Noise Analysis in
Switched-Capacitor Circuits

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Abstract

Switched-capacitor (SC) circuits are ubiquitous in CMOS mixed-signal ICs. The most fundamental performance limitation in these circuits stems from the thermal noise introduced by MOSFET switches and active amplifier circuitry. This tutorial reviews hand analysis techniques that allow the designer to predict the noise performance of switched-capacitor circuits at various levels of complexity. The material presented in this course focuses on practical examples ranging from basic passive and active track-and-hold circuits, integrators and SC delta-sigma modulators. Simulation examples are included to complement and verify the theoretical treatment.
Outline

- Motivation and Overview
- Preliminaries
- Circuit Examples
  - Elementary track-and-hold circuit
  - Charge redistribution stage
  - Delta-sigma modulator
Example: Delta-Sigma Modulator

Example: Pipeline ADC

Example: Pipeline ADC

Example: Sensor Interface

Charge Redistribution Stages

Integrator

Gain stage

Non-overlapping clock signals

$T_s = 1/f_s$
Elementary T/H Circuit

Elementary Track and Hold Circuit
Outline

• Motivation and Overview

• Preliminaries

• Circuit Examples
  – Elementary track-and-hold circuit
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  – Delta-sigma modulator
Types of Noise

- "Man made noise"
  - Supply coupling
  - Substrate coupling

- "Electronic noise"
  - Technology related
    - Flicker noise caused by lattice defects
  - Fundamental
    - Thermal noise caused by random motion of carriers
    - Focus of this short course
Significance of Thermal Noise

- The "fidelity" of electronic systems is often determined by their SNR
  - Examples: Audio systems, wireless transceivers, sensor interfaces
- Electronic noise directly trades with power dissipation and speed
- Electronic noise is a major concern in modern technologies with reduced $V_{DD}$

$$\text{SNR} \propto \frac{V_{\text{signal}}^2}{V_{\text{noise}}^2} \propto \frac{(\alpha V_{DD})^2}{V_{\text{noise}}^2}$$

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Physical Resistor

• "Thermal Noise" or "Johnson Noise"

• Can model random current component using a noise current source $i_n(t)$
Statistical Model

- The power spectral density (PSD) indicates how the power of a signal is spread across frequency.
- For thermal noise, the power is spread uniformly up to very high frequencies (about 10% drop at 2 THz).

\[ \text{PSD}(f) \]

\[ f \]
Thermal Noise Power

• Nyquist showed that

\[
\text{PSD}(f) = 4kT
\]

• The total average noise power of a resistor in a certain frequency band is therefore

\[
P_n = \int_{f_1}^{f_2} 4kT \cdot df = 4kT \cdot (f_2 - f_1) = 4kT \cdot \Delta f
\]
Equivalent Noise Generators

- We can model the noise using an equivalent voltage or current generator.

\[
\overline{v_n^2} = P_n \cdot R = 4kT \cdot R \cdot \Delta f
\]

\[
\overline{i_n^2} = \frac{P_n}{R} = 4kT \cdot \frac{1}{R} \cdot \Delta f
\]

For \( R = 1k\Omega \):

\[
\overline{v_n^2} = 16 \cdot 10^{-18} \frac{V^2}{Hz}
\]

\[
\sqrt{\overline{v_n^2}} = 4nV \frac{\sqrt{Hz}}{\sqrt{Hz}}
\]

\[
\sqrt{\overline{i_n^2}} = 4pA \frac{\sqrt{Hz}}{\sqrt{Hz}}
\]
MOSFET Thermal Noise

• The noise of a MOSFET operating in the triode region is approximately equal to that of a resistor.

• In the saturation region, the thermal noise can be modeled using a drain current source with power spectral density

\[
\bar{i_d^2} = 4kT \cdot \gamma \cdot g_m \cdot \Delta f
\]

• For an ideal MOSFET, we have \( \gamma = 2/3 \)
\( \gamma \) Parameter for Short Channels

- In moderate inversion, \( \gamma \approx 1 \ldots 1.5 \)
MOSFET Model in Saturation

Noiseless!

Merely a modeling resistor that lets us account for finite \( \frac{dI_D}{dV_{DS}} \)

\[
\frac{i_d^2}{\Delta f} = 4kT \cdot \gamma \cdot g_m + \frac{K_f}{C_{ox}} \cdot \frac{g_m^2}{W \cdot L} \cdot \frac{1}{f}
\]

We’ll ignore flicker noise in this short course
Noise in Circuits

• Most circuits have more than one relevant noise source

• In order to quantify the net effect of all noise sources, refer the noise sources to a single "interesting" port of the circuit
  – Usually the output or input
Output and Input Referred Noise

\[ v_{out} = H \cdot v_{in} \quad \Rightarrow \quad \frac{\overline{v^2}_{n,in}}{\overline{v^2}_{n,\text{out}}} = \frac{\overline{v^2}_{n,\text{out}}}{|H|^2} \]

- \( H \) can be a continuous time or discrete time transfer function
Outline

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  – Delta-sigma modulator
Elementary T/H Model

\[ \phi \]

\[ V_{in} \rightarrow V_{out} \]

\[ \tau = RC \]

\[ V_{in} \rightarrow \text{Track} \quad \text{Hold} \rightarrow t \]

\[ V_{out} \rightarrow t \]
Step Response

- Circuit is typically designed to settle for $N > 5\ldots10$ time constants.

<table>
<thead>
<tr>
<th>Settling Error at $t = T_s/2$</th>
<th>Required $N = (T_s/2) / \tau$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1%</td>
<td>4.6</td>
</tr>
<tr>
<td>0.1%</td>
<td>6.9</td>
</tr>
<tr>
<td>0.01%</td>
<td>9.2</td>
</tr>
</tbody>
</table>
T/H Noise Analysis

- Question: What is the noise at the output?
- Can’t really answer this question without being more specific…

\[ v_n^2 = 4kTR\Delta f \]
T/H Noise Analysis

• Two possible ways to look at the noise
  – Noise in the continuous time output waveform
  – Discrete time, i.e. “frozen” noise values

• In SC circuits, we think about the signal in terms of discrete time samples, hence we should do the same for noise

\[ V_{\text{out}}(n-1) \quad V_{\text{out}}(n) \quad V_{\text{out}}(n+1) \quad \ldots \]

Distribution of Samples
T/H Noise Analysis

- A discrete time noise sequence has two important properties
  - The variance of the samples
  - The spectrum of the sequence
    - Noise samples are uncorrelated $\rightarrow$ white spectrum
    - Samples are correlated $\rightarrow$ colored spectrum
The variance of the samples is equal to the noise PSD integrated over all frequencies (Parseval)

\[
\text{var}[V_{\text{out}}(n)] = v_{\text{out,tot}}^2 = \int_0^\infty 4kTR \cdot \left| \frac{1}{1 + j2\pi f \cdot RC} \right|^2 \, df = 4kTR \cdot \frac{1}{4RC}
\]

\[
\frac{v_{\text{out,tot}}^2}{C} = \frac{kT}{C}
\]
Effect of Varying R

- Increasing R increases the noise PSD, but decreases the bandwidth
  - R drops out
- For C=1pF the total integrated noise is $\sim 64\mu$V rms
- Integral converges for $f > 10x$ RC bandwidth
Alternative Derivation

• The equipartition theorem says that each degree of freedom of a physical system in thermal equilibrium holds an average energy of $kT/2$.

• In our system, the degree of freedom is the energy stored on the capacitor.

\[
\frac{1}{2} C V_{out}^2 = \frac{1}{2} kT
\]

\[
V_{out}^2 = \frac{kT}{C}
\]
Does kT/C Noise Matter?

- Let’s look at the SNR of the elementary T/H circuit with a 1V sinusoid applied

<table>
<thead>
<tr>
<th>SNR [dB]</th>
<th>C [pF]</th>
</tr>
</thead>
<tbody>
<tr>
<td>20</td>
<td>0.00000083</td>
</tr>
<tr>
<td>40</td>
<td>0.000083</td>
</tr>
<tr>
<td>60</td>
<td>0.0083</td>
</tr>
<tr>
<td>80</td>
<td>0.83</td>
</tr>
<tr>
<td>100</td>
<td>83</td>
</tr>
<tr>
<td>120</td>
<td>8300</td>
</tr>
<tr>
<td>140</td>
<td>830000</td>
</tr>
</tbody>
</table>

Required C is smaller than physically realizable

Designer will be concerned about thermal noise; component sizes often set by SNR

Difficult battle with thermal noise, often resort to oversampling or external components
Spectrum of Noise Samples

- The noise samples are instantaneous values (~T_s/2 apart) of the track mode noise process

- **Intuition**
  - If the \( \tau = RC \) is small relative to \( T_s/2 \) the samples should show little correlation \( \rightarrow \) white spectrum
  - For large RC \( \rightarrow \) significant correlation, colored spectrum
Transient Noise Simulation

\[ N = \frac{T_s}{2RC} = 10 \]

\( V_{out} \)

Time
Transient Noise Simulation

\[ V_{out} \]

\[ N = \frac{T_s}{2RC} = 1 \]

Time

\( x10^{-8} \)
Calculating the Spectrum

- Spectrum follows from Fourier transform of the noise process' autocorrelation function (Wiener-Khinchin)
- The derivation in the following slides consists of two steps
  - Calculate autocorrelation function of noise at the output of the RC filter
  - Calculate the spectrum by taking the discrete time Fourier transform of the autocorrelation function
Analysis

Autocorrelation of resistor noise (white)

\[ R_{xx}(\tau) = \delta(0) \cdot 2kTR \]

Impulse response of RC filter

\[ h(t) = \frac{1}{\tau} e^{-t/\tau} \]

Autocorrelation of filtered noise

\[ R_{yy}(\tilde{\tau}) = R_{xx}(\tilde{\tau}) \ast h(\tilde{\tau}) \ast h(-\tilde{\tau}) \]

\[ R_{yy}(\tilde{\tau}) = \frac{kT}{C} e^{-|\tilde{\tau}|/\tau} \]

Covariance of samples separated by \( n \) clock cycles

\[ R_{yy}(n) = \frac{kT}{C} e^{-|n \cdot 0.5T_s|/\tau} = \frac{kT}{C} e^{-|n \cdot N|} \]
Analysis

\[ X(\omega) = \sum_{-\infty}^{\infty} R_{yy}(n) e^{j\omega n T_s} \]

\[ X(f) = \frac{2 kT}{f_s C} \frac{1 - e^{-2N}}{1 - 2e^{-N} \cos \left( \frac{2\pi f}{f_s} \right) + e^{-2N}} \]

Spectrum is essentially "white" for \( N > 3 \)

\[ PSD \approx \frac{2 kT}{f_s C} \]
PNOISE Simulation (More Later)

C=1pF

Noise PSD

Noise Integral

64μVrms

N=1

N=7

N=1

N=7
Noise Aliasing Interpretation

- Noise PSD of the samples
  \[ PSD_S = \frac{2 \ kT}{f_s \ C} \]

- Noise PSD of the resistor
  \[ PSD_R = 4kTR \]

- Ratio of the PSDs
  \[ \frac{PSD_S}{PSD_R} = \frac{1}{2f_s} \frac{1}{RC} = \frac{T_s / 2}{\tau} = N \]

- The increase in the noise PSD is due to aliasing of noise across all frequencies into the band from 0…\(f_s/2\)

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Noise Aliasing Interpretation

Ken Kundert, “Simulating Switched-Capacitor Filters with SpectreRF,”
Charge-Redistribution T/H

- When $\phi_{1e}$ goes low, the signal charge ($Q_x$) is acquired at node X.
- During $\phi_2$, this charge is redistributed onto $C_f$.

(actual implementation is fully differential)
Charge Conservation Analysis

- During $\phi_1$
  \[ Q_{x1} = -C_s V_{in} \]

- During $\phi_2$
  \[ Q_{x2} = -C_f V_{out} \]

- Charge Conservation
  \[ Q_{x1} = Q_{x2} \]
  \[ -C_s V_{in} = -C_f V_{out} \]

\[ \therefore V_{out} = \frac{C_s}{C_f} V_{in} \]
Noise Analysis During $\phi_1$

- Need to find the total noise charge at node X after the $\phi_1e$ switch has turned off
- Tedious to calculate using piece-by-piece integration of all three noise sources
Equi-partition to the Rescue

\[ \frac{1}{2} \frac{q_x^2}{C_{\text{eff}}} = \frac{1}{2} \frac{q_x^2}{C_s + C_f + C_{\text{par}}} \]

\[ q_x^2 = kT \left( C_s + C_f + C_{\text{par}} \right) \]

\[ \frac{1}{2} \frac{q_x^2}{2C_s + C_f + C_{\text{par}}} = \frac{1}{2} kT \]

\( C_{\text{par}} \) can deteriorate noise performance!
Noise Analysis During $\phi_2$

- Signal and noise charge acquired during $\phi_1$ is redistributed onto $C_f$
- Additional noise from $R_{on}$ and OTA is added
Single Stage OTA Model

\[ i_{eq}^2 = 4kT \gamma g_{mn} + 4kT \gamma g_{mp} \]

\[ i_{eq}^2 = 4kT \gamma g_{mn} \left( 1 + \frac{g_{mp}}{g_{mn}} \right) = 4kT \gamma g_{mn} \alpha \]
Noise Analysis During $\phi2$

- Simplifying assumption: $R_{on}$ has negligible effect on the circuit’s frequency response
  - Typically true in a practical design; it would be wasteful to let the switch (rather than the OTA) limit the bandwidth
OTA Time Constant

\[ \beta = \frac{C_f}{C_f + C_s + C_{par}} \]

Feedback factor

\[ R_{eq} = \frac{1}{\beta g_{mn}} \]

\[ C_{eq} = C_L + C_f (1 - \beta) \]

\[ \tau_{OTA} = R_{eq} C_{eq} \]
Output Referred Noise

\[ H(s) = \frac{1}{1 + sR_{eq}C_{eq}} \]

\[ \frac{\overline{V_{out}^2}}{\Delta f} = 4kT R_{on} \left( \frac{C_s}{C_f} \right)^2 |H(s)|^2 + \overline{i_{eq}^2 \cdot R_{eq}^2} |H(s)|^2 \]

- A
- B
Output Referred Noise

\[
\frac{A}{B} = \frac{4kT R_{on} \left( \frac{C_s}{C_f} \right)^2}{i_{ieq}^2 \cdot R_{eq}^2} = \frac{R_{on}}{\frac{1}{g_{mn}}} \cdot \frac{\beta^2 \left( \frac{C_s}{C_f} \right)^2}{\gamma \alpha} < < 1
\]

\[
\therefore \frac{v_{out}^2}{\Delta f} \cong i_{eq}^2 \cdot R_{eq}^2 |H(s)|^2
\]

• OTA noise typically dominates, assuming \( R_{on} < < 1/g_{mn} \)
Output Referred Noise

\[
\overline{v_{\text{out},2}^2} = \int_{0}^{\infty} 4kT \gamma_\alpha g_{mn} R_{eq}^2 \cdot \frac{1}{1 + j\omega R_{eq} C_{eq}}^2 \ df
\]

\[
= 4kT \gamma_\alpha g_{mn} R_{eq}^2 \cdot \frac{1}{4R_{eq} C_{eq}}
\]

\[
\overline{v_{\text{out}}^2} = \frac{1}{\beta C_{eq}} \frac{kT}{\gamma_\alpha}
\]

• Noise introduced during \( \phi 2 \) is \( \propto kT/C_{eq} \)
  – Depending on feedback factor and OTA topology
Adding Up $\phi_1$ and $\phi_2$ Noises

$$V_{out,1}^2 = \frac{q_x^2}{C_f^2} = \frac{kT(C_s + C_f + C_{par})}{C_f^2}$$

$$V_{out,2}^2 = \frac{1}{\beta C_{eq}} \gamma \alpha$$

$$V_{out,tot}^2 = \frac{kT}{C_f} \left( C_f + C_s + C_{par} \right) + \frac{1}{\beta C_{eq}} \gamma \alpha$$

$$V_{in,tot}^2 = \frac{kT}{C_s} \left( 1 + \frac{C_f + C_{par}}{C_s} \right) + \frac{1}{\beta C_{eq}} \gamma \alpha \left( \frac{C_f}{C_s} \right)^2$$
Noise/Power Tradeoff

• Reducing the noise by 6 dB means
  – Increase all C by 4x
  – Increase all $g_m$ by 4x to preserve speed
  – Increase $I_D$ by 4x to maintain same $g_m/I_D$

• Bottom line
  – Improving the dynamic range in a noise limited circuit by 6 dB ("1bit") quadruples power dissipation!
Noise in Differential Circuits

- In differential circuits, the noise power is doubled (due to two half circuits), and the signal power increases by 4x

\[
\begin{align*}
DR_{\text{single}} & \propto \frac{\hat{V}_{\text{out}}^2}{kT/C} \\
DR_{\text{diff}} & \propto \frac{(2\hat{V}_{\text{out}})^2}{2kT/C} = 2\frac{\hat{V}_{\text{out}}^2}{kT/C}
\end{align*}
\]

- 3 dB win in dynamic range, but at the expense of twice the power dissipation
  - Can get the same DR/power in a single ended circuit by doubling all cap sizes and $g_m$
SC Noise Simulation

- There are at least three ways to simulate noise in switched capacitor circuits

- Basic .ac/.noise Spice simulations
  - Simulate noise in each clock phase separately
    - Activate $\phi_1$ switches, run .noise and integrate noise charge at relevant node over all frequencies and refer to output
    - Activate $\phi_2$ switches, run .noise and integrate noise over all frequencies at the output
    - Sum integrated noise from the two phases
  - This is analogous to the way we carried out the hand analysis
SC Noise Simulation

• Periodic Steady State Simulation
  – First run PSS analysis to find the periodic operating point
    • Analogous to .op for .ac/noise
  – Next run PNOISE analysis
    • Computes total noise, taking all clock phases, noise aliasing, noise correlations, etc. into account

• Transient Noise
  – Direct simulation of all noise sources using a transient simulation
  – Most physical way of simulating noise
Example T/H Circuit

\[ f_s = 100 \text{ MHz}, \quad \alpha = 2, \quad \gamma = 1 \]
\[ C_s = C_f = 100 \text{ fF}, \quad C_L = 500 \text{ fF}, \quad C_{par} \cong 0 \]

- OTA \( G_m \) chosen such than \( (T_s/2) / \tau_{OTA} = 10 \)
- Switches sized 5 times faster, i.e. \( N = 5 \cdot 10 = 50 \)
.Noise Simulation (ϕ₁)

*** Compute noise charge at node X and refer to output via Cf

\[
en \ vno \ 0 \ \text{vcvs} \ vol = ( \ cs* v(x,s) + cf* v(x,f) )/cf
\]

.ac dec 100 100 1000Gig

.noise v(vno) vdummy

(Showing half circuit for simplicity)
Noise Simulation ($\phi_1$)

![PSD and Integral Plots]

- PSD [V^2/Hz]
- Frequency [Hz]
- Integral [uVrms]

- 406 $\mu$Vrms
Noise Simulation ($\phi 2$)

\[
\sqrt{v_{out,tot}^2} = \sqrt{(406\mu V_{rms})^2 + (266\mu V_{rms})^2} = 485\mu V_{rms}
\]
PSS Simulation Setup

Set “tstab” if your circuit needs time to reach steady state (e.g. clock bootstrap circuits).

Under “options” set “maxacfreq” to the highest frequency from which you expect noise to fold down.
PSS Waveforms (Clocks)

Periodic Steady-State Analysis `pss`: time = (0 s -> 10 ns)

- /p1
- /p1e
- /p2

4.75ns
PNOISE Simulation Setup

Numsidebands \( \cong \frac{f_{\text{max}}}{f_s} \), where \( f_{\text{max}} \) is the maximum frequency from which you expect significant noise folding.

“timedomain” means simulator computes spectrum of discrete time noise samples at the specified sampling instant.

Sampling instant (4.75ns in this example)
How to Chose Parameters

• Maxacfreq must be set commensurate with the speed of the switches
  – Common pitfall: Use nice 1mΩ switches → must consider noise from “DC to daylight”

• In our example

\[
\text{Maxacfreq} \equiv 10 \cdot \frac{1}{2\pi R_{on} C} = \frac{10}{\pi} \cdot N \cdot f_s
\]

\[
\text{Maxacfreq} \equiv 3 \cdot 50 \cdot 100\text{MHz} = 15\text{GHz}
\]
How to Chose Parameters

\[
\text{Numsidebands} \approx \frac{\text{Maxacfreq}}{f_s} = \frac{15\text{GHz}}{100\text{MHz}} = 150
\]

- In traditional PSS/PNOISE simulators (such as SpectreRF), simulation time increases rapidly for large values of Numsidebands.
- Berkeley Design Automation (BDA) Analog FastSPICE (AFS) automatically includes an infinite number of sidebands at significantly reduced simulation times.
PNOISE Result (SpectreRF)

Sampled Noise PSD

Integrated Noise
Transient Noise Using BDA AFS

- Simulated 1000 samples
- Only one critical setting: noisefmax = 15 GHz
Comparison

<table>
<thead>
<tr>
<th>Method</th>
<th>$\phi_1$ Noise [µVrms]</th>
<th>$\phi_2$ Noise [µVrms]</th>
<th>Total [µVrms]</th>
<th>Comment</th>
</tr>
</thead>
<tbody>
<tr>
<td>Calculation</td>
<td>406</td>
<td>245</td>
<td>474</td>
<td>Neglected switch noise during $\phi_2$ $\rightarrow$ smaller value than .NOISE</td>
</tr>
<tr>
<td>.NOISE</td>
<td>406</td>
<td>266</td>
<td>485</td>
<td>Somewhat smaller than .NOISE due to finite Maxacfreq and Numsidebands</td>
</tr>
<tr>
<td>PNOISE</td>
<td>-</td>
<td>-</td>
<td>475</td>
<td>Somewhat smaller than .NOISE due to statistical fluctuations (can use more samples)</td>
</tr>
<tr>
<td>TRAN</td>
<td>-</td>
<td>-</td>
<td>477</td>
<td></td>
</tr>
<tr>
<td>NOISE</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

- Very good agreement between hand calculation and all three simulation approaches
Advantages of TRAN NOISE

• Takes advantage of rapid advancements in very fast, spice-accurate transient simulators (such as BDA AFS)
  – Can handle much larger circuits compared to PSS/PNOISE (PSS tends to have convergence issues for large circuits)

• Intuitive inspection of waveforms

• No need to combine noises manually
  – Compared to .ac/.noise simulation flow

• Applicable also to non-periodic circuits
Second-Order $\Delta \Sigma$ Modulator

\[
\text{STF} = \frac{Y(z)}{X(z)} = z^{-2} \quad \text{NTF} = \frac{Y(z)}{E(z)} = (1 - z^{-1})^2
\]
Integrator

- Input referred noise similar to T/H stage
  - Output referred noise is colored due to integration → best to work with input noise

\[ V_{in, tot}^2 = \frac{kT}{C_s} \left( 1 + \frac{C_{par}}{C_s} \right) + \frac{1}{\beta C_{eq}} \frac{kT}{C_{eq}} \gamma \alpha \left( \frac{C_f}{C_s} \right)^2 \]
Second-Order $\Delta\Sigma$ Modulator

$$\frac{Y(z)}{N_1(z)} = z^{-2}$$

$$\frac{Y(z)}{N_2(z)} = 2 \left( 1 - z^{-1} \right) z^{-1}$$
In-Band Noise

\[
\text{OSR} = \frac{f_s / 2}{f_b}
\]
In-Band Noise

Total in-band thermal noise at modulator input:

\[ V_{\text{in, tot}}^2 = V_{\text{in, tot1}}^2 \cdot \frac{1}{\text{OSR}} + V_{\text{in, tot2}}^2 \cdot \frac{\pi^2}{3} \frac{1}{\text{OSR}^3} \]
Summary (1)

• General
  – Significance of $kT/C$ noise and steep trade-off with power dissipation
  – Whiteness of $kT/C$ noise samples
  – Noise folding

• Charge redistribution circuit analysis
  – Charge domain analysis of track mode noise
  – Watch out for parasitics at charge redistribution node
Summary (2)

• Analysis of $\Delta\Sigma$ modulator
  – Convenient to work with input referred, equivalent white noise generators and corresponding noise transfer functions
  – The same approach works for SC filters

• SC noise simulation
  – Recent developments in transient simulators have made TRAN NOISE simulation an attractive alternative to more conventional methods