CMOS Sigma-Delta Converters – From Basics to State-of-the-Art

Basic Concepts and Architectures

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   - Quantization noise shaping
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4. Continuous-Time ΣΔ Modulators
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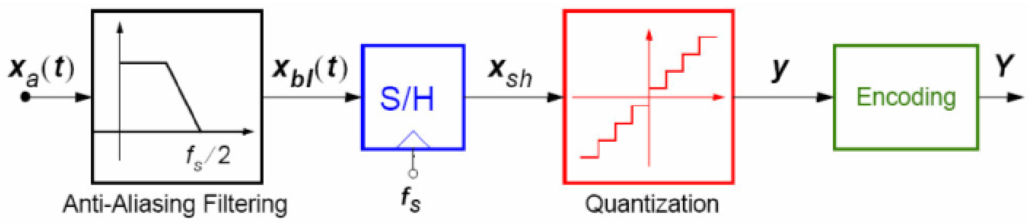
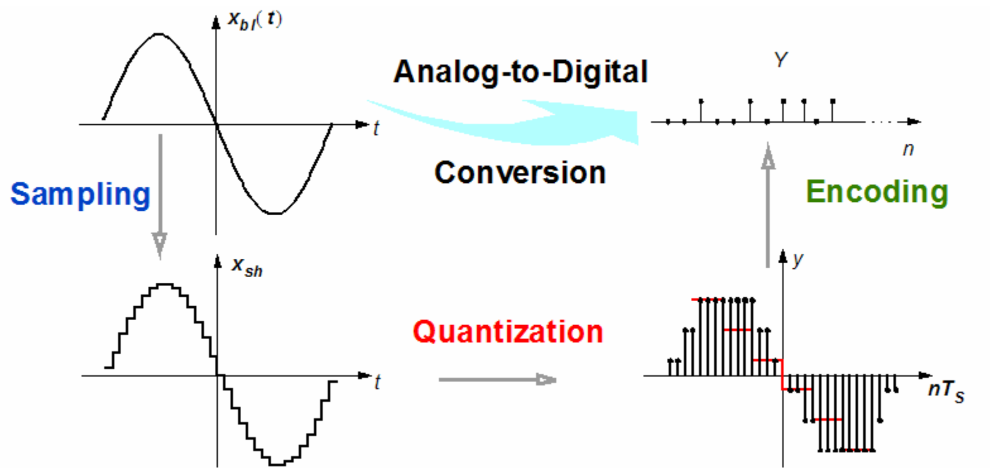
   - Synthesis methods

Introduction: Basic ADC process

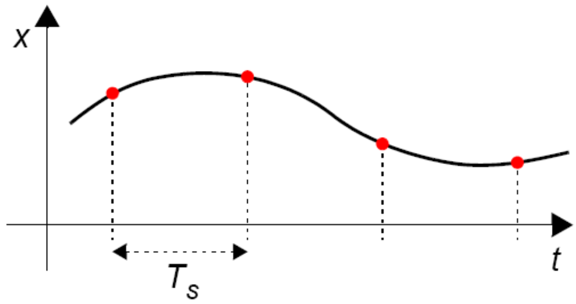




Introduction: Basic ADC process

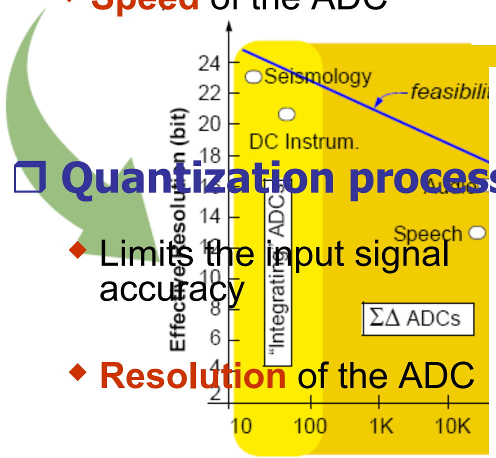
❒ Sampling process

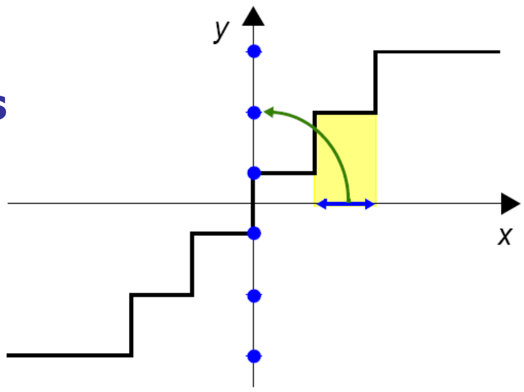
- Limits the input signal frequency
- **Speed** of the ADC



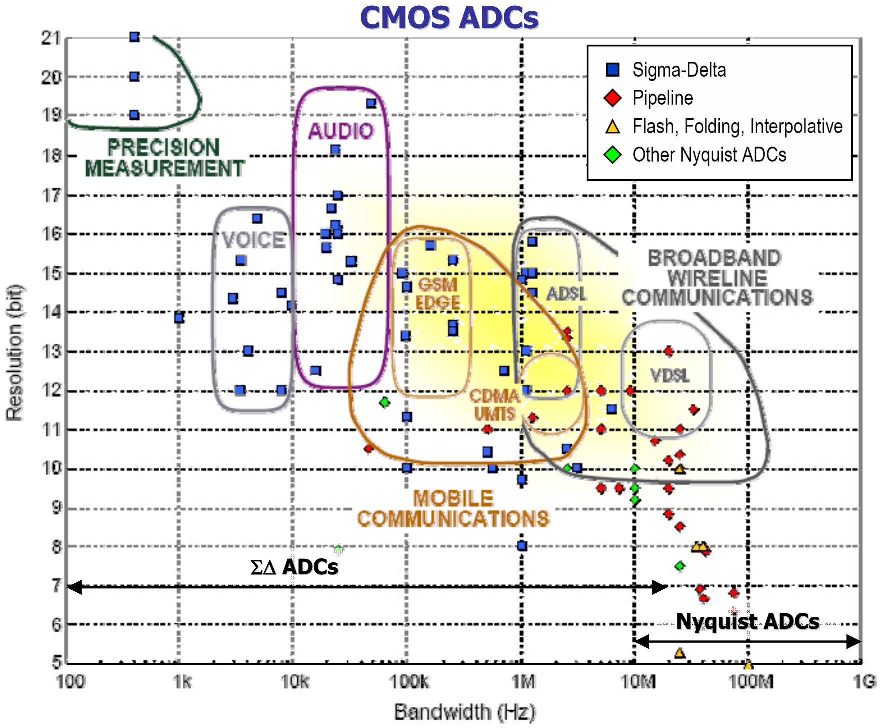
❒ Quantization process

- Limits the input signal accuracy
- **Resolution** of the ADC



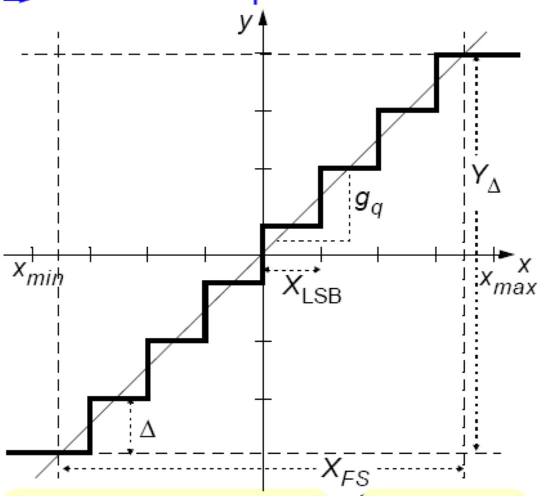


Introduction: Resolution vs. conversion rate

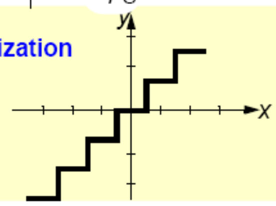


Introduction: Quantization

Midrise uniform quantization



Midtread quantization



- **Resolution (bits):**

$$B = \log_2(\#\text{ levels})$$

- **Separation between adjacent input levels:**

$$X_{LSB} = \frac{X_{FS}}{(2^B - 1)}$$

- **Separation between adjacent output levels:**

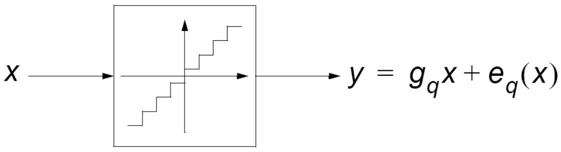
$$\Delta = \frac{Y_\Delta}{(2^B - 1)}$$

- **Full-scale input range:** X_{FS}
- **Gain:**

$$g_q = \frac{\Delta}{X_{LSB}} = \frac{Y_\Delta}{X_{FS}}$$

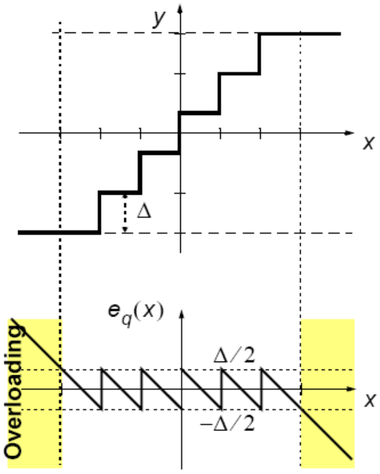
Introduction: Quantization

- **Quantization input-output characteristic**

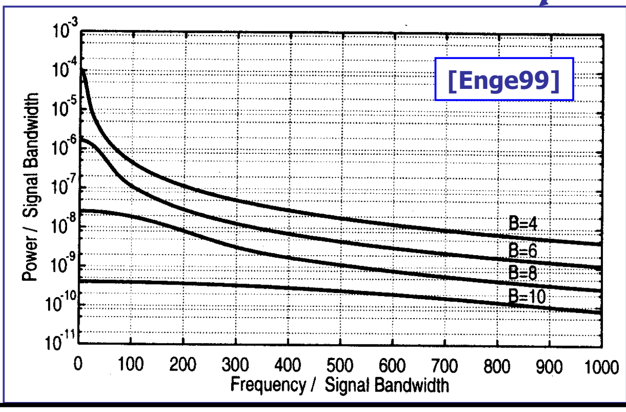


$$x \rightarrow \boxed{\text{quantizer}} \rightarrow y = g_q x + e_q(x)$$

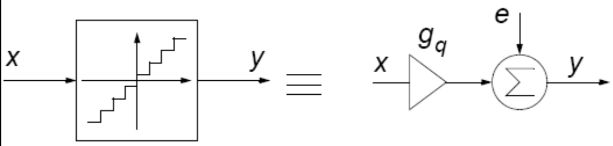
- **Quantization error**



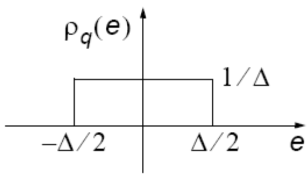
- **White noise model**
  - If x varies randomly from sample to sample
  - If the # of quantizer levels is high



Introduction: Quantization - white noise model



- **Probability Density Function**

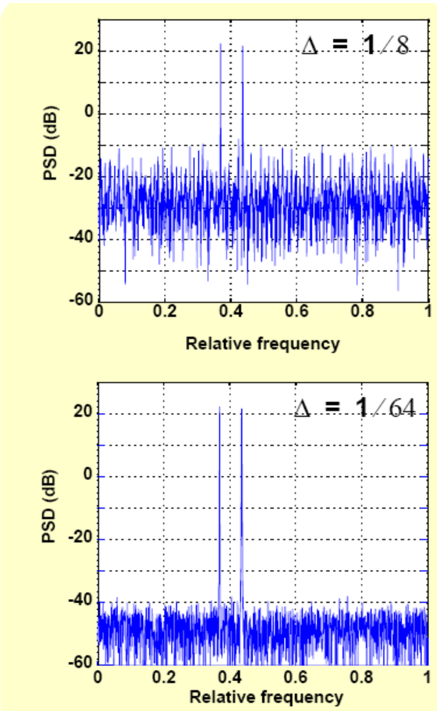


- **Quantization error power**

$$\sigma^2(e) = \left[\frac{1}{\Delta}\int_{-\Delta/2}^{\Delta/2} e^2 de\right] = \frac{\Delta^2}{12}$$

- **Quantization error Power Spectral Density**

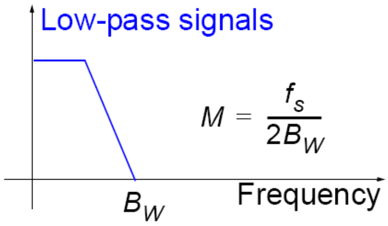
$$S_E(f) = \frac{\sigma^2(e)}{f_s} = \frac{\Delta^2}{12 f_s}$$



Introduction: Sampling

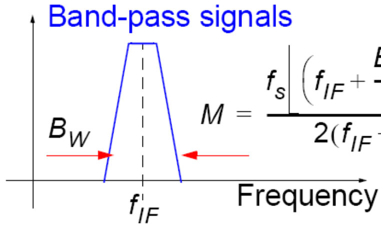
❒ Oversampling

Low-pass signals



$$M = \frac{f_s}{2B_W}$$

Band-pass signals

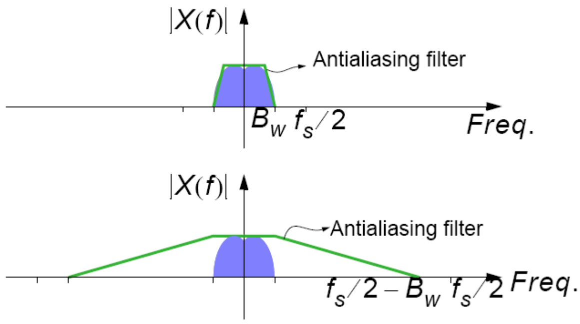


$$M = \frac{f_s \left\lfloor \left(f_{IF} + \frac{B_W}{2}\right) / B_W \right\rfloor}{2(f_{IF} + B_W)} \cong \frac{f_s}{2B_W}$$

$OSR \equiv M \equiv$ Oversampling Ratio

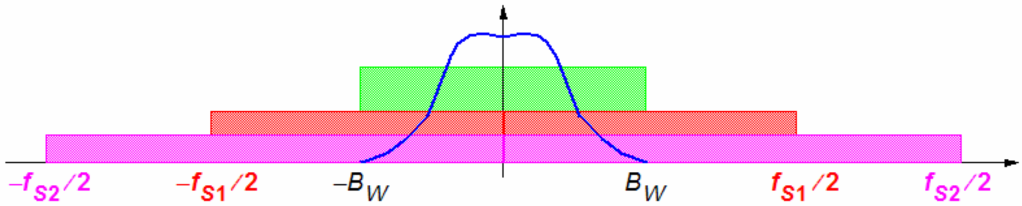
❒ Classification of ADCs

- Nyquist-rate ADCs (M~1)
- Oversampling ADCs (M>>1)



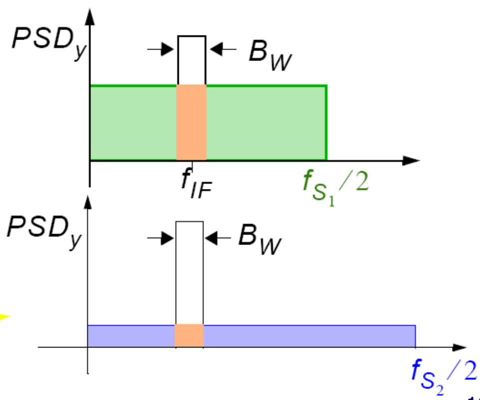
Fundamentals of ΣΔ ADCs: Oversampling

- **PSD of oversampled quantization noise**

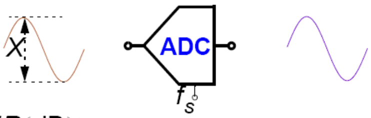


- **In-Band Noise power (IBN or P_Q)**

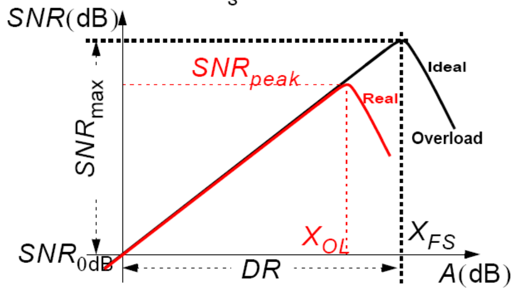
$$P_Q = \int_{f_{IF}-B_w/2}^{f_{IF}+B_w/2} 2S_E(f)df = \frac{B_w}{6f_s}\Delta^2 = \frac{\Delta^2}{12M}$$



Fundamentals of ΣΔ ADCs: Performance Metrics



$P_{sig} = \frac{X^2}{8}$ +  $P_Q = \frac{B_w}{6f_s}\Delta^2$



$$SNR(\mathrm{dB}) = 10\log_{10}\left(\frac{P_{sig}}{P_Q}\right) =$$

$$= 10\log_{10}\left[\frac{3}{2}M(2^B-1)^2\left(\frac{X}{X_{FS}}\right)^2\right]$$

$$SNR_{max}(\mathrm{dB}) = 10\log_{10}\left[\frac{3}{2}M(2^B-1)^2\right]$$

$$DR = 10\log_{10}\left[\frac{(X_{FS}/2)^2}{2P_Q}\right]$$

- **N-bit Nyquist-Rate ADC**
  - $f_{s1} = f_N \equiv 2B_w$
  - $SNR_{max} = 10\log_{10}\left[\frac{3}{2}(2^N-1)^2\right]$
- **B-bit Oversampled ADC**
  - $f_{s2} = Mf_N (M > 1)$
  - $SNR_{max} = 10\log_{10}\left[\frac{3}{2}M(2^B-1)^2\right]$

⇨ ***ENOB*** $N \cong \frac{SNR_{max} - 1.76}{6.02} \cong \log_2(2^B-1) + \frac{1}{2}\log_2(M)$ $(N > 1)$

Fundamentals of ΣΔ ADCs: Performance Metrics

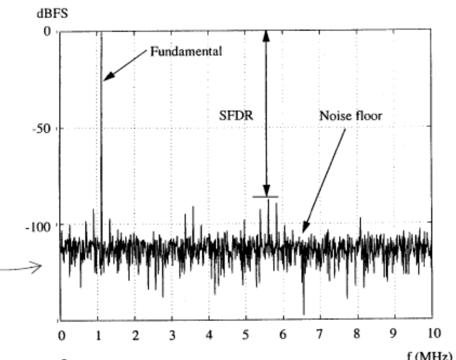
- *SNDR* / *SINAD*:

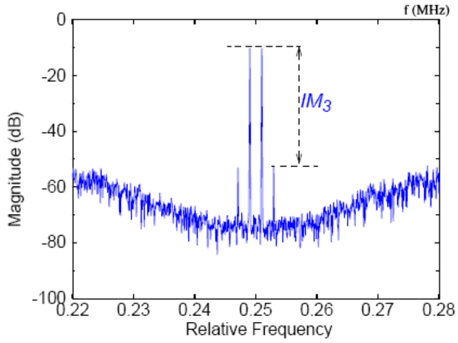
  $$SNDR(\mathrm{dB}) = 10\log_{10}\left(\frac{P_{sig}}{P_Q + P_H}\right)$$

- **Effective Number Of Bits** *ENOB*:

  $$ENOB \cong \frac{SNDR - 1.76}{6.02}$$

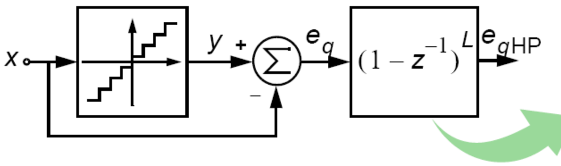
- *SFDR*: **Spurious-Free Dynamic Range**
- **Harmonic Distortion:**
  - HD_k, THD,
  - IM_3, IP_3
  - ...





- **Processing of the quantization error**

- If $f_i \ll f_s$, $|e_q(n) - e_q(n-1)| \ll |e_q(n)|$



$L = 1$

$e_{qHP}(n) = e_q(n) - e_q(n-1)$

$L = 2$

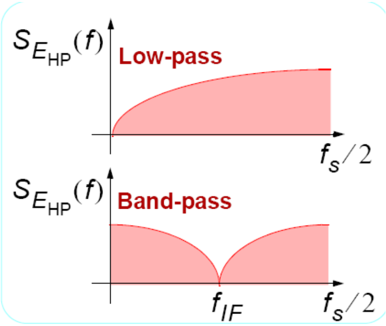
$e_{qHP}(n) = e_q(n) + e_q(n-2) - 2e_q(n-1)$

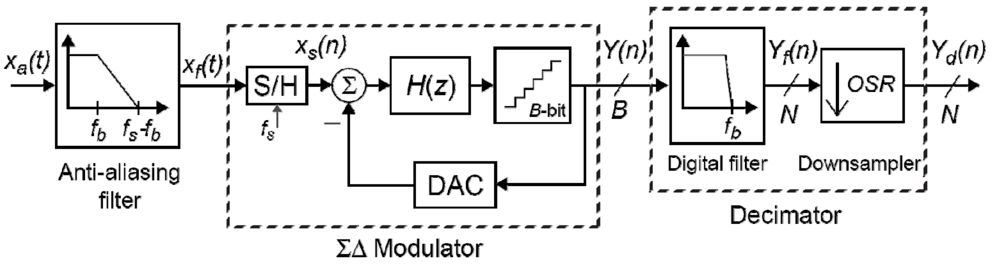
- **In-band noise power and effective resolution**

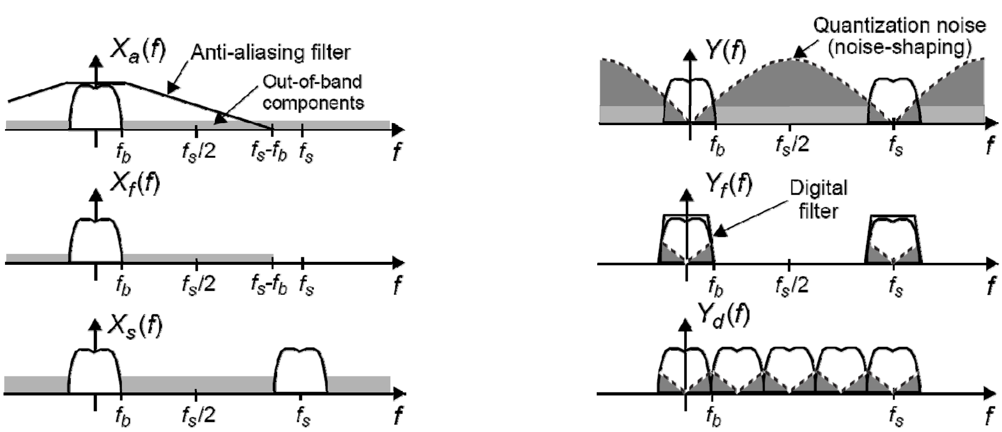
$$N_{TF}(z) = (1 - z^{-1})^L \Rightarrow S_{E_{HP}} = |N_{TF}(f)|^2 S_E$$

$$P_{E_{HP}} = \int_0^{B_w} S_{E_{HP}}(f)df \cong \frac{\Delta^2}{12}\frac{\pi^{2L}}{(2L+1)M^{2L+1}}$$

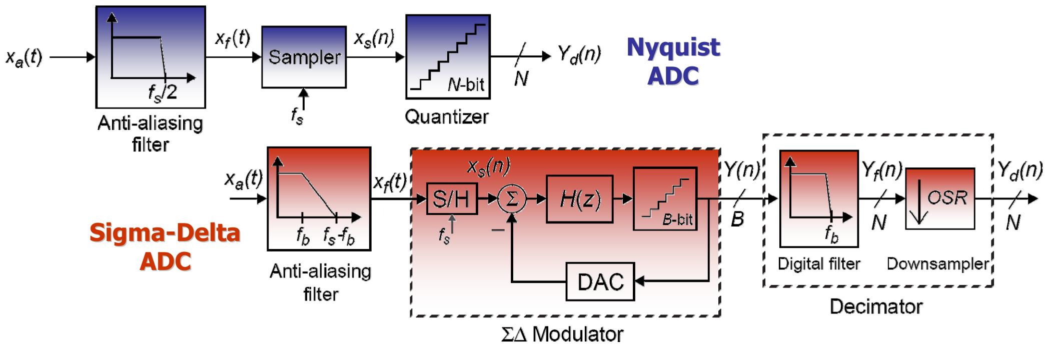
$$\boxed{N \cong \log_2\left[\frac{(2^B - 1)(2L+1)}{\pi^{2L}}\right] + \left(L + \frac{1}{2}\right)\log_2(M)}$$







Fundamentals of ΣΔ ADCs: Nyquist-rate vs. ΣΔ ADCs

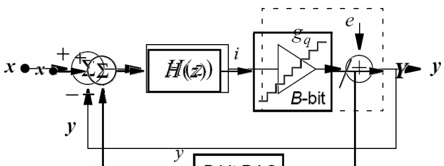


- HIGH-SELECTIVITY ANALOG FILTER for anti-aliasing
- Overall resolution obtained using HIGH-ACCURACY ANALOG BLOCKS

- LOW-SELECTIVITY ANALOG FILTER for anti-aliasing (1st/2nd order)
- High overall resolution obtained using LOW/MODERATE-ACCURACY ANALOG BLOCKS
- HIGH-SELECTIVITY DIGITAL FILTER

EASIER AND MORE ROBUST IN MODERN CMOS

Fundamentals of ΣΔ ADCs: Basic ΣΔM architecture



$$Y(z) = STF(z)X(z) + NTF(z)E(z)$$

H(z) with large gain within the signal band

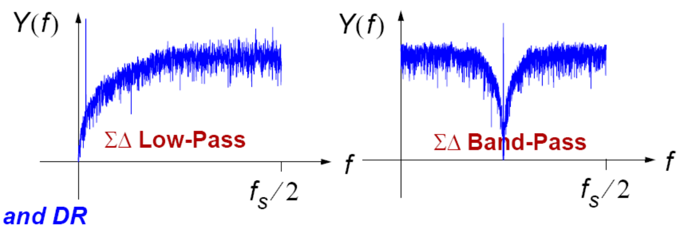
$$STF(z) = \frac{g_q H(z)}{1+g_q H(z)} \rightarrow STF(z) \approx 1$$

$$NTF(z) = \frac{1}{1+g_q H(z)} \rightarrow NTF(z) \approx \frac{1}{g_q H(z)} \ll 1$$

Within the signal bandwidth

$$|S_{TF}(z)| = \frac{H(z)}{1+H(z)} \rightarrow 1$$

$$N_{TF}(z) = \frac{1}{1+H(z)} \rightarrow 0$$



In-band noise power, SNR and DR

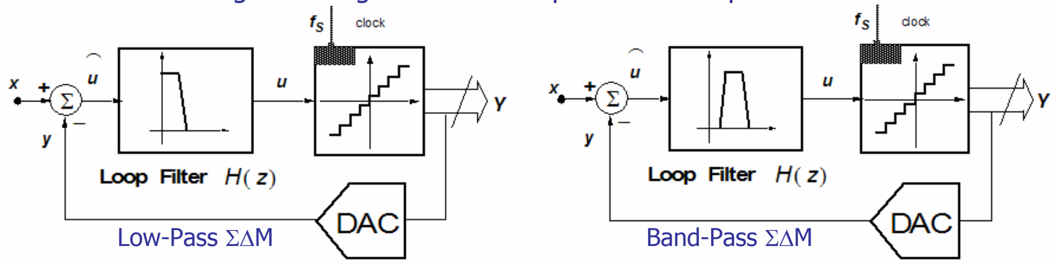
$$P_Q = \begin{cases} \dfrac{\Delta^2}{6f_s}\displaystyle\int_0^{B_w} |N_{TF}(f)|^2 df & \text{for LP}\Sigma\Delta \\ \dfrac{\Delta^2}{6f_s}\displaystyle\int_{f_n-B_w/2}^{f_n+B_w/2} |N_{TF}(f)|^2 df & \text{for BP}\Sigma\Delta \end{cases}$$

$$SNR = 10\log_{10}\left(\frac{A^2/2}{P_Q}\right)$$

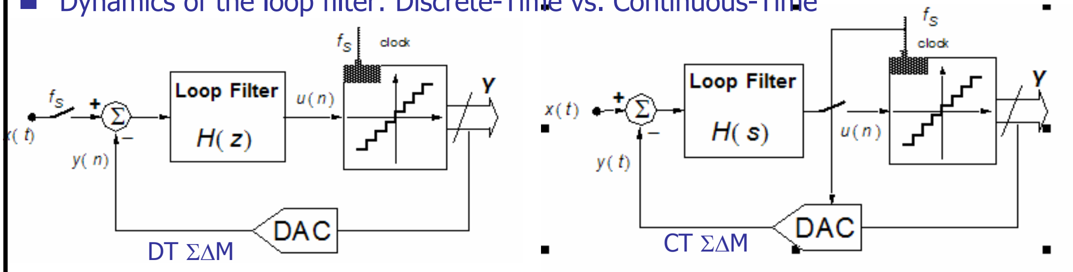
$$DR = 10\log_{10}\left[\frac{(X_{FS})^2/2}{P_Q}\right]$$

Fundamentals of ΣΔ ADCs: Classification of ΣΔMs

- Nature of the signals being handled: Low-pass vs. Band-pass

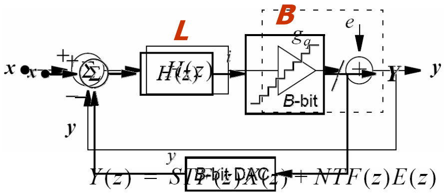


- Dynamics of the loop filter: Discrete-Time vs. Continuous-Time



- Number of bits of the embedded quantizer: single-bit vs. multi-bit
- Number of quantizers employed: single-loop, cascade, etc..
- Type of primitives available in the fabrication technology…

Fundamentals of ΣΔ ADCs: Basic control parameters



$$Y(z) = STF(z)X(z) + NTF(z)E(z)$$

$$STF(z) = \frac{g_q H(z)}{1 + g_q H(z)} \rightarrow STF(z) \approx 1$$

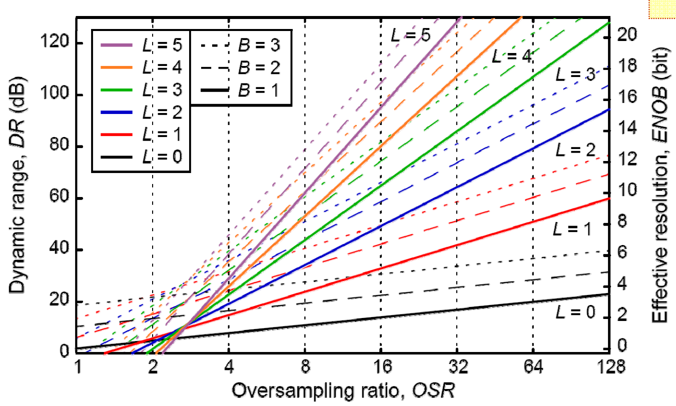
$$NTF(z) = \frac{1}{1 + g_q H(z)} \rightarrow NTF(z) \approx \frac{1}{g_q H(z)} \ll 1$$

H(z) with large gain within the signal band

L th-order ΣΔM

$$Y(z) = z^{-L}X(z) + (1 - z^{-1})^L E(z)$$

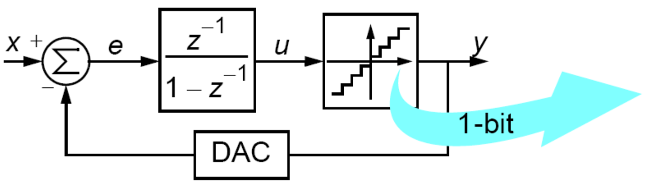
$$DR \approx \frac{3}{2}(2^B - 1)^2 \cdot \frac{(2L+1)OSR^{(2L+1)}}{\pi^{2L}}$$



- **Oversampling, *OSR***
  Speed of analog circuitry
- **Order of the shaping, *L***
  Stability of the ΣΔM
- **Resolution of the internal quantizer, *B***
  Linearity of the DAC

DT-ΣΔMs: 1st-order LP ΣΔ Modulator

$$N_{TF}(z)\big|_{z=1} \to 0 \quad \Rightarrow \quad \left.\frac{1}{1+H(z)}\right|_{z=1} \to 0 \quad \Rightarrow \quad \boxed{H(z) = \frac{1}{z-1}}$$

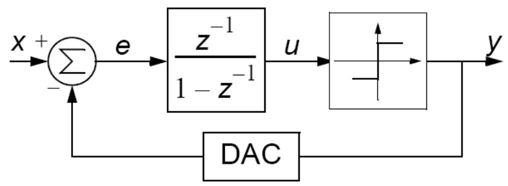


$$\begin{aligned} e(n) &= x(n) - y(n) \\ u(n) &= u(n-1) + e(n) \\ y(n) &= \mathrm{sgn}[u(n)] \end{aligned}$$

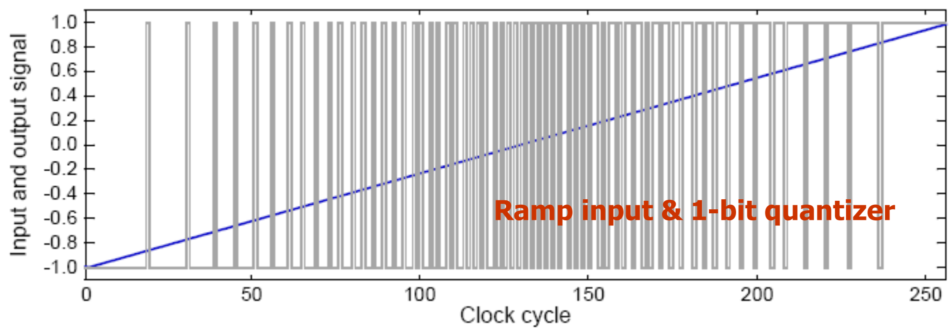
- Using a linear model for the quantizer

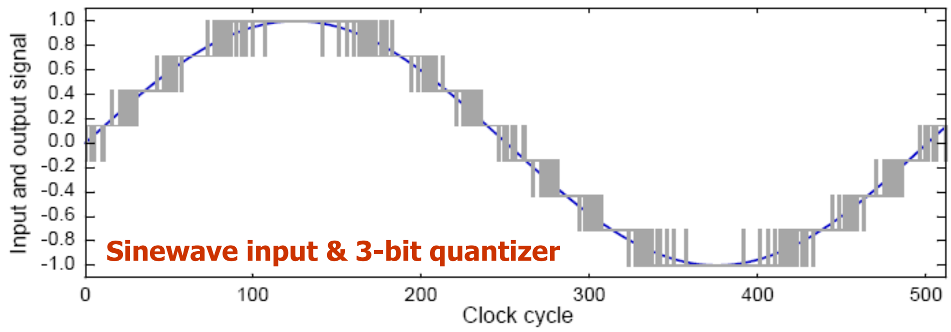
$$Y(z) = z^{-1}X(z) + (1-z^{-1})E(z)$$

$$DR(\mathrm{dB}) \cong 10\log_{10}\left(\frac{9M^3}{2\pi^2}\right)$$

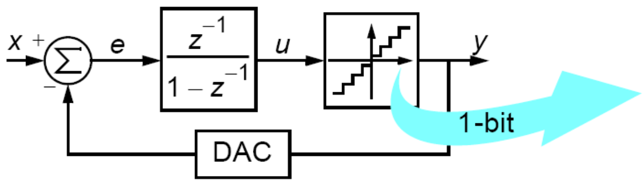


DT-ΣΔMs: 1st-order LP ΣΔ Modulator



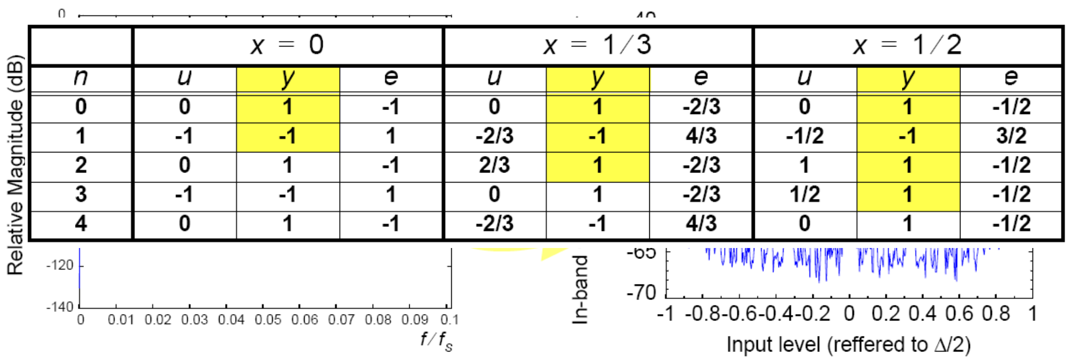


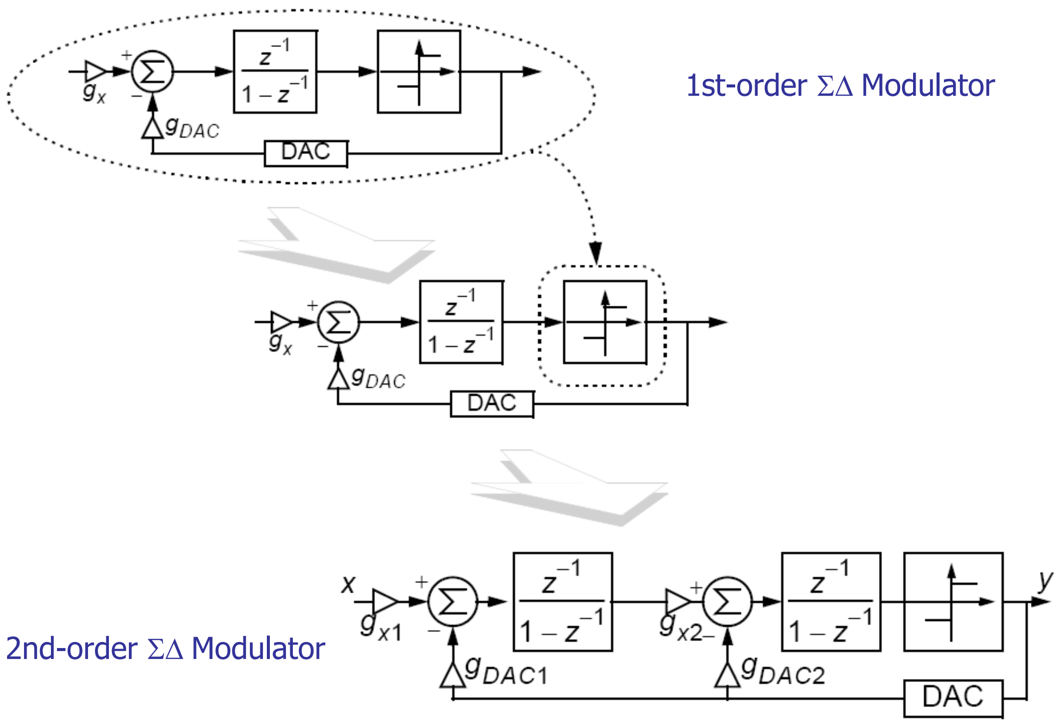
- Noise pattern



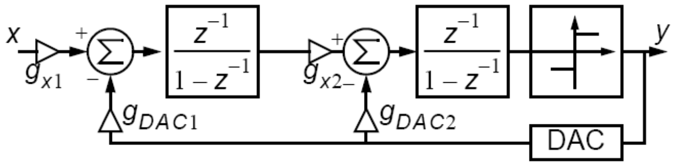
$$e(n) = x(n) - y(n)$$
$$u(n) = u(n-1) + e(n)$$
$$y(n) = \mathrm{sgn}[u(n)]$$

	$x = 0$			$x = 1/3$			$x = 1/2$		
n	u	y	e	u	y	e	u	y	e
0	0	1	-1	0	1	-2/3	0	1	-1/2
1	-1	-1	1	-2/3	-1	4/3	-1/2	-1	3/2
2	0	1	-1	2/3	1	-2/3	1	1	-1/2
3	-1	-1	1	0	1	-2/3	1/2	1	-1/2
4	0	1	-1	-2/3	-1	4/3	0	1	-1/2





DT-ΣΔMs: 2nd-order LP ΣΔ Modulator



Stability conditions:

$$g_{DAC1} g_{x2} g_q = 1$$

$$g_{DAC2} = 2 g_{DAC1} g_{x2}$$

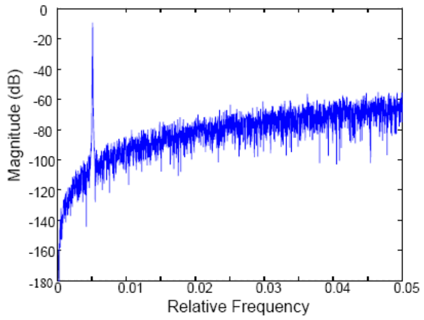
- Linear analysis

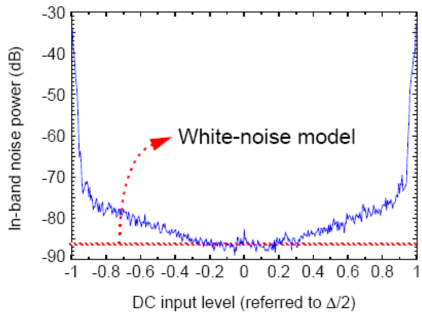
$$Y(z) = z^{-2}X(z) + (1-z^{-1})^2 E(z)$$

$$P_Q \cong \frac{\Delta^2 \pi^4}{60M^5} \Rightarrow DR \cong \frac{15M^5}{2\pi^4}$$

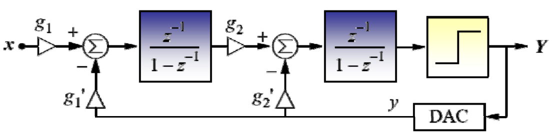
- **Dependence on M: 15 dB/oct.**
- **Example: digitize a 10kHz signal with 16 bits**
  - M = 150 (f_s = 3 MHz) for a 2nd-order ΣΔM
  - M = 1500 (f_s = 30 MHz) for a 1st-order ΣΔM

- Output spectrum and noise pattern





DT-ΣΔMs: High-order Single-loop ΣΔ Modulators

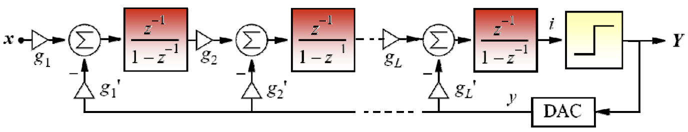


2nd-order ΣΔM

$$Y(z) = z^{-2}X(z) + (1-z^{-1})^2 E(z)$$

$$g_1' g_2 g_q = 1 \qquad g_2' = 2 g_1' g_2$$

Stable for inputs in $[-0.9\Delta/2, +0.9\Delta/2]$ if $g_2' > 1.25 g_1' g_2$ [Candy85]

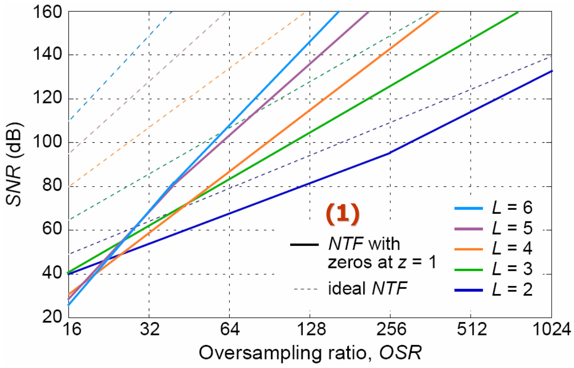


L th-order ΣΔM

$$Y(z) = z^{-L}X(z) + (1-z^{-1})^L E(z)$$

pure-differentiator FIR *NTF*

$\|NTF\|_\infty = 2^L$ Prone to instability



High-order ΣΔ loops are only conditionally stable [OptE90]

IIR *NTF*s [Lee87]

$$NTF(z) = \frac{(z-1)^L}{D(z)} \quad \textbf{(1)}$$

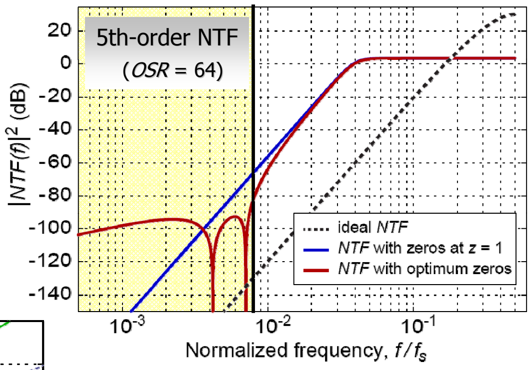
- Zeros at $z = 1$
- Butterworth/Chebyshev poles

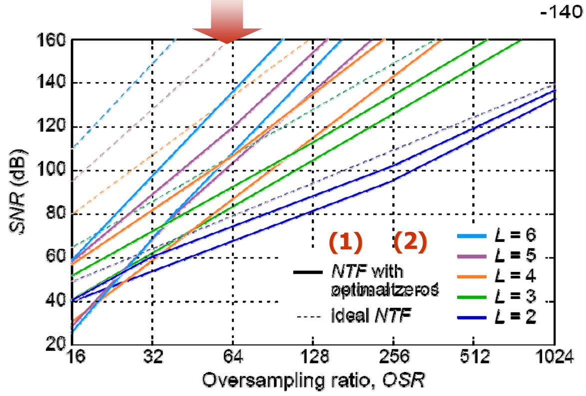
Gain adjusted to satisfy $\|NTF\|_\infty \sim 1.5$

OPTIMIZED IIR *NTFs* [Schr93]

$$\min\left[\int_0^{f_b} |NTF(f)|^2 df\right] \quad \textbf{(2)}$$

- Complex zeros at $|z| = 1$ with optimal positions within the signal band
- Butterworth/Chebyshev poles





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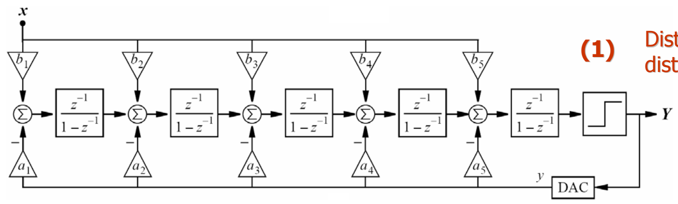
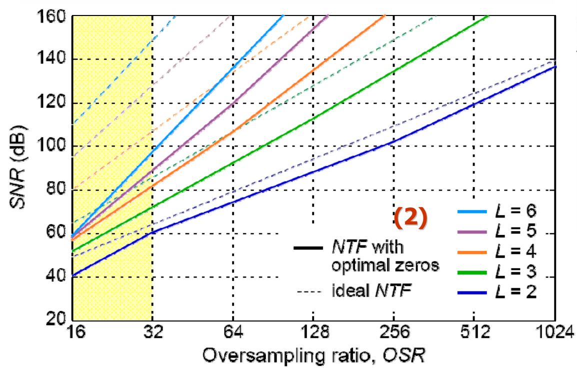
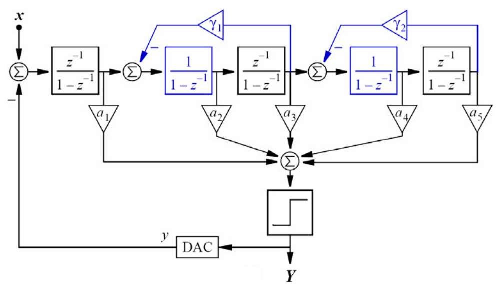


(1) Distributed feedback and distributed feedforward input

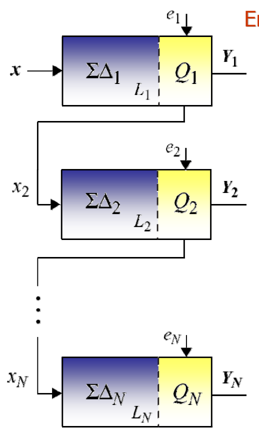
- Complexity (many feedback/feedforward coeffs)
- Large spread of coeffs (area, power)

(2) Feedforward summation + local resonators





DT-ΣΔMs: High-order Cascade ΣΔ Modulators



Error Cancellation Logic (ECL)

$$Y(z) = z^{-L}X(z) + d_{2N-3}(1-z^{-1})^L E_N(z)$$

$$L = L_1 + L_2 + \ldots + L_N$$

- HIGH-ORDER STABLE OPERATION is ensured by cascading low-order stages (L_i = 1, 2).
- Relationships among ECL and ΣΔM to be fulfilled for perfect cancellation (NOISE LEAKAGE).

d > 1, interstage coupling

$$P_Q \cong d_{2N-3}^2 \cdot \frac{\Delta_N^2}{12} \cdot \frac{\pi^{2L}}{(2L+1)OSR^{(2L+1)}}$$

Systematic loss of resolution, but:

- Smaller than for single loops
- Independent of *OSR*

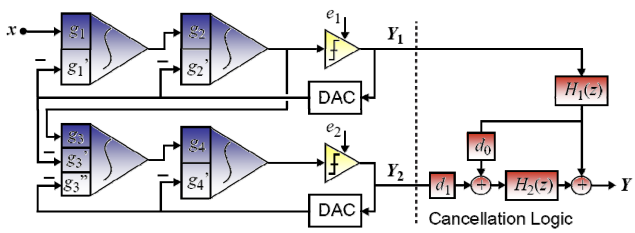
MASH ΣΔMs

- Each stage re-modulates a signal containing the quantization error in the previous one.
- Digital processing is used to cancel out all quantization errors, but that in the last stage.

$$NTF_i(z) = 0 \quad , i = 1, \ldots, N-1$$

- Small spread of analog coeffs
- ECL can be easily implemented
- Performance close to ideal
- Suited at low oversampling

DT-ΣΔMs: High-order Cascade ΣΔ Modulators



Analog	Digital	
$g_2' = 2g_1'g_2$	$d_0 = \frac{g_3'}{g_1'g_2g_3} - 1$	$H_1(z) = z^{-2}$
$g_4' = 2g_3''g_4$	$d_1 = \frac{g_3''}{g_1'g_2g_3}$	$H_2(z) = (1-z^{-1})^2$

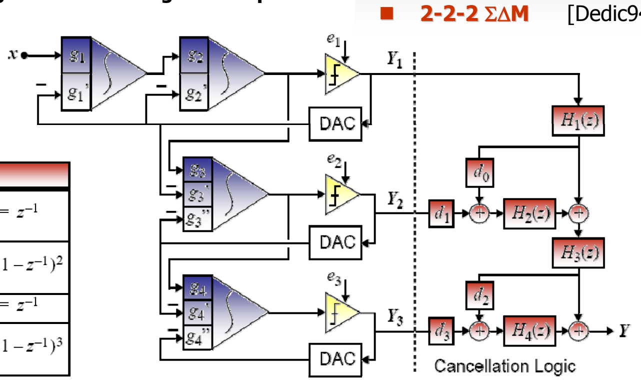
2-2 ΣΔM [Kare90]
4th-order 2-stage cascade

Noise leakage precludes the cascading of a large number of stages to be practical

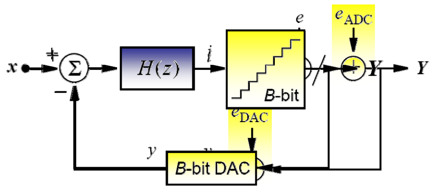
- **1-1-1 ΣΔM** [Mats87]
- **2-1 ΣΔM** [Longo88]
- **2-2-1 ΣΔM** [Vleu01]
- **2-1-1-1 ΣΔM** [Rio00]
- **2-2-2 ΣΔM** [Dedic94]

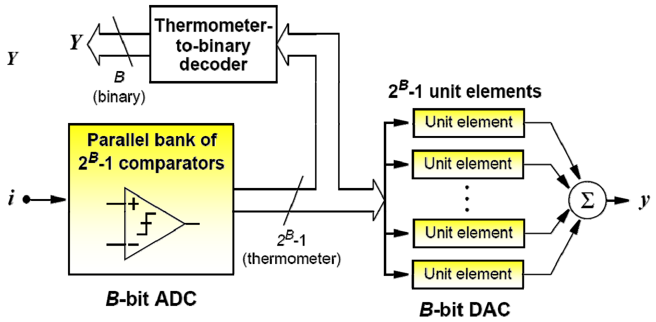
2-1-1 ΣΔM [Yin94]
4th-order 3-stage cascade

Analog	Digital	
$g_2' = 2g_1'g_2$	$d_0 = \frac{g_3'}{g_1'g_2g_3} - 1$	$H_1(z) = z^{-1}$
$g_4' = g_3''g_4$	$d_1 = \frac{g_3''}{g_1'g_2g_3}$	$H_2(z) = (1-z^{-1})^2$
	$d_2 = 0$	$H_3(z) = z^{-1}$
	$d_3 = \frac{g_4''}{g_1'g_2g_3g_4}$	$H_4(z) = (1-z^{-1})^3$



DT-ΣΔMs: Multi-bit ΣΔ Modulators





- Increased dynamic range
  B can trade for *OSR* (wideband)
- Better stability properties
  More aggressive high-order *NTF*s
- DAC non-linearities are directly added to the input
  The linearity of the ΣΔM will be no better than that

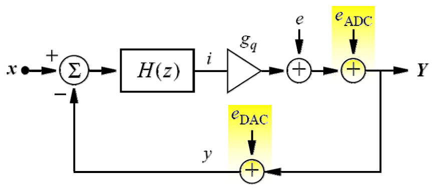
FULL-PARALLEL ADC/DAC
(Typically $B < 6$)

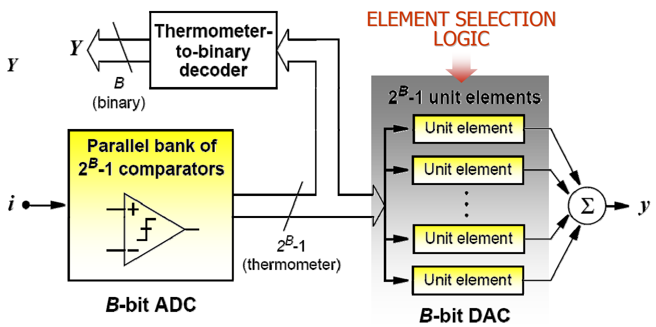
DAC linearity limited by component mismatch

POSSIBLE APPROACHES

- Correcting DAC errors
  - Element Trimming
  - Analog Calibration
  - Digital Correction
- Decorrelating DAC errors from the input
  - DEM techniques
- Introducing DAC errors at a non-critical position
  - Dual quantization

DT-ΣΔMs: Multi-bit ΣΔ Modulators





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FULL-PARALLEL ADC/DAC
(Typically $B < 6$)

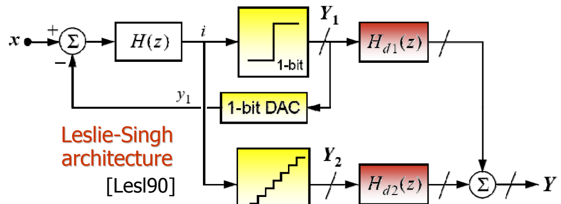
DAC linearity limited by component mismatch

Dynamic Element Matching (DEM)

- Elements selected to make DAC errors independent of the input signal
- Algorithms that try to average the error in each DAC level to zero (to push DAC errors to high freq.)
  - Randomization: Distortion transforms into white noise
  - Rotation: Distortion moves out of band (CLA)
  - Mismatch-shaping: 1st/2nd order (ILA, DWA, DDS)

Dual Quantization

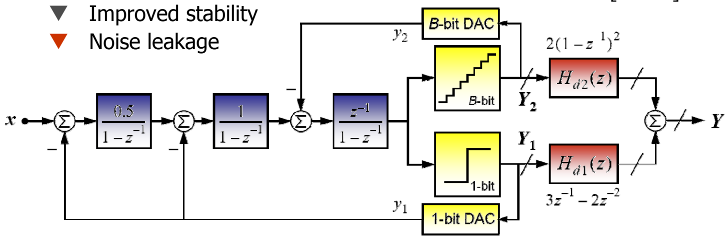
- Combines 1-bit and multi-bit quantizers (linearity/reduced error)



- L-0 cascade ΣΔM
- Suffers from noise leakage
- Multi-bit quantization does not improve stability

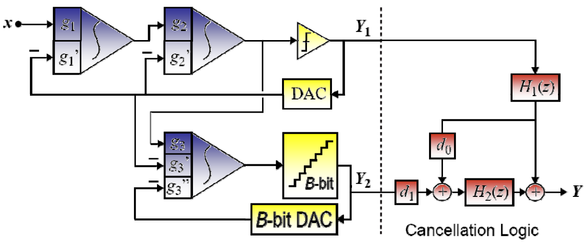
Concept applied to single-loop ΣΔMs [Hair91]

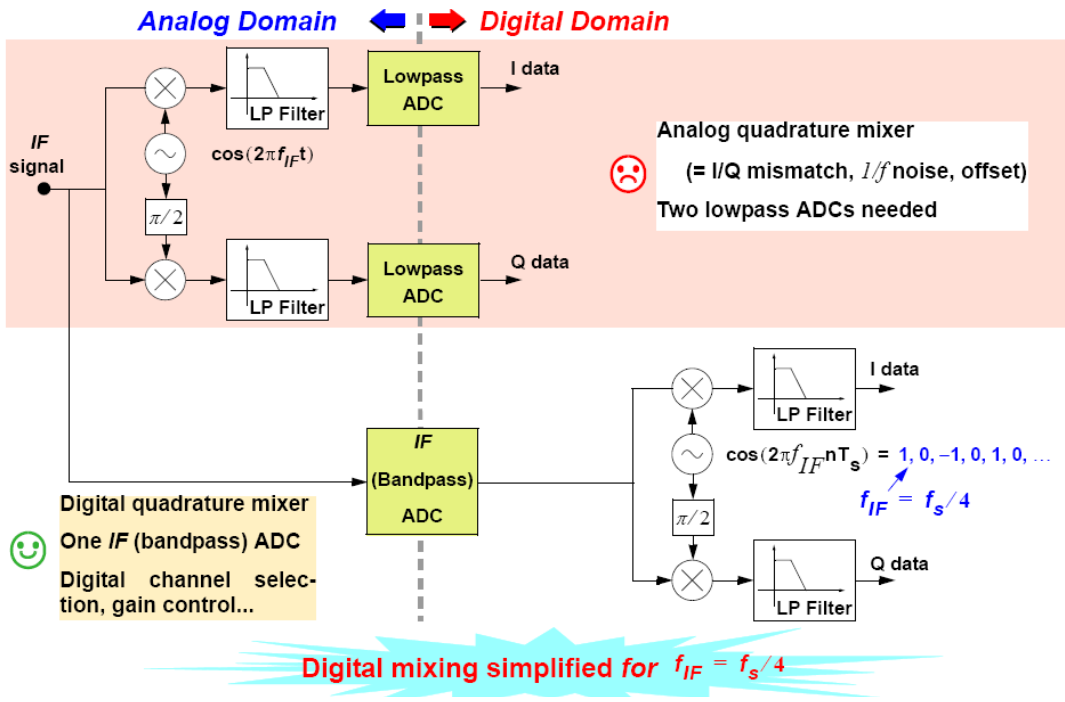
- Improved stability
- Noise leakage



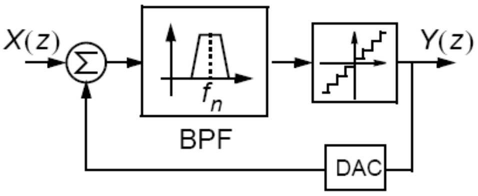
Concept applied to cascade ΣΔMs [Bran91]

- Multi-bit quantization usually applied only in the last stage
- DAC errors shaped by L-L_N
  Relaxes DAC requirements
- Noise leakage (inherent to cascades)

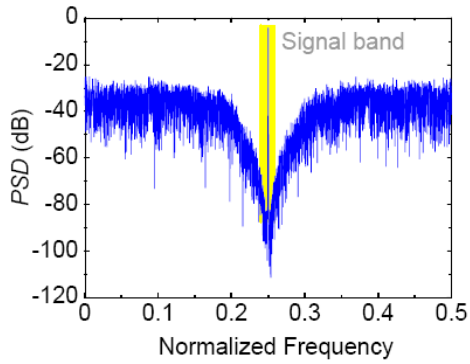


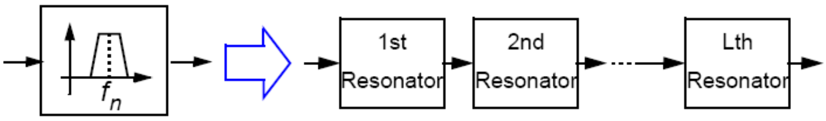


DT-ΣΔMs: Bandpass ΣΔ Modulators



$$Y(z) = S_{TF}(z)X(z) + N_{TF}(z)E(z)$$





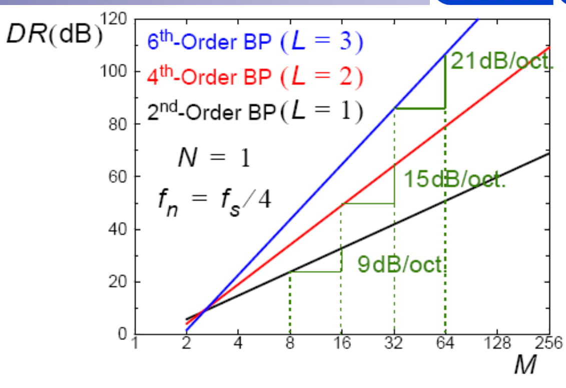
$$H_{bp}(z) = \left[\frac{N_{RES}(z)}{(1-z^{-1}z_n)(1-z^{-1}z_n^*)}\right]^L \qquad (z_n = e^{2\pi f_n T_s})$$

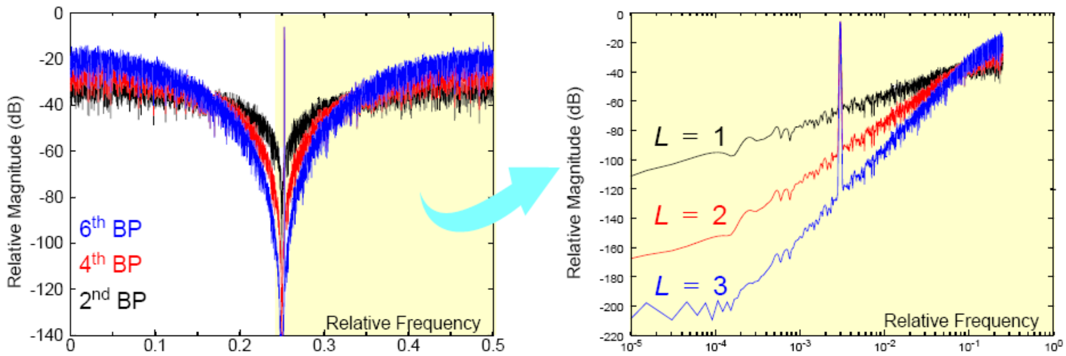
$$(N_{RES}(z) + (1-z^{-1}z_n)(1-z^{-1}z_n^*) = 1) \Rightarrow \boxed{N_{TF}(z) = [1 - 2\cos(2\pi f_n T_s)z^{-1} + z^{-2}]^L}$$

DT-ΣΔMs: Bandpass ΣΔ Modulators

$$P_Q \cong \frac{(\sin[2\pi f_n T_s])^{2L}\pi^{2L}X_{FS}^2}{12(2^N-1)^2(2L+1)M^{(2L+1)}}$$

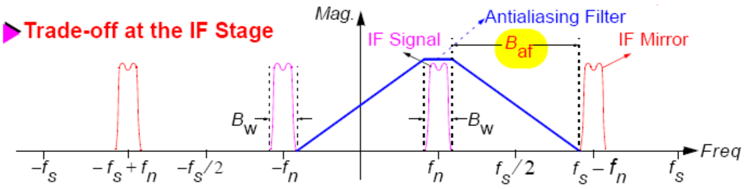
$$DR \cong \frac{3(2^N-1)^2(2L+1)M^{2L+1}}{2\pi^{2L}(\sin[2\pi f_n T_s])^{2L}}$$





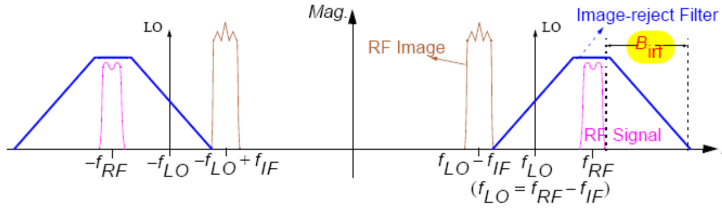
DT-ΣΔMs: Bandpass ΣΔMs - Signal band location

▶ **Trade-off at the IF Stage**



☹ The larger f_n the lower B_{af}

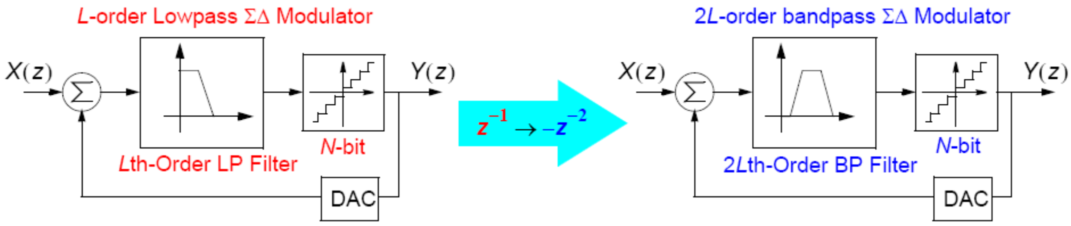
▶ **Trade-off at the RF Stage**

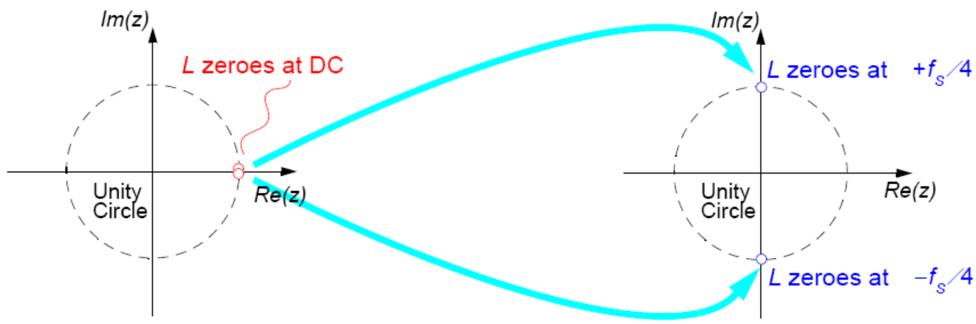


☹ The lower f_n the lower B_{irf}

- **Optimum location for** $f_n = f_s/4$ **(at the middle of Nyquist band)**
  - Forward path (analog) modulator filter realization can be simplified
  - Simplifies LP-to-BP transformation, $z^{-1} \rightarrow -z^{-2}$
  - Digital mixing to baseband is notoriously simplified: $\cos(2\pi f_{IF} n T_s) = 1, 0, -1, 0, 1, 0, \ldots$

DT-ΣΔMs: LP-to-BP Transformation Method



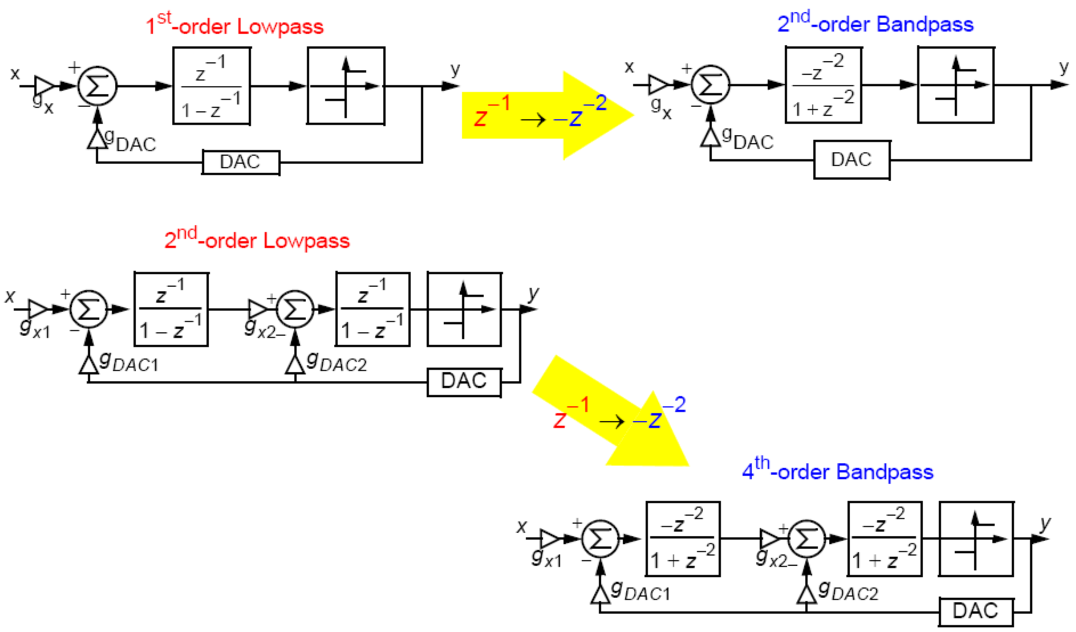


$$S_{TF} = z^{-L}$$
$$N_{TF} = (1-z^{-1})^L$$

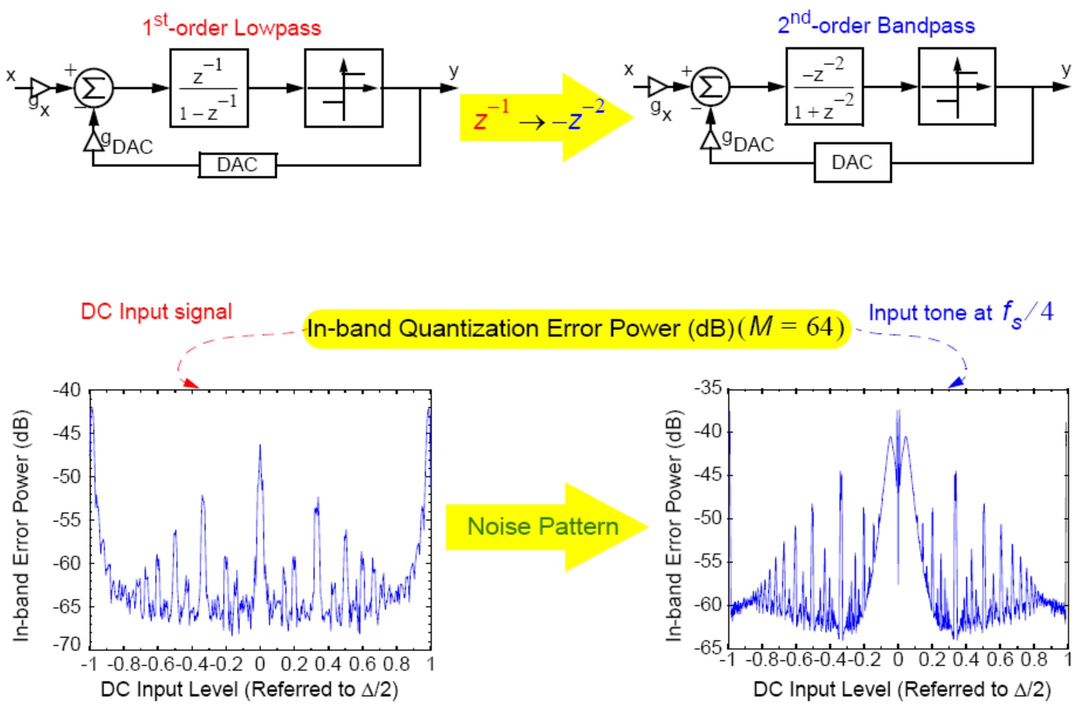
$$SNR_Q = \frac{12(2^B-1)^2(2L+1)M^{2L+1}}{8\pi^{2L}}$$

$$S_{TF} = (-z^{-2})^L$$
$$N_{TF} = (1+z^{-2})^L$$

DT-ΣΔMs: LP-to-BP Transformation Method

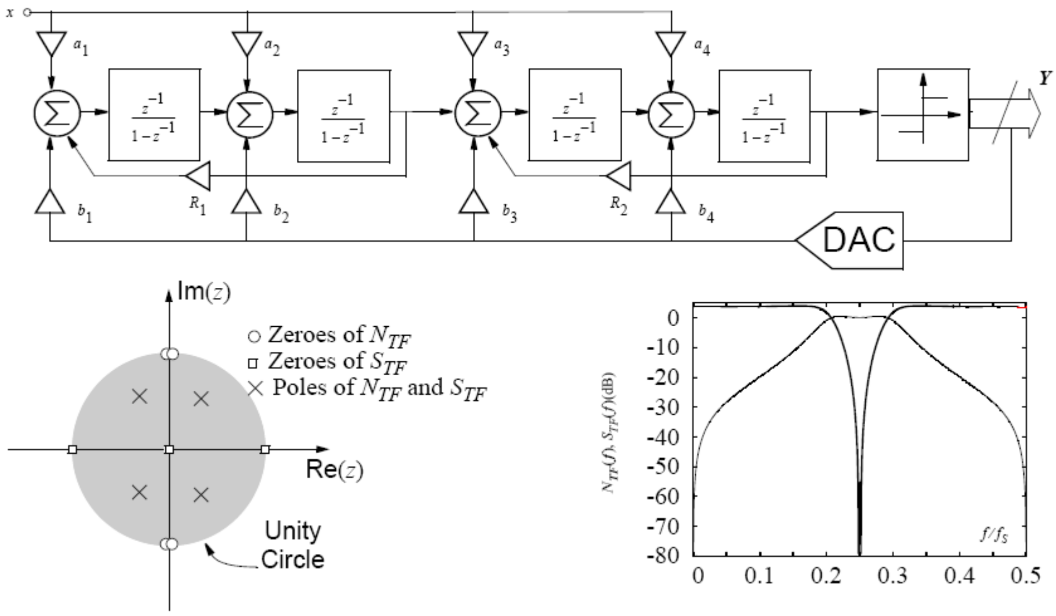


DT-ΣΔMs: LP-to-BP Transformation Method



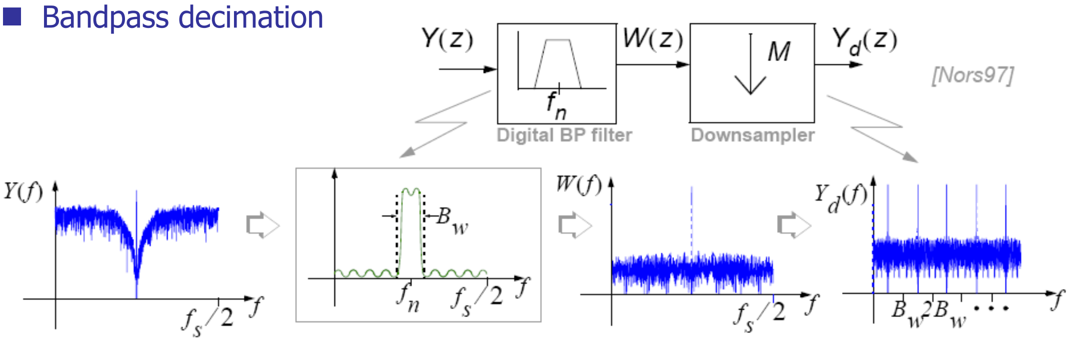
DT-ΣΔMs: Bandpass ΣΔ Modulators

- Other BP-ΣΔM architectures

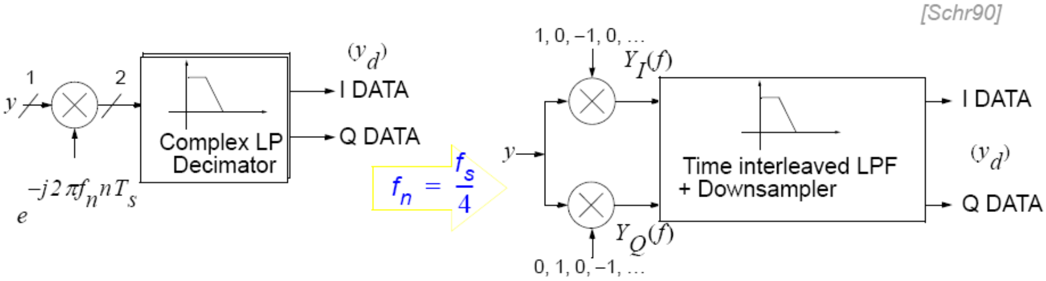


DT-ΣΔMs: Bandpass ΣΔ ADCs - Decimation

- Bandpass decimation

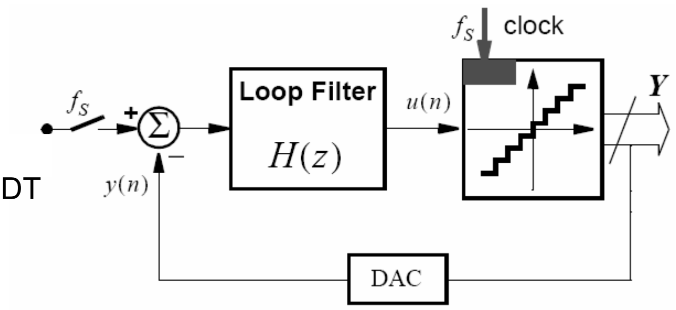


- Efficient decimation



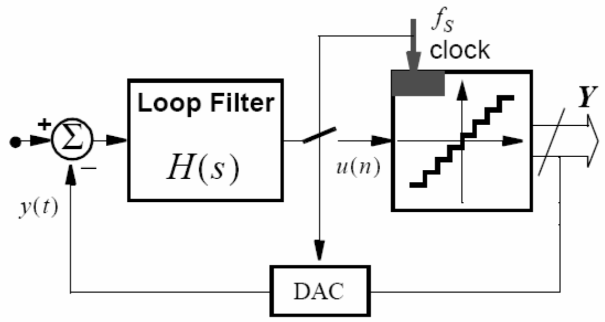
❒ Discrete-Time ΣΔMs

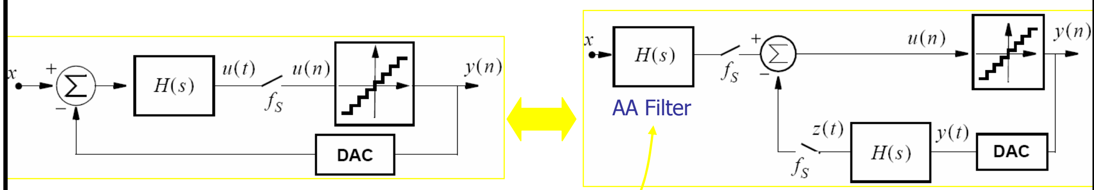
- DT loop filter
- All internal signals are DT
- Sampling at the input



❒ Continuous-Time ΣΔMs

- CT front (loop filter) part
- DT back (quantizer) part
- Sampling inside the loop





❒ Pros of CT-ΣΔMs

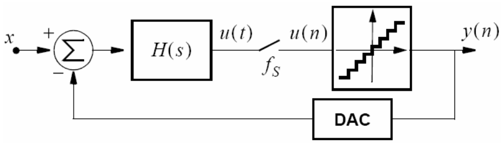
- Implicit anti-aliasing filter
- Less impact of sampling errors
- No input switches – potentially better for low-voltage supply
- No “settling” error at the loop filter circuitry
- Potentially larger operation speed with less power consumption
- No sampling of the noise at the input capacitors
- Reduced digital noise coupling

❒ Counters of CT-ΣΔMs

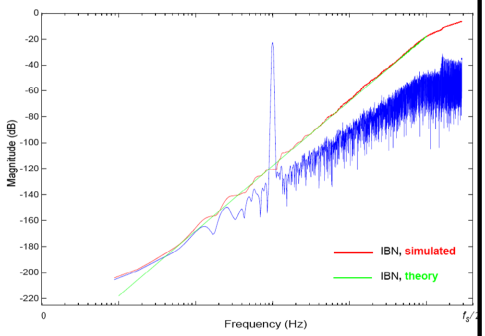
- Very involved dynamic due to the combination of non-linearity, CT and DT
- larger impact of circuit non-linearities
- Time constant tuning is needed for correct loop filtering
- Large sensitive to time uncertainty (“jitter”)

CT-ΣΔMs: Basic Concepts

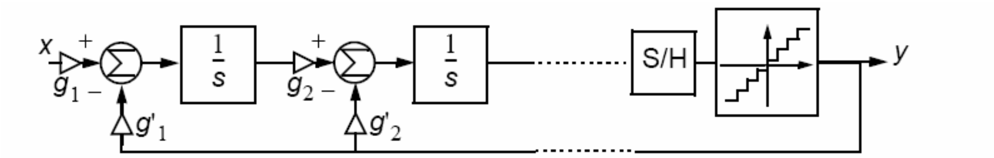
- Linear analysis of CT-ΣΔMs, assuming [Bree01]:
  - Linear model for the quantizer
  - DAC gain is unity in the signal bandwidth



$$Y(f) \cong \frac{H(f)}{1+H(f)} \cdot X(f) + \frac{1}{1+H(f)} \cdot E(f)$$



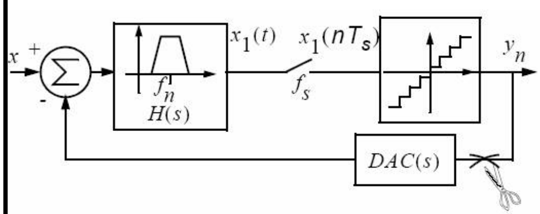
- Example: *L*th-order, *B*-bit single-loop architecture



$$Y(f) \cong \frac{g_1}{g'_1} \cdot X(f) + (2\pi j f \tau)^L \cdot E_q(f) \quad \Rightarrow \quad DR = \frac{3(2^B-1)^2(2L+1)M^{2L+1}}{2\pi^{2L}}$$

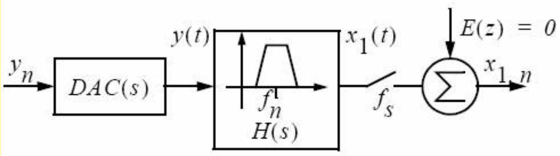
CT-ΣΔMs: Synthesis Methods

- DT-to-CT synthesis method: pulse invariant transformation (freq. domain)
  - Find an equivalent DT ΣΔM that fulfils the required specifications
  - Based on a DT-to-CT equivalence [Cher00]



DAC	$H(z)$	$H(s)$
NRZ	$\frac{z^{-1}\cdot(1-z^{-1})}{1+z^{-2}}$	$\frac{\omega_o \cdot s}{s^2+\omega_o^2}$
RZ	$\frac{\left(1-\frac{\sqrt{2}}{2}\right)\cdot z^{-1} - \left(\frac{\sqrt{2}}{2}\cdot z^{-2}\right)}{1+z^{-2}}$	
HRZ	$\frac{\frac{\sqrt{2}}{2}\cdot z^{-1} - \left(\left(1-\frac{\sqrt{2}}{2}\right)\cdot z^{-2}\right)}{1+z^{-2}}$	

Open-loop configuration

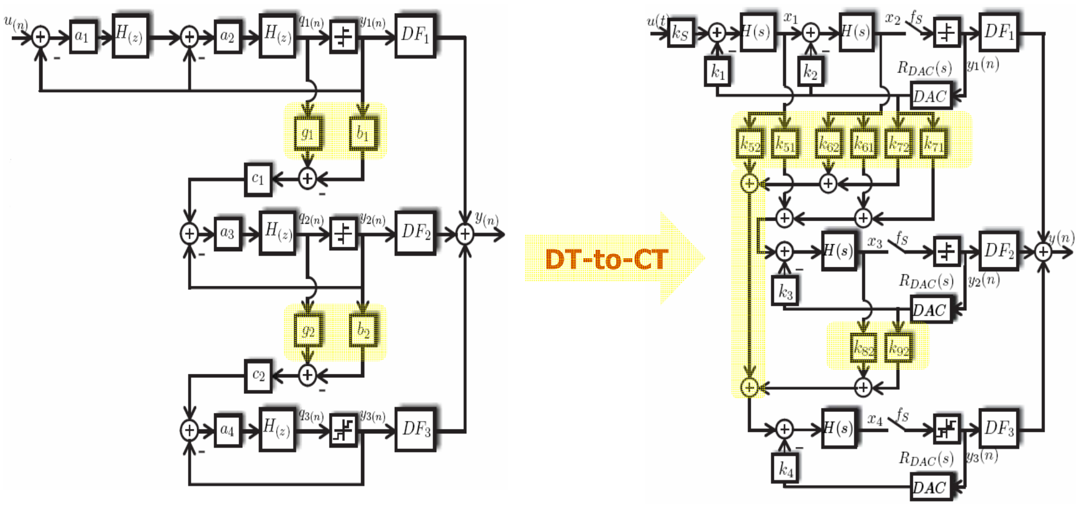


$$\frac{X_1(z)}{Y(z)} = Z\left[L^{-1}[DAC(s)H(s)]\big|_{t=nT_s}\right]$$

CT-ΣΔMs: Synthesis Methods

❒ Application of DT-to-CT method to cascade CT ΣΔMs

- Every state variable and DAC output must be connected to the integrator input of the ulterior stages in the cascade [Ortm01]
- Increases the number of analog components (transconductors and amplifiers)



CT-ΣΔMs: Synthesis Methods

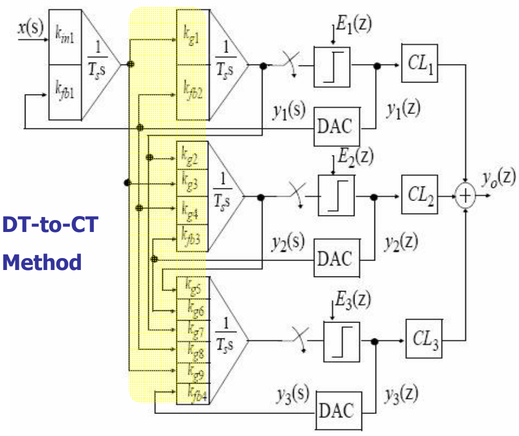
❒ Direct synthesis method [Bree01]

- Uses the desired NTF as a starting point, (as for the DT case)
- An Inverse Chevychev distribution of the NTF zeros has advantages in terms of SNR and stability

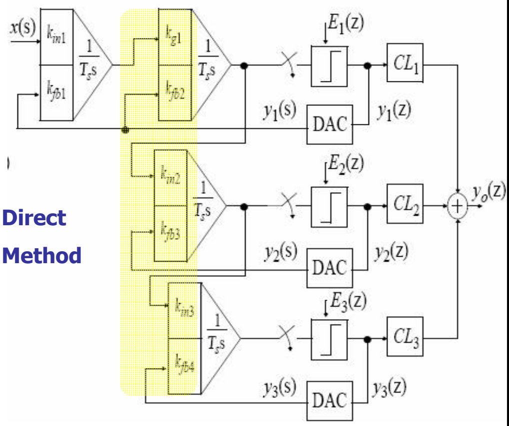
❒ Application to cascade architectures [Tort06]

- Optimum placement of poles/zeroes of the NTF
- Synthesis of both analog and digital part of the cascade CT ΣΔ Modulator
- Reduced number number of analog components

DT-to-CT Method

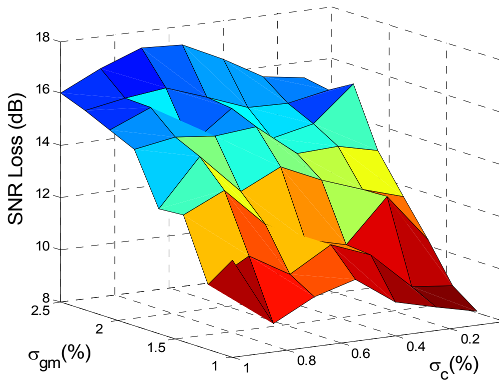


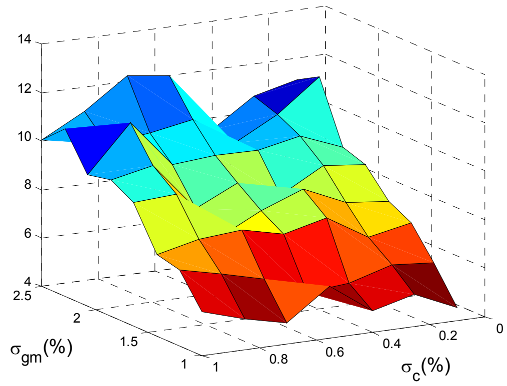
Direct Method



❒ Direct synthesis of cascade architectures (I) [Tort06]

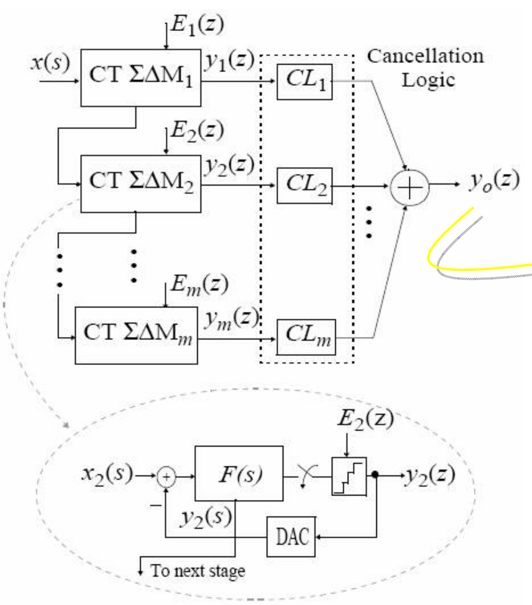
- Sensitivity to mismatch (gm,C)
- A 2-1-1 example





- DT-to-CT synthesis method
- Direct synthesis method

❒ Direct synthesis of cascade architectures (II) [Tort06]



$$y_o(z) = \sum_{k=1}^{m} y_k(z) CL_k(z)$$

$$y_k(z) = \frac{E_k(z) + \sum_{i=1}^{k-1} Z_{ik} y_i(z)}{1 - Z_{kk}}$$

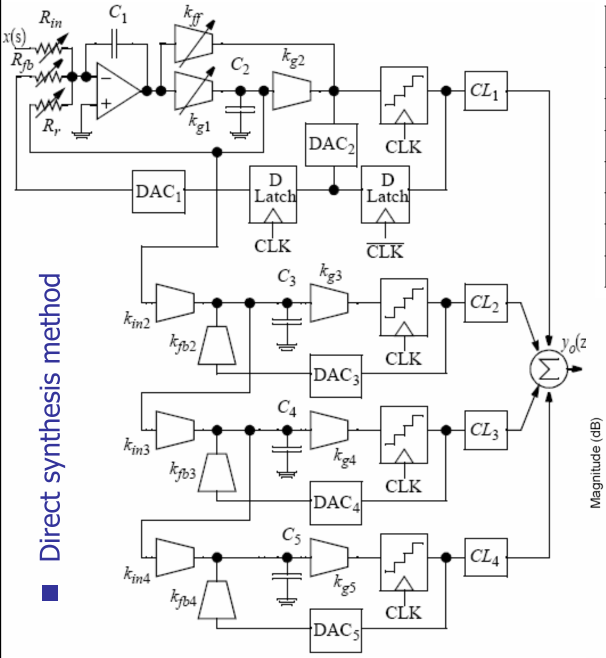
$$CL_k(z) = \frac{-Z_{km} CL_m}{1 - Z_{mm}}$$

$$\left[Z_{km} \equiv Z\left(L^{-1} (H_D F_{km}) \Big|_{nT_s} \right) \right]$$

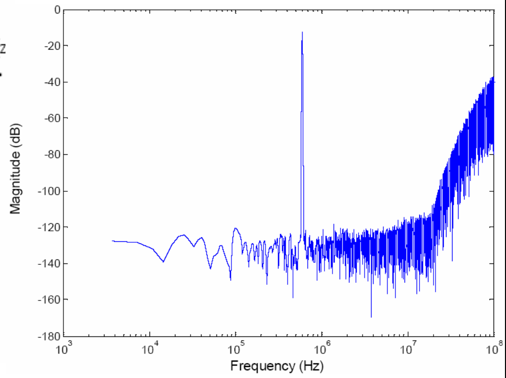
CT-ΣΔMs: Synthesis Methods

- A case study: A 12-bit@20MHz, 4-b, 2-1-1 CT ΣΔM for VDSL [Tort06]

- Direct synthesis method



Parameter	Value
R_{in},R_{fb}	1kΩ
R_r	2.9kΩ
$k_{g2}\cdots k_{g5}$	50µA/V
k_{g1}	500µA/V
k_{ff}	120µA/V
$k_{in2}\cdots k_{in4}$,$-k_{fb2}\cdots -k_{fb4}$	450µA/V
C_1	7.5pF
$C_2\cdots C_5$	1.875pF



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