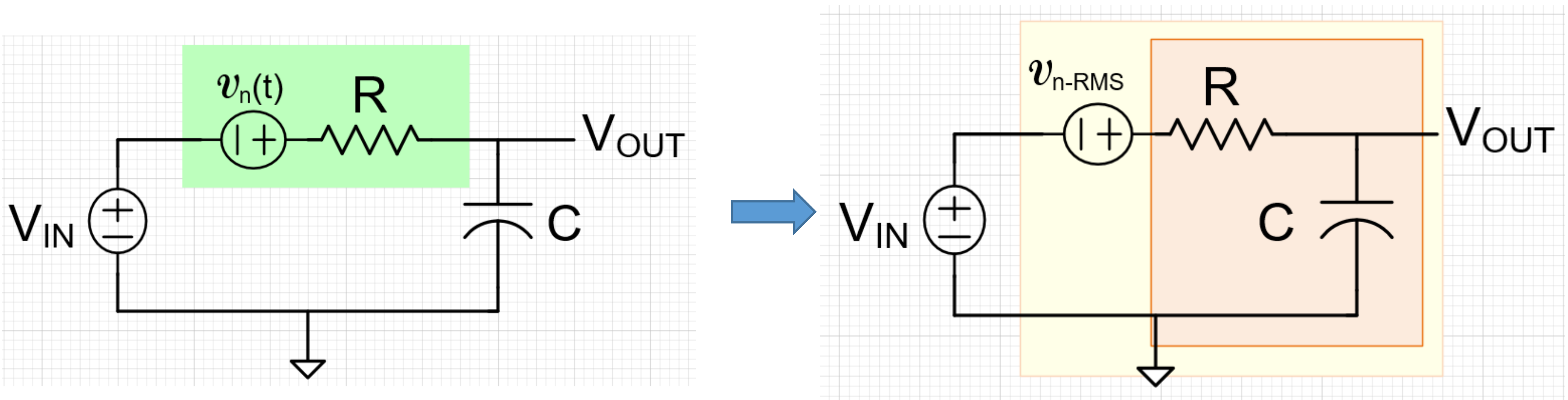
$$v_{n_{RMS}} = \sqrt{\int_{f=0}^{\infty} S_{V_{OUT}}\,df} = \sqrt{\int_{f=0}^{\infty} \frac{4kTR}{1+\omega^2 R^2 C^2}\,df}$$

From a standard change of variable with a trig identity, it follows that

$$v_{n_{RMS}} = \sqrt{\int_{f=0}^{\infty} S_{V_{OUT}}\,df} = \sqrt{\frac{kT}{C}}$$

- Note the continuous-time noise voltage has an RMS value that is independent of R
- The noise contributed by the resistor is dependent only upon the capacitor value C
- This is often referred to at kT/C noise and it can be decreased at a given T only by increasing C

Example: Noise in First-Order RC Network

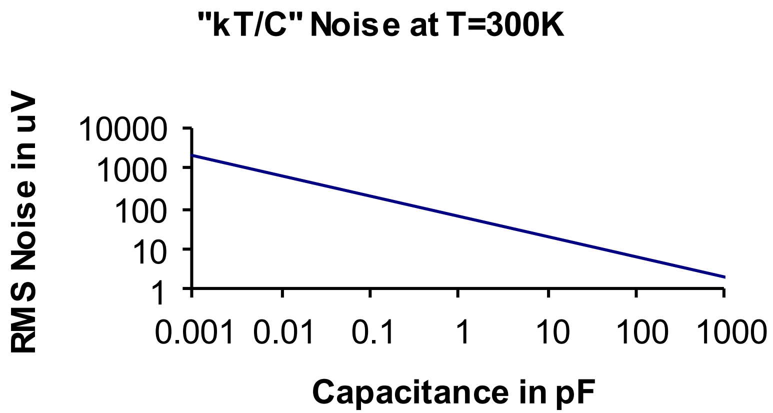


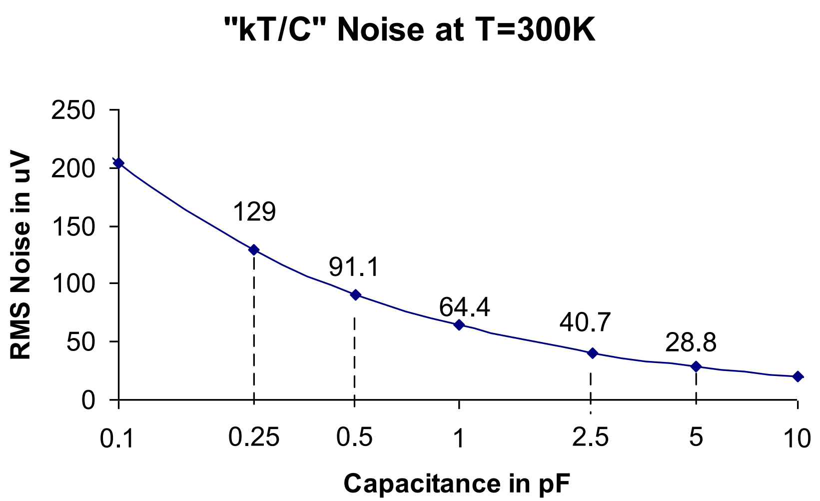
From a standard change of variable with a trig identity, it follows that

$$v_{n_{RMS}} = \sqrt{\int_{f=0}^{\infty} S_{V_{OUT}} \mathrm{df}} = \sqrt{\frac{kT}{C}}$$

- Note the continuous-time noise voltage has an RMS value that is independent of R
- The noise contributed by the resistor is dependent only upon the capacitor value C
- This is often referred to at kT/C noise and it can be decreased at a given T only by increasing C

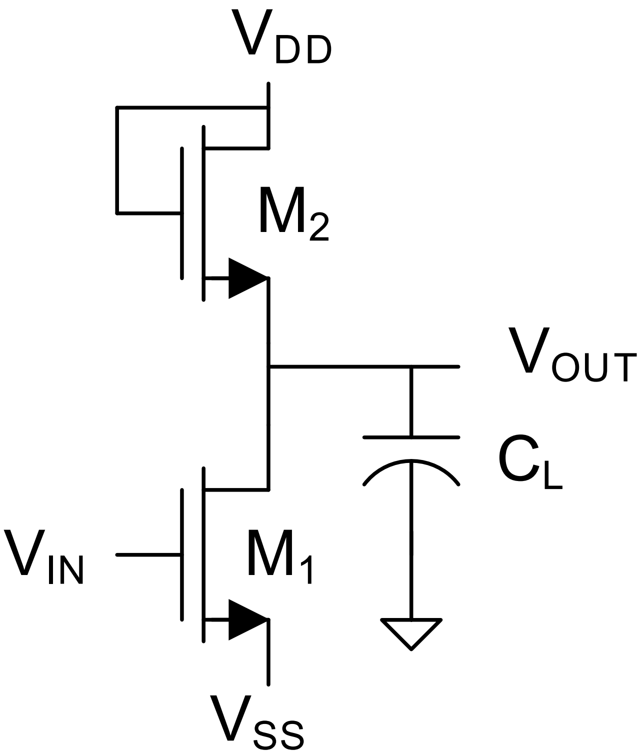
Noise Associated with Capacitors





Example: Noise in Simple Common-Source Amplifier

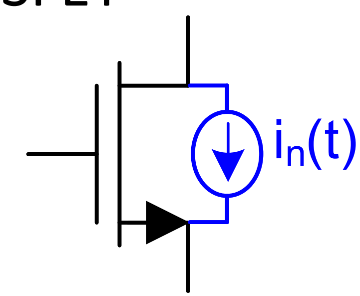
Determine the RMS noise voltage in the output in the frequency band from dc to 1MHz (neglect C_L)



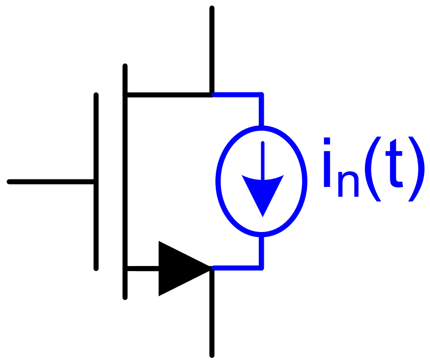
$$A_V = \frac{v_{OUT}}{v_{IN}} \simeq -\frac{g_{m1}}{g_{m2}}$$

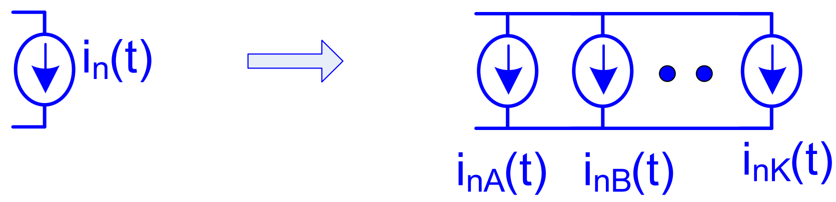
Will determine the noise voltage at the output and the corresponding input-referred noise voltage

Need noise model for MOSFET



Example: Noise Model for MOSFET



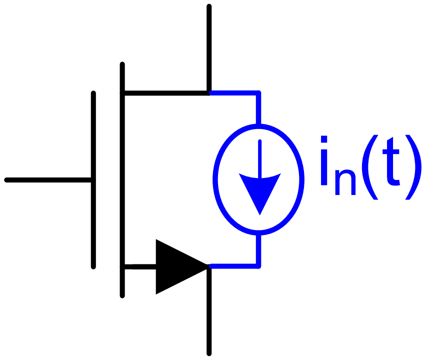


- $i_n(t,T)$ is comprised of several uncorrelated noise sources
- Since uncorrelative, they can be added in the RMS sense
- Spectral densities thus are simply additive
- Two most important are usually due to thermal noise and 1/f noise

In this lecture, will consider only thermal noise in MOSFET

Example: Noise Model for MOSFET

Thermal Noise in MOSFET



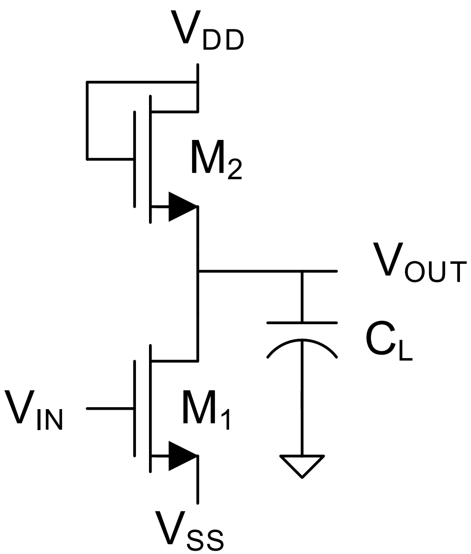
$$S_{I\_Thermal} = \begin{cases} \dfrac{4kT}{R_{FET}} & \textit{ohmic region} \\ \dfrac{8kTg_m}{3} & \textit{saturation region} \end{cases}$$

- Power Spectral Density is Flat
- White Noise
- Highly T dependent

$8kT/3 = 1.1\text{x}10^{-20}$ V•A•sec at T=300K

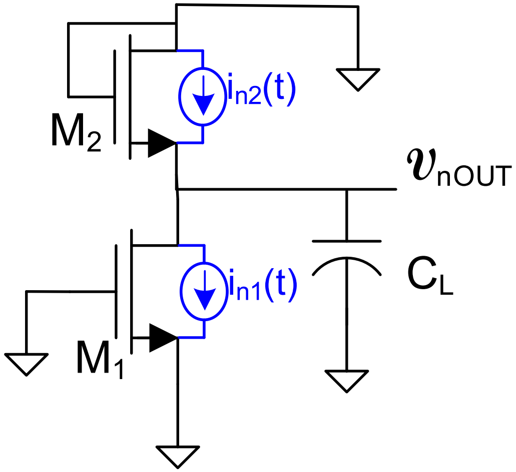
Example: Noise in Simple Common-Source Amplifier

Determine the RMS noise voltage in the output in the frequency band from dc to 1MHz (neglect C_L)



$$A_V = \frac{v_{OUT}}{v_{IN}} \simeq -\frac{g_{m1}}{g_{m2}}$$

Small-signal noise model for thermal noise

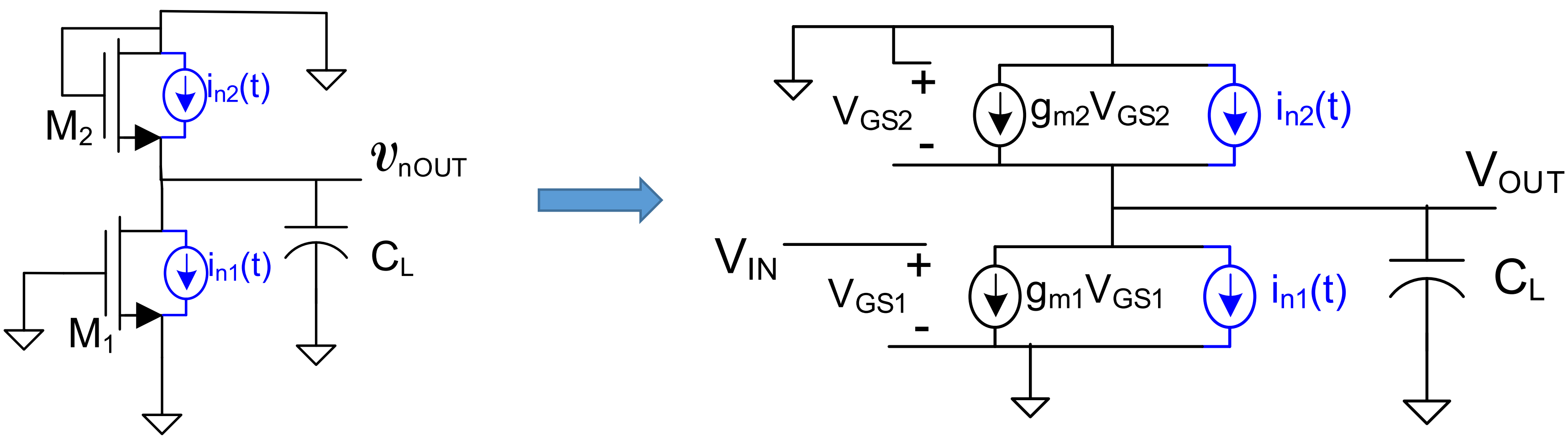


$$S_{n1} = \frac{8kTg_{m1}}{3}$$

$$S_{n2} = \frac{8kTg_{m2}}{3}$$

Example: Noise in Simple Common-Source Amplifier

Determine the RMS noise voltage in the output in the frequency band from dc to 1MHz (neglect C_L)



From KCL:

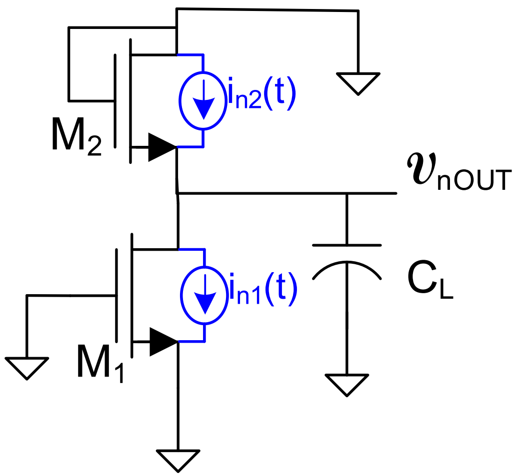
$$sC_L V_{OUT} + i_{n1} + g_{m1} V_{IN} = i_{n2} + g_{m2}(-V_{OUT})$$

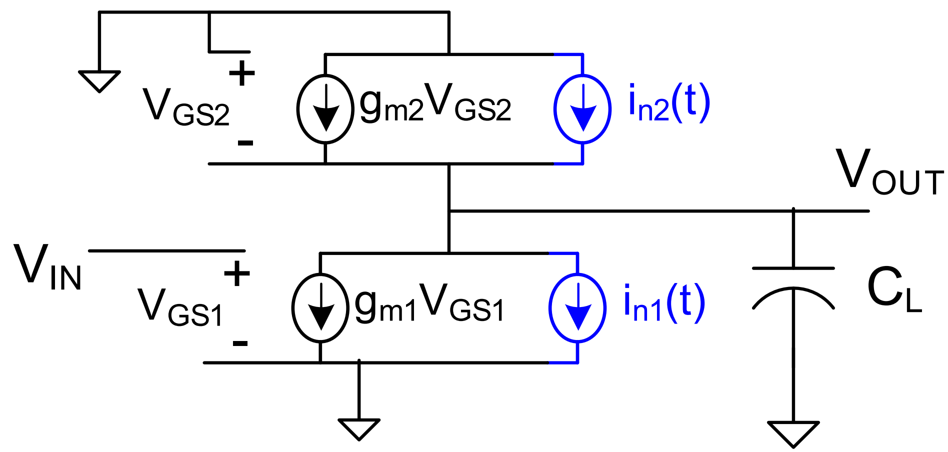
It thus follows that:

$$V_{OUT} = -\frac{g_{m1}}{sC_L + g_{m2}} V_{IN} - \frac{1}{sC_L + g_{m2}} i_{n1} + \frac{1}{sC_L + g_{m2}} i_{n2}$$

Example: Noise in Simple Common-Source Amplifier

Determine the RMS noise voltage in the output in the frequency band from dc to 1MHz (neglect C_L)





$$V_{OUT} = -\frac{g_{m1}}{sC_L + g_{m2}} V_{IN} - \frac{1}{sC_L + g_{m2}} i_{n1} + \frac{1}{sC_L + g_{m2}} i_{n2}$$

$$S_{OUT} = \sum_{i=1}^{N} S_i \bullet \left| T_i(j\omega) \right|^2$$

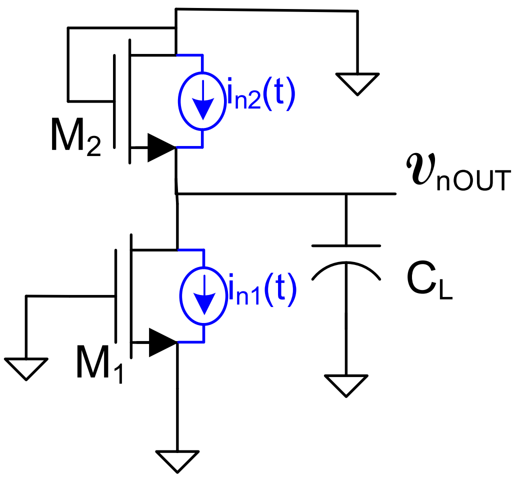
$$S_{n1} = \frac{8kTg_{m1}}{3}$$

$$S_{n2} = \frac{8kTg_{m2}}{3}$$

$$S_{OUT} = S_1 \frac{1}{\omega^2 C_L^2 + g_{m2}^2} + S_2 \frac{1}{\omega^2 C_L^2 + g_{m2}^2} = \frac{8kT}{3}\left[\frac{g_{m1}}{\omega^2 C_L^2 + g_{m2}^2} + \frac{g_{m2}}{\omega^2 C_L^2 + g_{m2}^2} \right]$$

Example: Noise in Simple Common-Source Amplifier

Determine the RMS noise voltage in the output in the frequency band from dc to 1MHz (neglect C_L)



$$S_{OUT} = \frac{8kT}{3}\left[\frac{g_{m1}}{\omega^2 C_L^2 + g_{m2}^2} + \frac{g_{m2}}{\omega^2 C_L^2 + g_{m2}^2}\right]$$

$$v_{OUT\_RMS} = \sqrt{\int_{f=0}^{\infty} S_{OUT}\,df} = \sqrt{\int_{f=0}^{\infty} \sum_{i=1}^{N} S_i \bullet \left|T_i(j\omega)\right|^2 df}$$

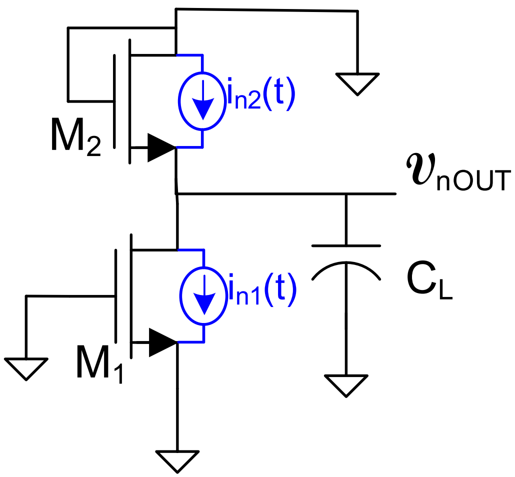
$$v_{OUT\_RMS} = \sqrt{\int_{f=0}^{10^6} \frac{8kT}{3}\left[\frac{g_{m1}}{\omega^2 C_L^2 + g_{m2}^2} + \frac{g_{m2}}{\omega^2 C_L^2 + g_{m2}^2}\right] df}$$

Neglecting C_L, we obtain

$$v_{OUT\_RMS} = \sqrt{\int_{f=0}^{10^6} \frac{8kT}{3}\left[\frac{g_{m1}}{g_{m2}^2} + \frac{g_{m2}}{g_{m2}^2}\right] df}$$

Example: Noise in Simple Common-Source Amplifier

Determine the RMS noise voltage in the output in the frequency band from dc to 1MHz (neglect C_L)



$$v_{OUT\_RMS} = \sqrt{\int_{f=0}^{10^6} \frac{8kT}{3}\left[\frac{g_{m1}}{g_{m2}^2} + \frac{g_{m2}}{g_{m2}^2}\right] df}$$

But g_{m1} and g_{m2} are independent of f

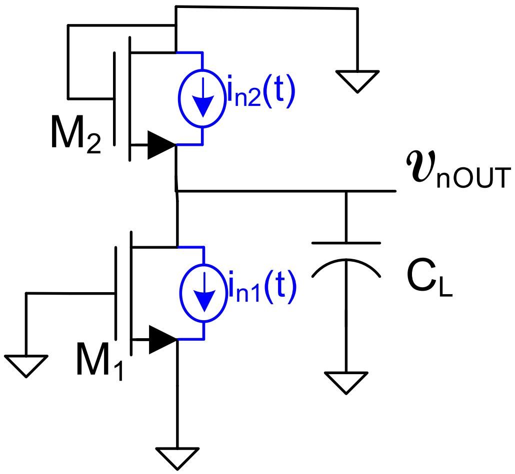
$$v_{OUT\_RMS} = \sqrt{\frac{8kT}{3}\left[\frac{g_{m1}}{g_{m2}^2} + \frac{g_{m2}}{g_{m2}^2}\right] 10^6}$$

$$v_{OUT\_RMS} = \sqrt{\frac{8kT}{3}\left[\frac{1}{2I_{DQ}}\frac{V_{EB2}^2}{V_{EB1}} + \frac{1}{2I_{DQ}}V_{EB2}\right]10^6} = \sqrt{\frac{1}{I_{DQ}}}\sqrt{\frac{4kT}{3}10^6\left(\frac{V_{EB2}^2}{V_{EB1}} + V_{EB2}\right)}$$

Note that noise voltage decreases with sqrt I_{DQ}

Example: Noise in Simple Common-Source Amplifier

Determine the RMS noise voltage in the output in the frequency band from dc to 1MHz (neglect C_L)



$$v_{OUT\_RMS} = \sqrt{\frac{1}{I_{DQ}}}\sqrt{\frac{4kT}{3}10^6\left(\frac{V_{EB2}^2}{V_{EB1}} + V_{EB2}\right)}$$

Consider Case 1: $V_{EB1}=V_{EB2}=0.5V$, $T=300K$, $I_{DQ}=1uA$

$$v_{OUT\_RMS} = 74\mu V$$

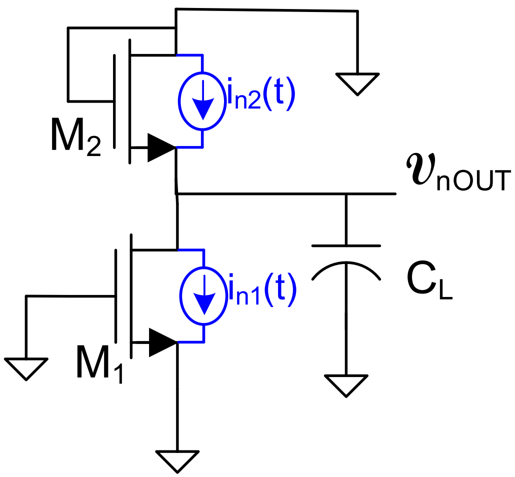
Consider Case 2: $V_{EB1}=V_{EB2}=0.5V$, $T=300K$, $I_{DQ}=1mA$

$$v_{OUT\_RMS} = 2.3\mu V$$

- Note noise strongly dependent upon biasing conditions !
- Including C_L would not have complicated analysis very much and for large C_L, it will actually attenuate noise

Example: Noise in Simple Common-Source Amplifier

Determine the RMS noise voltage in the output in the frequency band from dc to 1MHz (neglect C_L)



Consider Case 1: $V_{EB1}=V_{EB2}=0.5V$, T=300K, $I_{DQ}=1uA$

$$v_{OUT\_RMS} = 74\mu V$$

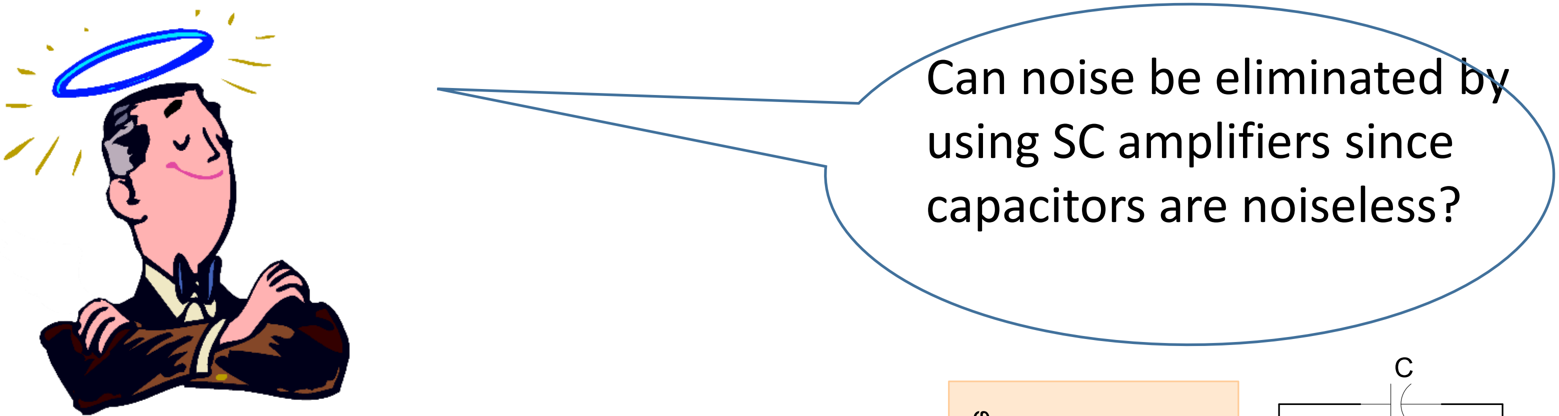
SNR can be easily calculated at a given signal level

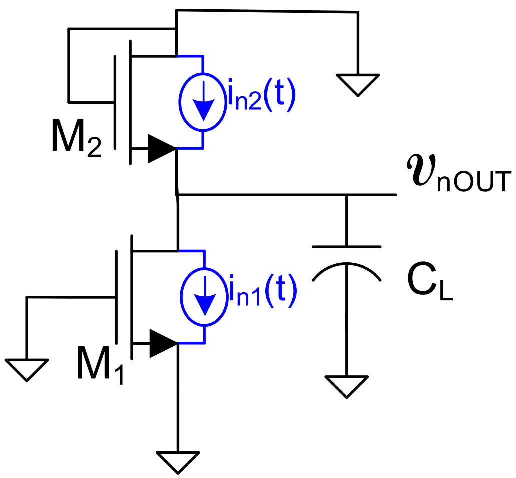
Dynamic Range can be calculated at an acceptable distortion level

1/f noise may increase output noise voltage

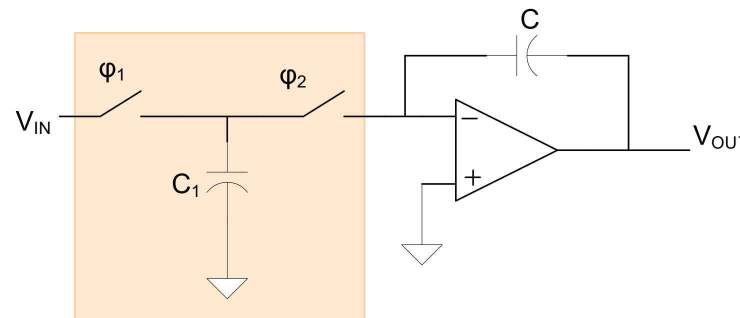
Input referred noise voltage can be calculated by dividing by gain

C_L effects can easily be added (see equation above)

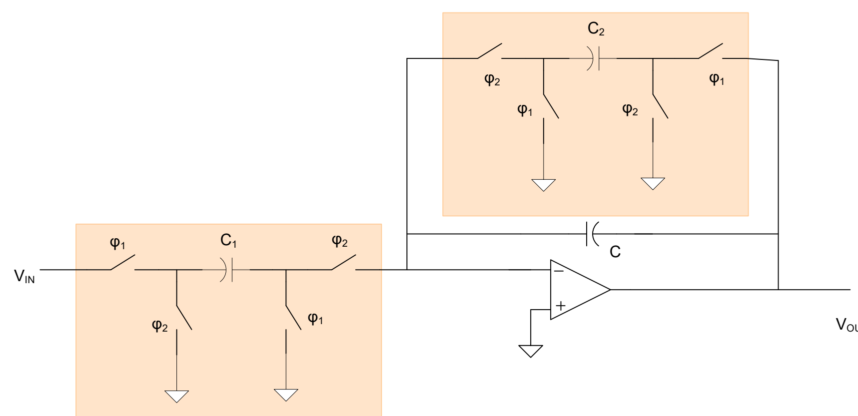




g_m-C Lowpass Filter



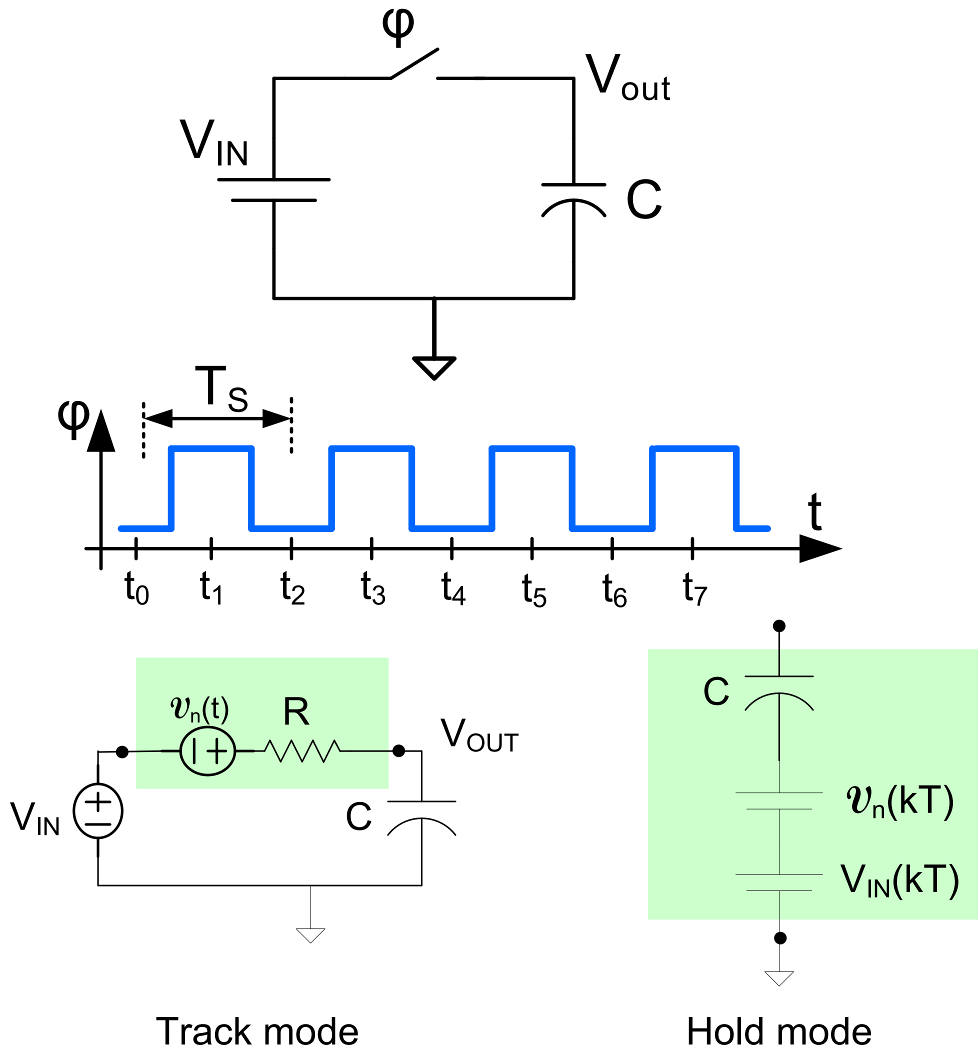
SC Integrator



SC Lowpass Filter

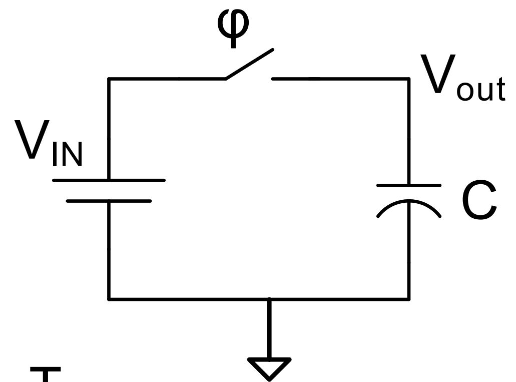
Both SC filter structures have a SC sampler

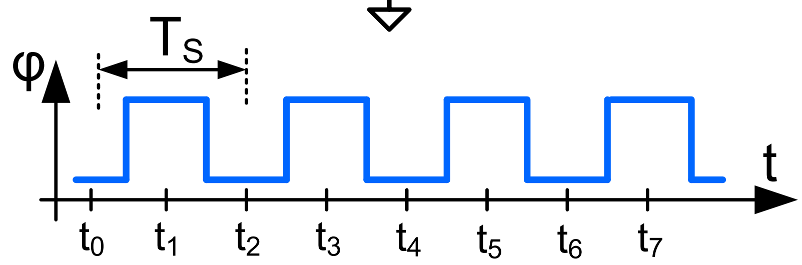
Example: Switched Capacitor Sampler

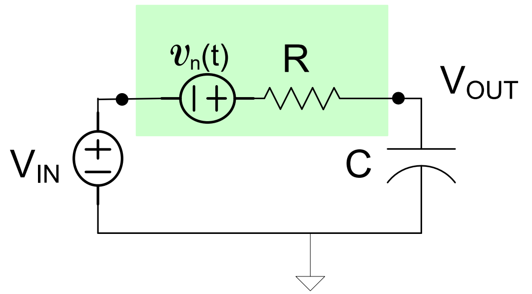


Resistor present during track mode !

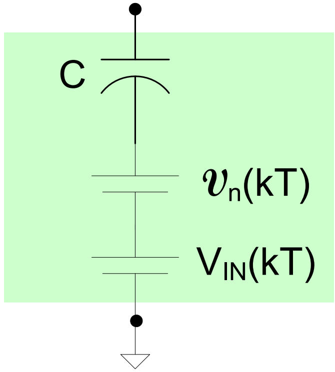
Example: Switched Capacitor Sampler





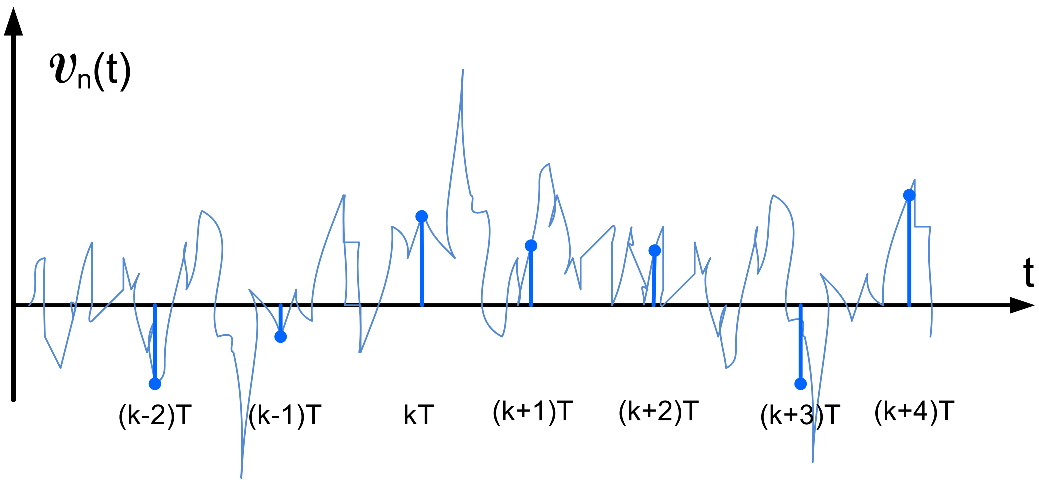


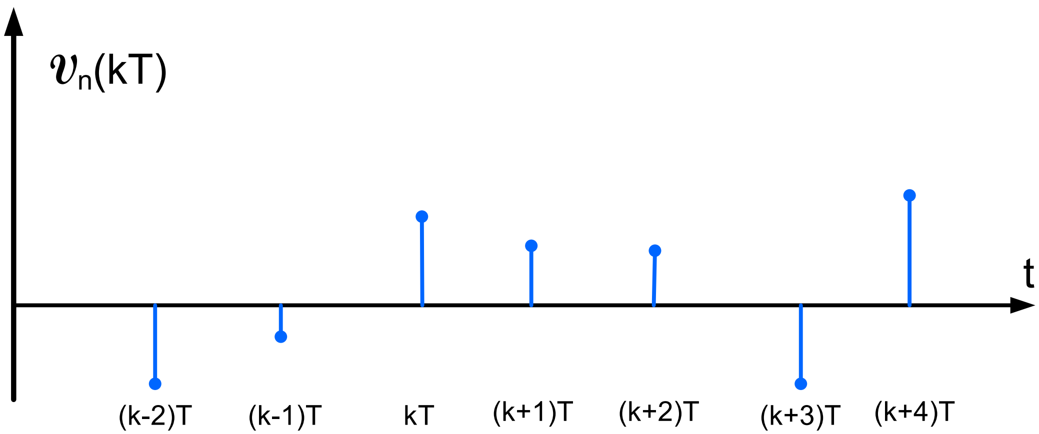
Track mode



Hold mode

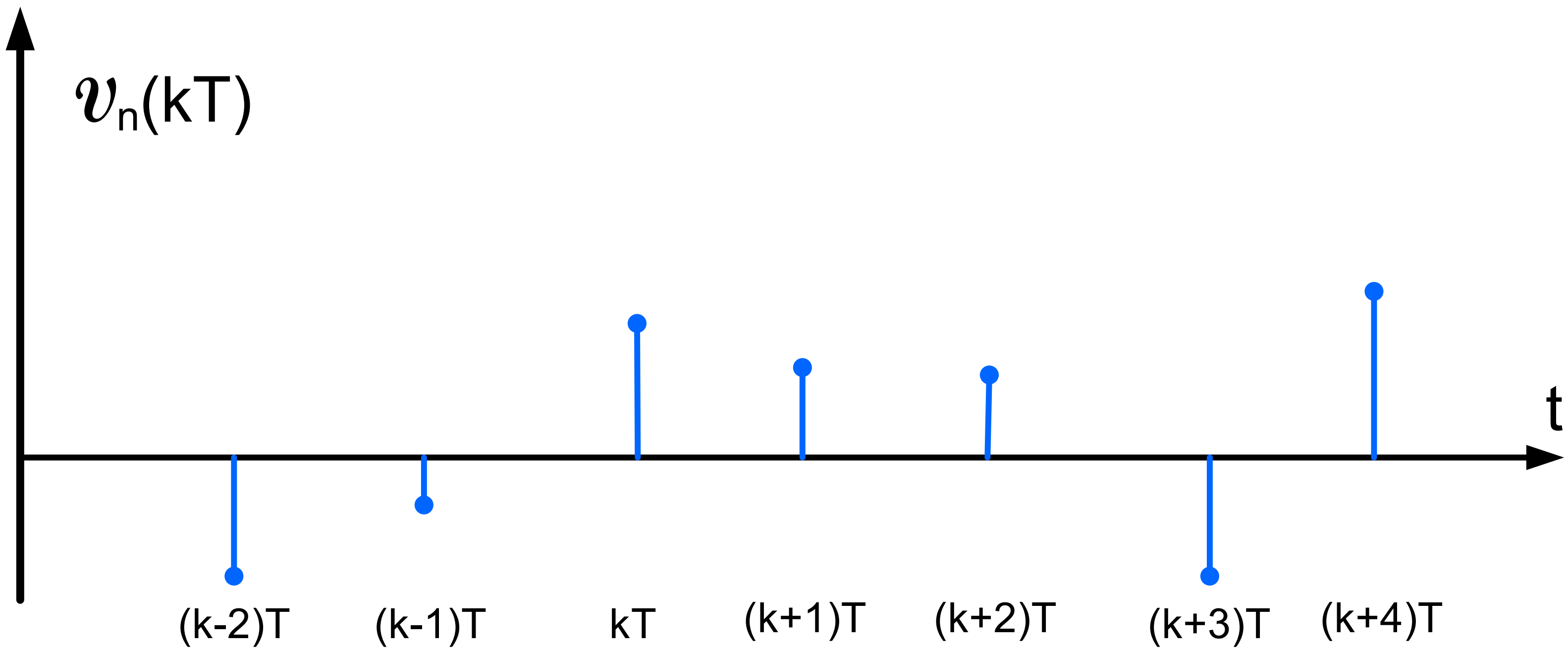
Example: Switched Capacitor Sampler





$v_n(kT)$ is a discrete-time sequence obtained by sampling a continuous-time noise waveform

Characterization of a noise sequence

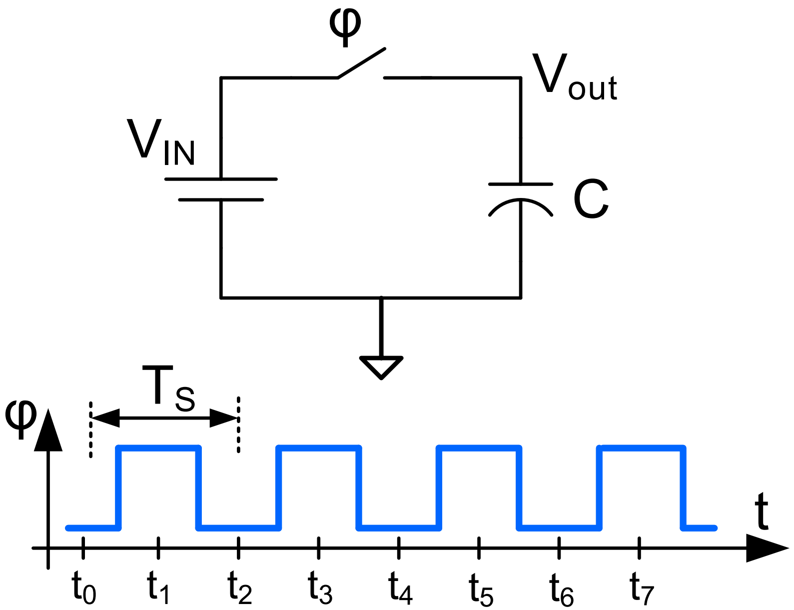


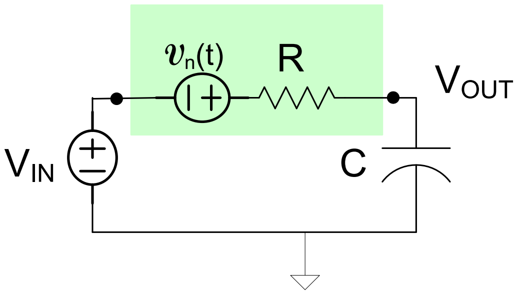
$$\hat{v}_{RMS} = E\left(\sqrt{\lim_{N\to\infty}\left(\frac{1}{N}\sum_{k=1}^{N} v^2(kT)\right)}\right) \underset{N\, large}{\simeq} \sqrt{\frac{1}{N}\sum_{k=1}^{N} v^2(kT)}$$

Theorem If $\mathcal{V}(t)$ is a continuous-time zero-mean noise source and <$\mathcal{V}$(kT)> is a sampled version of $\mathcal{V}(t)$ sampled at times T, 2T, …. then the RMS value of the continuous-time waveform is the same as that of the sampled version of the waveform. This can be expressed as $\mathcal{V}_{\text{RMS}} = \hat{\mathcal{V}}_{\text{RMS}}$

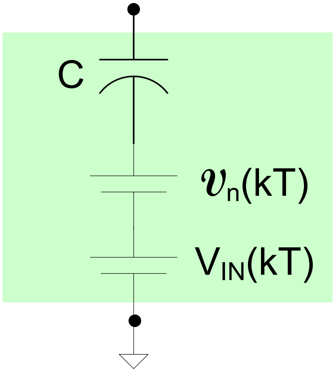
Theorem If $\mathcal{V}(t)$ is a continuous-time zero-mean noise signal and <$\mathcal{V}$(kT)> is a sampled version of $\mathcal{V}(t)$ sampled at times T, 2T, …. then the standard deviation of the random variable $\mathcal{V}(kT)$, denoted as $\sigma_{\hat{V}}$ satisfies the expression $\sigma_{\hat{V}} = \mathcal{V}_{\text{RMS}} = \hat{\mathcal{V}}_{\text{RMS}}$

Example: Switched Capacitor Sampler





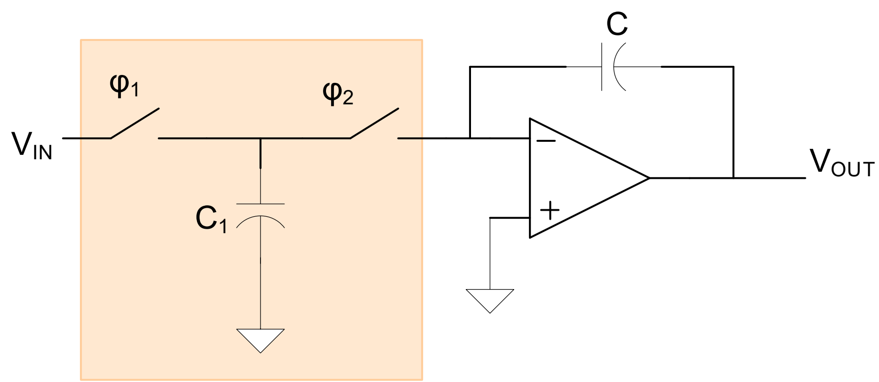
Track mode



Hold mode

$$v_{n_{RMS}} = \sqrt{\frac{kT}{C}}$$

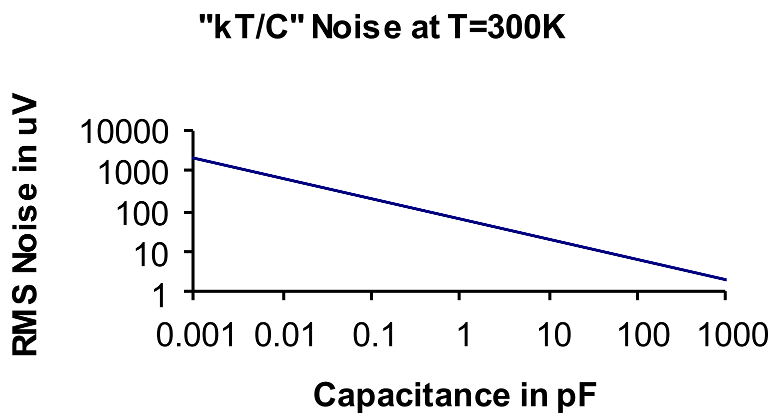
Can noise be eliminate by using SC amplifiers since capacitors are noiseless?

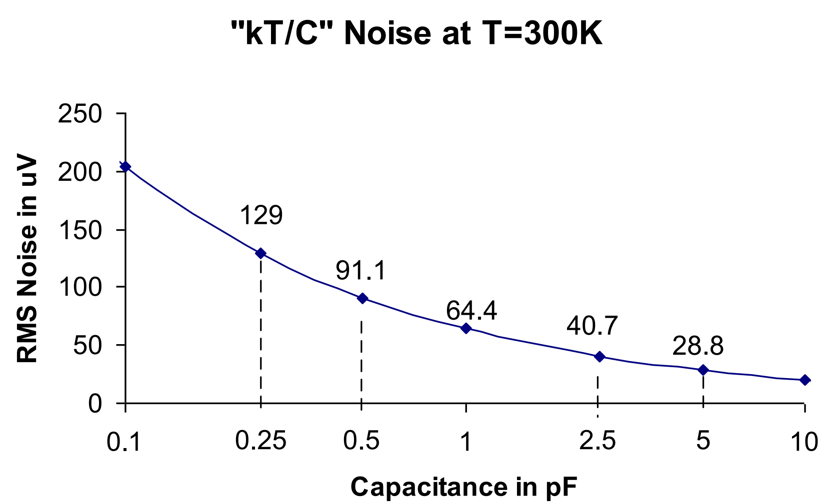


$$v_{n_{RMS}} = \sqrt{\frac{kT}{C_1}}$$

Even though capacitors are noiseless, the RMS noise voltage due to the switch will be dependent upon C independent of the size of the switch !!

Noise Associated with Capacitors





Stay Safe and Stay Healthy !

End of Lecture 42