Laboratory 2

Delta-Sigma Modulation

The Art of Oversampling, Noise Shaping and Averaging

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Abstract. Quantization noise will be investigated in unmodulated and ΔΣ modulated form. Mathematical roots to understand the matter are discussed before practical evidence is given. Quantization noise power within the baseband 0...f_B is proportional to Δ^2/OSR^{2L+1}, with Δ being the quantizer's minimum step, $OSR=f_s/2f_B$ the oversampling ratio, f_s the sampling frequency and L the modulator's order.

1 Introduction

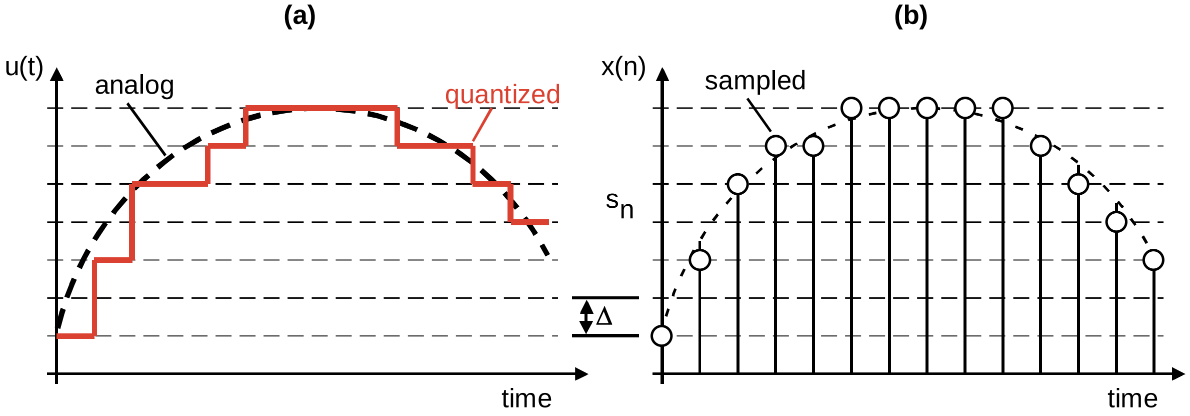


Fig. 1: (a) Analog and quantized signal, **(b)** sampled digital signal

Any signal real is noisy. Mostly we try to reduce digital noise by a higher bit-width of the processed numbers. The ΔΣ modulator goes the opposite direction: It translates a long bit-vector or an analog quantity into a fast data stream with few parallel bits. The data stream encodes the input value within its average. This requires quantization [1]-[4].

Quantization is inevitable for analog-to-digital (A/D) conversion. Quantization and quantization noise correspond to the mathematical process of rounding and round-off noise. The minimum step of quantization is termed Δ. It can be a bit, a voltage, a current, etc.

The power of noise shaping demonstrates Table 1. To improve the signal-to-noise ratio of the averaged signal by 60dB (= factor 1000 in voltage, 10^6 in power or ca. 8 bit signal width) an unmodulated signal must be $OSR=10^6$ times faster than required acc. to Nyquist. For a 1st order modulator an *OSR* of 126 is enough and for a 2nd order modulator an *OSR* of only 27! (PS: This calculation assumes that there are no other noise sources than quantization noise.)

Table 1: *OSR* is the oversampling ratio and *L* the modulator's order. Data was taken from [3].

SNR_{dB}	20 dB	40 dB	60 dB	80 dB	100 dB
OSR @ L=2	5	11	27	67	168
OSR @ L =1	6	28	126	578	2657
OSR @ L =0	100	10322	1,05E6	1,06E8	1,08E10

The theoretical part of ΔΣ modulation is relatively difficult [1]-[4]. But it is of high practical importance and a useful device to demonstrate numerous signal processing theorems.

The practical part of this laboratory is relatively easy and many results are more or less given. Some documents help to understand equipment [5]-[8] and control loops [9].

The organization of this laboratory is as follows:

Section 2	presents some very fundamental theoretical background on signal processing which most students should already know.
Section 3	presents particular statements on ΔΣ modulation.
Section 4	offers experimental verification.
Section 5	draws relevant conclusion.

2 Some Fundamentals on Signal Processing

2.1 Correlation

- Correlated signals depend on each other
- Uncorrelated signals do not depend on each other

Correlated signals sum in amplitude: $$y_{sum,corr} = x_1 + x_2 + x_3 + ... + x_N \quad (2.1)$$

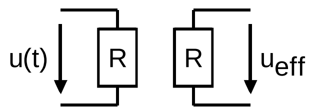
Uncorrelated signals sum in power: $$y_{sum,uncorr} = \sqrt{x_1^2 + x_2^2 + x_3^2 + ... + x_N^2} \quad (2.2)$$

Different frequencies are always uncorrelated. (2.3)

2.2 Root-Mean-Square Computation of Effective Amplitudes

2.2.1 Effective Voltages and Currents and their RMS Computation

Fig. 2.2.1: The effective DC voltage $u_{eff}\{u(t)\}$ dissipates the same power at a resistor R like the $u(t)$.



The effective DC value of voltage $u(t)$ is termed $u_{eff}\{u(t)\}$. It dissipates the same power at a resistor R as the $u(t)$. Good voltmeters either heat a reference resistor to the same temperature as $u(t)$ or compute u_{eff} by the **Root Mean Square** (rms) method: Square a signal, average it over time (mean) and then compute the square root of it.

Time continuous: $$u_{eff} = \sqrt{\overline{u^2}} = \sqrt{\frac{1}{T}\int_0^T u^2(t)dt} \quad (2.4)$$

Time discrete for samples x_i, i=1...N: $$x_{eff} = \sqrt{\overline{x^2}} = \sqrt{\frac{1}{N}\sum_{i=1}^{N} x_i^2} \quad (2.5)$$

Note: The frequently observed expression $\overline{u}^2$ is 0 for an AC signal, because its mean value is 0 and 0^2=0, too. Correct is averaging after squaring: $u_{eff}^2 = \overline{u^2}$.

2.2.2 Effective Values of Some Particular Waveforms

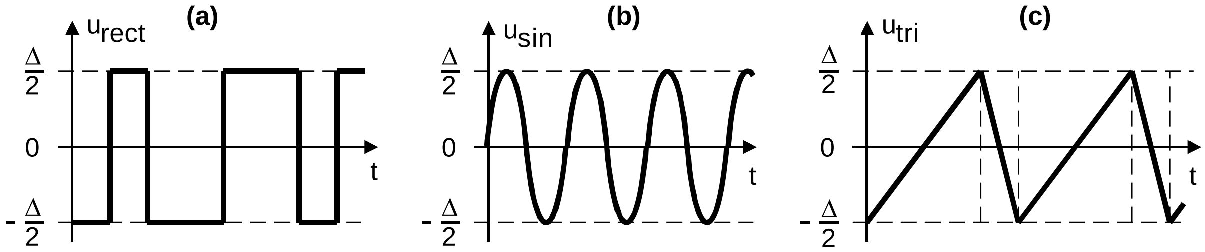


Fig. 2.2.2: Particular waveforms: **(a)** rectangular, **(b)** sinusoidal, **(c)** triangular.

Fig. 2.2.2 shows (a) a rectangular, (b) a sinusoidal und (c) a triangular signal oscillating between the values Δ/2 and -Δ/2. Its total power at 1Ω and its effective voltage are given by

Rectangular: $$\overline{u_{rect}^2} = \frac{(\Delta/2)^2}{1} = \frac{\Delta^2}{4} \quad \leftrightarrow \quad u_{rect,eff} = \frac{\Delta}{2} = \frac{\Delta/2}{\sqrt{1}} \;, \qquad (2.6)$$

Sinusoidal: $$\overline{u_{\sin}^2} = \frac{(\Delta/2)^2}{2} = \frac{\Delta^2}{8} \quad \leftrightarrow \quad u_{\sin,eff} = \frac{\Delta}{\sqrt{8}} = \frac{\Delta/2}{\sqrt{2}} \;, \qquad (2.7)$$

Triangular: $$\overline{u_{tri}^2} = \frac{(\Delta/2)^2}{3} = \frac{\Delta^2}{12} \quad \leftrightarrow \quad u_{tri,eff} = \frac{\Delta}{\sqrt{12}} = \frac{\Delta/2}{\sqrt{3}} \;. \qquad (2.8)$$

Adding an offset: The offset voltage is frequency 0Hz. Adding in power delivers

Combined: $$\overline{u_{comb}^2} = \overline{u_{offset}^2} + \overline{u_{xxx}^2} \quad \leftrightarrow \quad u_{comb,eff} = \sqrt{u_{offset}^2 + u_{xxx,eff}^2} \qquad (2.9)$$

with *xxx* standing for *rect*, *sin*, *tri* or others.

2.3 Estimating the Total Power of Quantization Noise

Quantization is the process of expressing quantities as integral multiples of a minimum step, also termed Δ in signal processing. Quantization corresponds to the mathematical process of rounding and the quantization error e_q corresponds to the mathematical round-off error.

2.3.1 Worst Case: Signal Power << Quantization Noise Power

Fig. 2.3.1: Averaged analog signal $u(t)$ is tiny compared to the quantizer's Δ/2.

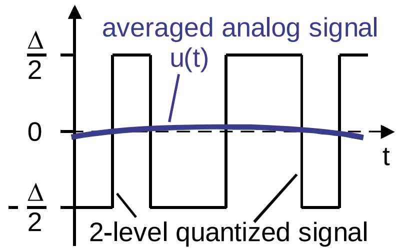


Fig. 2.3.1 illustrates the worst case scenario for a 2-level (1-bit) quantizer and a signal $u(t)$ obtained from the bit-stream by averaging. This case is of very high practical relevance!

Quantization noise is obtained from $$\boxed{e_q(t) = u_{ideal}(t) - u_{quantized}(t)} \;. \qquad (2.10)$$

It is also clear that in the worst-case (2.10) becomes in the limit of $u(t) \to 0$

If $|u(t)| << \Delta/2 \quad \to$ $$e_q(t) \approx -u_{quantized}(t) \;. \qquad (2.11)$$

The total noise power and effective noise voltage is then given by (2.6):

Worst Case: $|u(t)/\Delta| \to 0$: $$\boxed{\overline{e_{n,q,worst}^2} \xrightarrow{|u(t)/\Delta| \to 0} \frac{\Delta^2}{4}} \quad \leftrightarrow \quad \boxed{e_{n,q,worst,eff} \xrightarrow{|u(t)/\Delta| \to 0} \frac{\Delta}{2}} \;. \qquad (2.12)$$

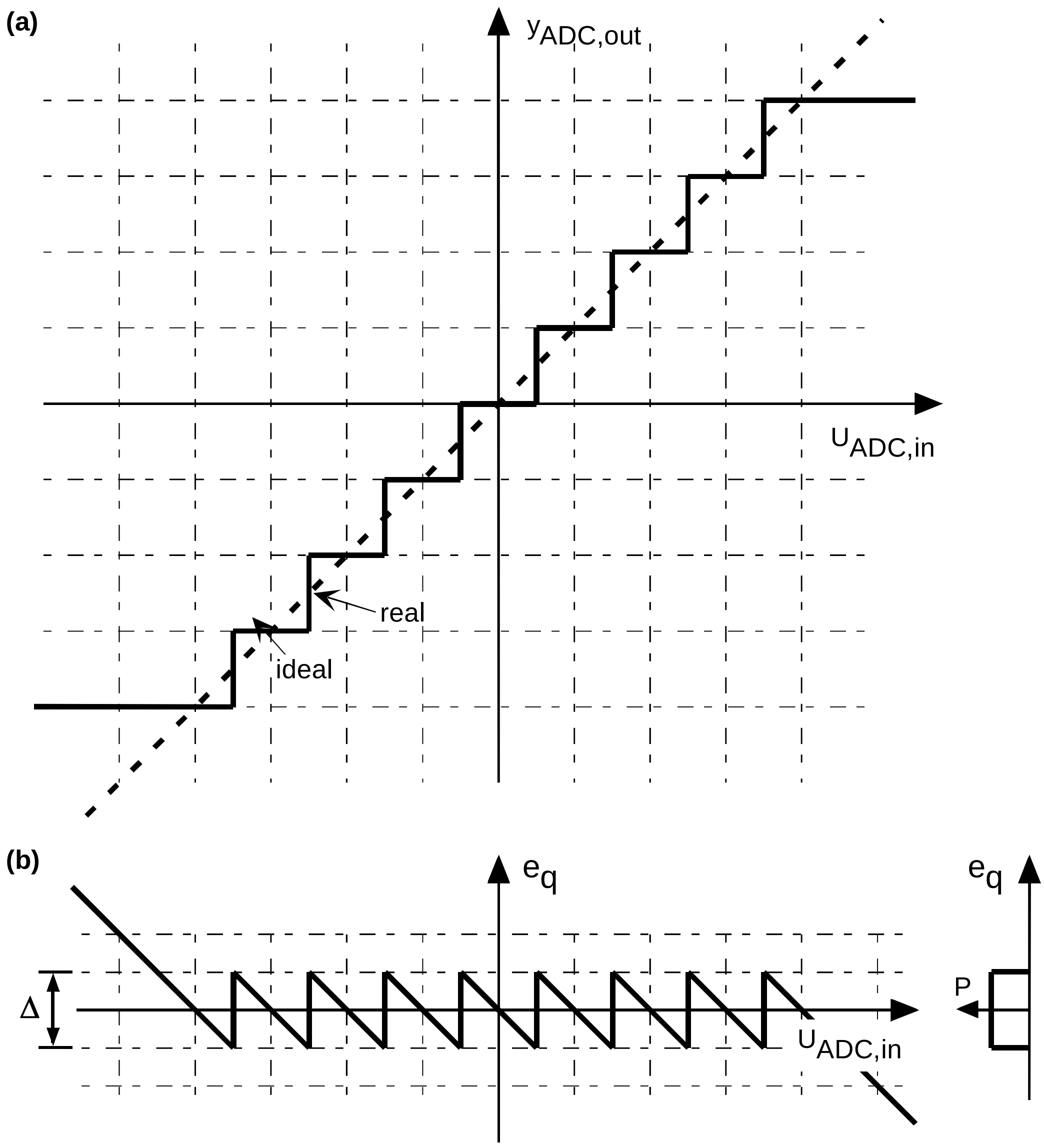
2.3.2 Best Case: Quantization Noise is Equidistributed Over Interval Δ

Fig. 2.3.2: Illustrating the best-case quantization situation

(a) ideal and real charac-teristics of a quan-tizer

(b) left : quanti-zation noise e_q .

(c) right: e_q equidis-tributed over Δ.



From signal processing literature [1]-[4] we know (without proof) that

- If the input signal is sufficiently busy (no DC!) and (2.13)
- the quantization error $e_q(t)$ is equidistributed over the interval Δ (2.14)

then it can be shown [1]-[4], that the total noise power is described by (2.8), the effective value of a triangular oscillation:

Best Case: e_q equidistributed over Δ : $\boxed{\overline{e_{n,q,best}^2} = \frac{\Delta^2}{12}} \leftrightarrow \boxed{e_{n,q,best,eff} = \frac{\Delta}{2\sqrt{3}}}$. (2.15)

The reality will be somewhere between worst and best case, i.e. $\boxed{\frac{\Delta^2}{12} \le e_{n,q}^2 \le \frac{\Delta^2}{4}}$. (2.16)

2.4 Bel and Dezibel

In honor of Graham Bell a factor 10 in signal power is termed a Bel, and 1 B = 10 dB, just as 1 m = 10 dm or 1 liter = 10 dl. As power corresponds to square of amplitude ($p=u^2/R=i^2\cdot R$) and $\log(x^2)=2\cdot\log(x)$ we get

$$\text{Signal-Ratio} = \log_{10}\frac{p_2}{p_1}B = 10\log_{10}\frac{p_2}{p_1}dB = 20\log_{10}\frac{u_2}{u_1}dB = 20\log_{10}\frac{i_2}{i_1}dB \quad . \tag{2.6}$$

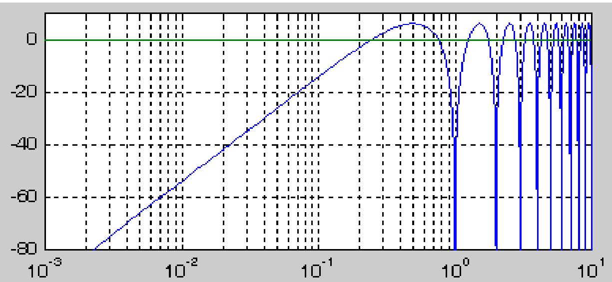
Question: 10dB is what factor in signal power? 10dB is what factor in effective voltage?

1 B = 10dB is a factor 10 in power → a factor $\sqrt{10}$ in voltage.

2.5 Estimating Signal Averages on Logarithmic Scale

Printing averaged values on logarithmic axis delivers values in the upper range of the signal.

- The average value of signal $2\sin^2(\pi F)$ shown in Fig. 2.5(a) is 1 or 0dB.
- The average value of signal $2\sin^2(\pi F)\ 1/(1+\pi F)$ in Fig. 2.5(b) is 0.1061 or –19.5dB.



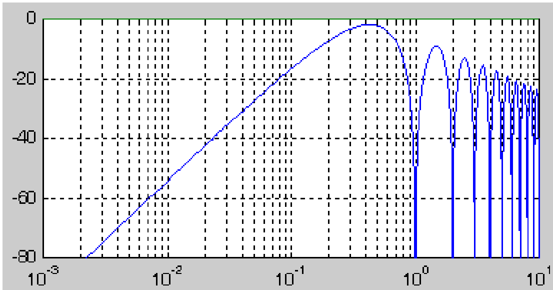


Fig. 2.5: **(a)** Bode-plot of $2\sin^2(\pi F)$ **(b)** Bode-plot of $2\sin^2(\pi F)\ 1/(1+\pi F)$

2.6 Time-Continuous versus Time-Discrete Models

- Time-continuous systems are modeled in s (, i.e. by Laplace transformation).
- Time-discrete systems are modeled in $z=e^{sT}$ with $T=1/f_s$ and f_s sampling frequency.
- In both cases – time-continuous and time-discrete – transfer functions are computed by setting $s=j\omega$.

In signal processing we are rather interested in the relative frequencies $F=f/f_s$ and $\Omega = \omega / f_s$ than in the absolute frequencies f and $\omega=2\pi f$.

If $H(z)$ is a transfer functions in z, then $|H(z)|$ is periodic in f_s. It is enough to know $|H(z)|$ in the range f=0...½ f_s corresponding to F=0...½ and Ω=0...π. the rest is clear from symmetry considerations.

In the time-discrete domain we cannot represent signals at $f > f_s/2$. Therefore we assume the total quantization noise to be distributed within the interval $f = 0 \ldots f_s/2$.

In the time-continuous world we definitively have frequencies $f > f_s/2$.

Ideal time-continuous differentiator with transit frequency ω_T: $$H_{c,\text{int}}(j\omega) = \frac{j\omega}{\omega_T} = \frac{jf}{f_T} \quad (2.7)$$

Ideal time-discrete differentiator with $F=fT=f/f_S$: $$H_{d,\text{int}}(z) = z^{\frac{1}{2}} - z^{-\frac{1}{2}} = 2j\sin(\pi F) \quad (2.8)$$

2.7 Frequency Range to be Measured for Sampled Signals

Fig. 2.6: *NTF* of 2nd order ΔΣ modulator with a measured frequency range up to F=5 corresponding to $f=5f_s$, with sampling frequency f=100KHz.

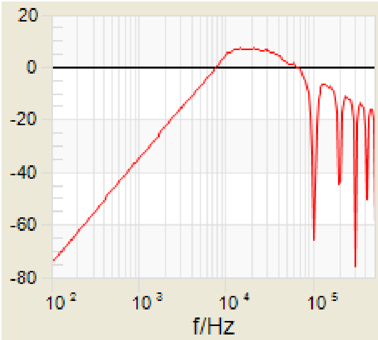


Fig. 4.2.6(d) illustrates the measured NTF of the 2nd order ΔΣ modulator in the range f=10Hz...50KHz for f_s=100KHz. Fig. 2.6 illustrates what happens if we extend the frequency range by a factor 10. The measurement interval was extended to 500KHZ corresponding to $5f_s$ or F=5. With constant C the measured function should be

$$|H(F)| = C\sin^2(\pi F)$$

if the output samples were Dirac pulses. Fig. 2.6. shows 5 maxima of the $\sin^2$ - function. Caused by the finite pulse width τ we see a –20dB/dec weighting for high frequencies, which can be approximated as

$$|H(F)| = C\sin^2(\pi F) \cdot 1/(1+\pi F f_s \cdot \tau)$$

Like in most cases the pulse width here is $\tau = 1/f_s$, so that the approximation becomes

$$|H(F)| = C\sin^2(\pi F) \cdot 1/(1+\pi F).$$

As $\sin(x)\approx x \;\rightarrow\; \sin^2(x)\approx x^2$ for $x<<1$ we see a slope of 40dB/dec in Fig. 2.6 for f<<50KHz.

2.8 Check your Knowledge

How do you sum correlated signals? How do you sum uncorrelated signal? How do you sum signals with different frequencies? What are the formulas to compute effective signal values by the rms method, for time-continuous and time-discrete signals? What are effective values of rectangular, sinusoidal and triangular waveforms, with and without offset? How much is 10dB as factor in amplitude and in power? How do determine the mean value of signals on logarithmic axis? What frequency range do we observe for sampled signals?

3 Theory of Delta-Sigma Modulation

3.1 Delta-Sigma Modulation

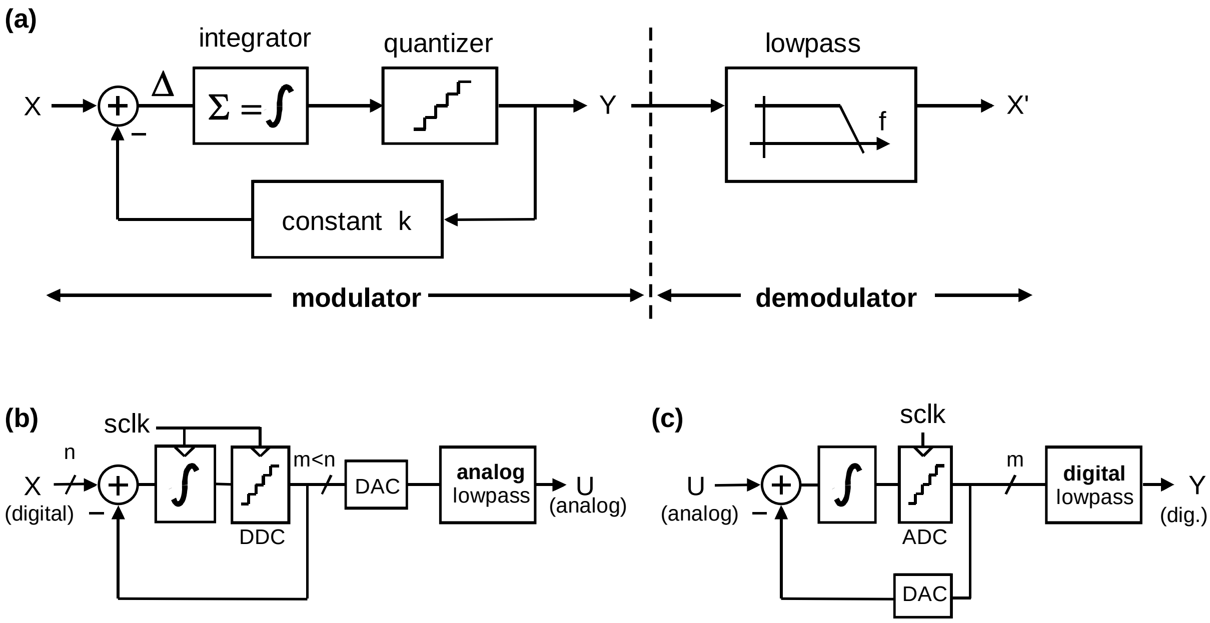


Fig. 3.1: **(a)** Basic principle of ΔΣ modulation and demodulation **(b)** ΔΣ DAC, **(c)** ΔΣ ADC.

Fig. 3.1(a) illustrates the basic principle of ΔΣ modulation: A quantizer reduces the resolution of the input signal. The errors which occur are memorized by the integrator and the average output error is minimized by the feedback loop. For high loop amplification the behavior of the modulator can be approximated by k^{-1} with k being constant. To allow for averaging the output samples the modulator must sample high above the Nyquist rate, i.e. at high oversampling ratio (OSR). Demodulation is averaging the high-data rate output samples. This is done by a lowpass, which can be seen as a weighted averager.

Fig. 3.1(b) illustrates how to use a ΔΣ modulator as digital-to-analog converter (DAC): A digital-to-digital converter (DDC) reduces the sampling rate and drives a high quality, low-resolution (often one bit) DAC. After averaging by an analog lowpass we have an accurate analog signal. The DDC is easiest realized by using some of the most significant bits of the integrator's state vector.

Fig. 3.1(c) illustrates how to use a ΔΣ modulator as analog-to-digital converter (ADC): We use an ADC in the forward network and a DAC in the feedback branch, both DAC and ADC with low resolution (often one bit). After averaging by a digital lowpass we have an accurate digital signal.

Check your knowledge:
What is the basic topology of a ΔΣ modulator?
What is the more particular topology of a ΔΣ modulator for D/A conversion?
What is the more particular topology of a ΔΣ modulator for A/D conversion?

3.2 The Building Blocks

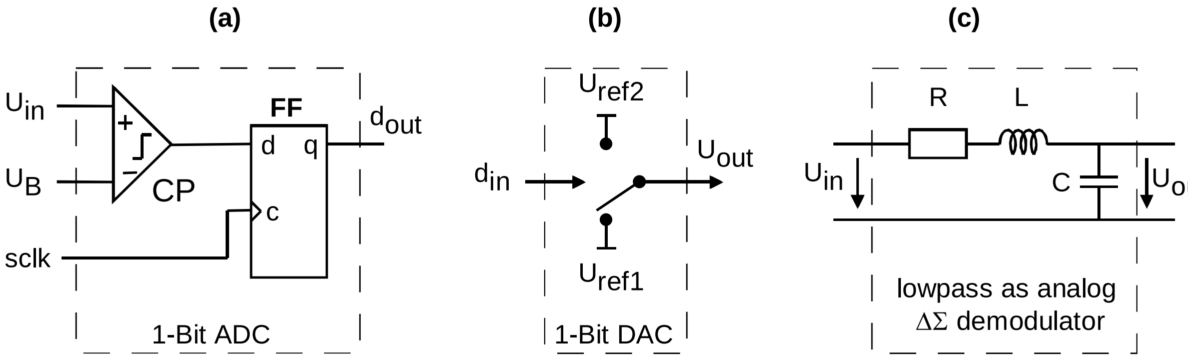


Fig. 3.2: **(a)** 2-level ADC. **(b)** 2-level DAC, **(c)** Lowpass as ΔΣ demodulator.

The difference signal symbolized by Δ (which is not the minimum step Δ!) and the integrator by Σ are the name-giving building blocks of the ΔΣ modulator. The other two important things are the quantizer and the fact, that the transfer function of the feedback branch is constant over frequency. The integrator realizes high loop gain for frequencies low compared to the sampling frequency f_s. This is important because signal- and noise transfer functions

$$STF = \frac{Y}{X} = \frac{A}{1+kA} \xrightarrow{|kA|>>1} k^{-1} \ , \qquad (3.1)$$

$$NTF = \frac{1}{1+kA} \xrightarrow{|kA|\to\infty} 0 \qquad (3.2)$$

require the condition |kA|>>1 to remove the forward network A out of the formula [1]-[4]. The STF in (3.1) now depends on the feedback branch k only. As this is a DAC in Fig. 3.1(c), we can realize a good ΔΣ-ADC from a poor ADC and a good DAC. As good DACs are significantly easier to realize than good ADCs this modulator transfers the quality properties of the DAC to the ΔΣ ADC. (Note that k^{-1} is an ADC when k is a DAC.)

Fig. 3.2(a) illustrates a 2-level quantizer making the circuit simple and cheap. A/D conversion is a two-dimensional process by discretization of both value und time, realized by comparator *CP* and flipflop *FF*, respectively, in this laboratory.

Fig. 3.2(b) illustrates a 2-level DAC, often realized as flipflop using its output transistors as switches. It is simple, cheap, accurate and energy-efficient.

- Negative: Quantization is extremely rough.
- Strong dependence on the signal integrity of U_{ref1}, U_{ref2}; in case of a FF: V_{DD} and *gnd*.
+ DAC is 100% linear, as a line through 2 points is straight. More levels by → DEM [1], [2].
+ Cheap: Very high energy efficiency obtainable, no need for analog devices.

Fig. 3.2(c) shows how easy and cheap an analog ΔΣ demodulator can be realized as lowpass. Positive: Such a lowpass can always deliver an accurate average of U_{in}, independent from tolerances of R, L, C. Precondition is that the data rate of U_{in} is sufficiently higher than the lowpasses cutoff frequency. Again, we have trade-off between accuracy and speed. Accurate devices are a problem, but high-speed, time-precise clock signals are available in electronics.

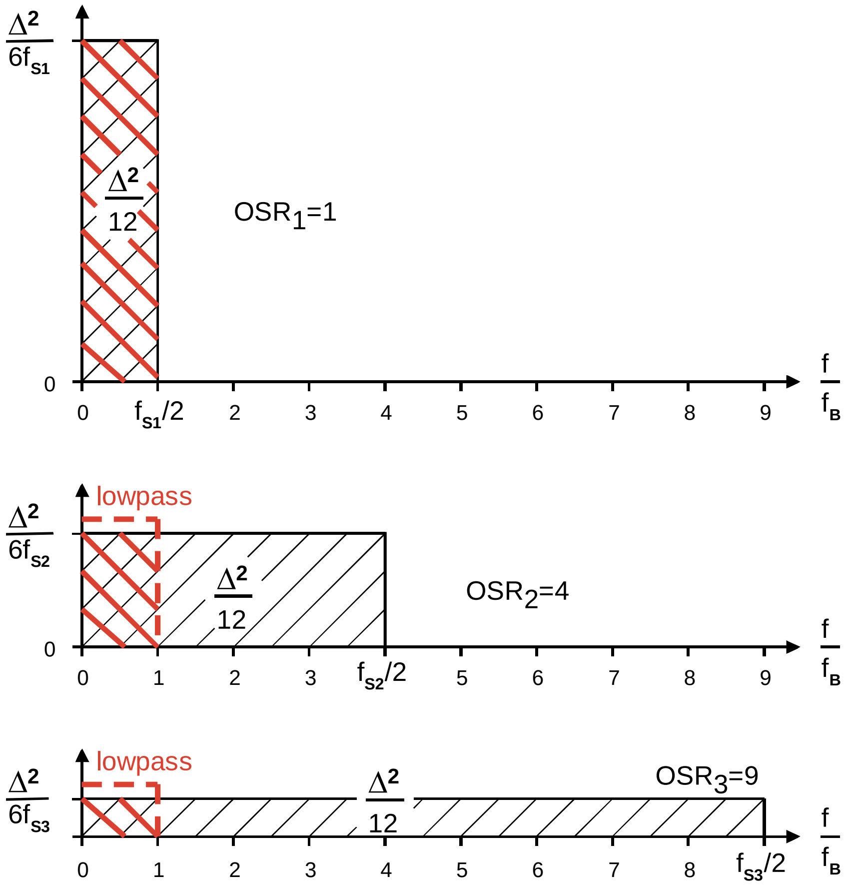
3.3 White Quantization Noise (0^{th} Order Noise Shaping)

Fig. 3.3:

(a) no oversampling

(b) Using an OSR=4 and an ideal lowpass at f_B.

(c) Using an OSR=9 and an ideal lowpass at f_B.



A ΔΣ modulator is a quantizer operated by a control loop to optimize the quantization noise profile. Understanding the process of quantization is essential!

Definitions:

- Sampling rate: $\boxed{\mathbf{f_s}}$.
- Baseband: $\boxed{\mathbf{0...f_B}}$.
- Oversampling ratio: $\boxed{\mathbf{OSR = f_s/2f_B}}$. (3.4)

Thesis: If the input signal is sufficiently busy the total noise power is distributed with constant power density over the frequency range 0...f_s/2. As uncorrelated signals sum in power and different frequencies are always uncorrelated, it is noise power (not noise voltage) that equidistributes over the frequency range 0...f_s/2.

Fig. 3.3 illustrates these statements. A signal within the baseband 0...f_B is sampled with same Δ but different sampling frequencies f_{s1}, f_{s2}, f_{s3}:

(a) $f_{s1}=2f_B$ has no oversampling: $OSR_1=f_{s1}/2f_B=1$.

(b) $f_{s2}=8f_B \rightarrow OSR_2=f_{s2}/2f_B=4$. Noise power (voltage) within baseband lower by factor 4 (2).

(c) $f_{s2}=18f_B \rightarrow OSR_2=f_{s2}/2f_B=9$. Noise power (voltage) within baseband lower by factor 9 (3).

Taking advantage of the $1/f_s$-reduction in baseband-noise power a lowpass with cutoff-frequency f_B is required. The red dashed lines in Fig. 3.3 show ideal lowpasses with cutoff-frequency f_B.

We matter we have to check by measurement:

- Quantization noise **power** density (in V^2/Hz) is proportional to Δ^2/f_s. (3.5)
- Effective quantization noise **voltage** density (in V/$\sqrt{Hz}$) is proportional to $\Delta/\sqrt{f_s}$. (3.6)

(In data sheets we often find the acronym *rtHz* for $\sqrt{Hz}$.)

3.4 First Order ΔΣ Modulator (1st Order Noise Shaping)

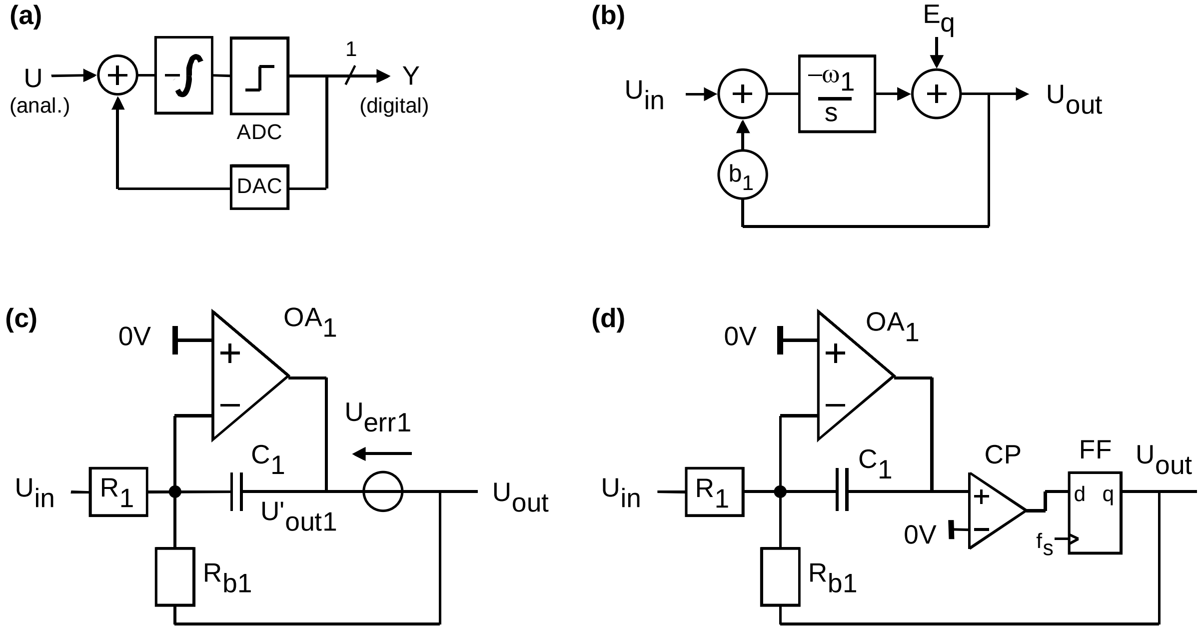


Fig. 3.4-1: **(a)** High-level schematics, **(b)** depicted for modeling, **(c)** OpAmp realization with $\omega_1=1/R_1C_1$, $b_1=R_1/R_{b1}$ and error source U_{err1}, **(d)** U_{err1} replaced by a 2-level quantizer.

While the signal transfer function (STF) is flat, the noise transfer function (NTF) shapes the quantization noise according to *1 / integrator = differentiator*. The transfer function of an ideal time-continuous differentiator – modeled in the *s*-domain – is

$$|H_{s,diff}(f)| = f/f_T,$$

with transit frequency f_T. Due to the clocked quantizer we have a time-discrete situation here. Consequently, the transfer function of a differentiator – modeled in the *z*-domain – becomes

$$|H_{z,diff}(F)| = C_1 \cdot \sin(\pi F),$$

with C_1 being a constant and $F=f/f_s$. As $\sin(x) \approx x$ for $x \ll 1$ we have a 20dB/dec slope in the Bode diagram for small F.

Fig. 3.4-2(a) sketches how quantization noise within the baseband (i.e. f=0...f_B corresponding to F=0...F_B with F_B=f_B/f_s) is reduced compared to a situation without noise shaping.

Fig. 3.4-2(b) illustrates what happens, when the order of the demodulating lowpass filter is equal to the order of the modulator (given by the number of its integrators): The noise power distribution above the filter's cutoff frequency becomes constant rather than being cut off by the filter.

Fig. 3.4-2(c) sketches the situation with the inevitable noise floor, because there are several noise sources other than the quantizer, too. The noise in the baseband is increased compared to Fig. part (a). Red dashed lines symbolize an ideal lowpass filter.

According to the theory and in absence of other noise the noise power in the baseband for the 1st order modulator is proportional $1/OSR^3$, → eff. noise voltage is proportional $1/OSR^{1.5}$.

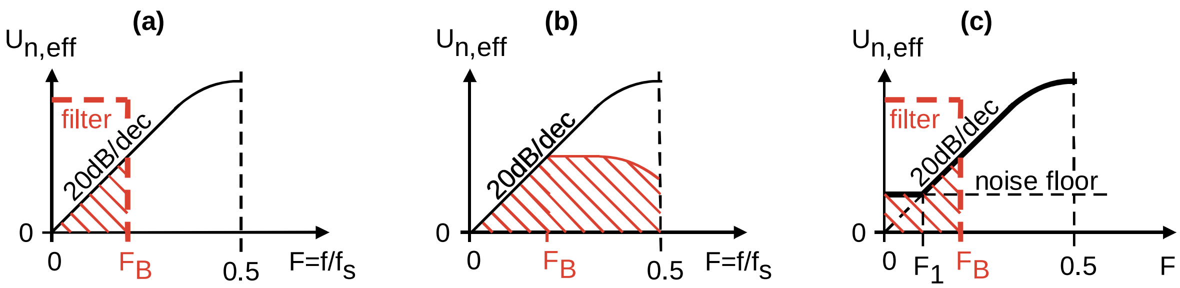


Fig. 3.4-2: **(a)** ideal situation, **(b)** lowpass-order=modulator-order, **(c)** with noise floor.

3.5 Second Order ΔΣ Modulator (2nd Order Noise Shaping)

According to the theory and in absence of other noise the noise power in the baseband for the 2nd order modulator is proportional $1/OSR^5$, → eff. noise voltage is proportional $1/OSR^{2.5}$.

While the signal transfer function (STF) is flat, the noise transfer function (NTF) shapes the quantization noise according to *1 / integrator² = differentiator²*. The transfer function of a time-continuous second order differentiator – modeled in the *s*-domain – is

$$|H_{s,diff}(f)| = (f\,/\,f_T)^2.$$

with transit frequency f_T. Due to the clocked quantizer we have a time-discrete situation here. Consequently, the transfer function of a second order differentiator – modeled in the *z*-domain – becomes

$$|H_{z,diff}(F)| = C_2 \cdot \sin^2(\pi F),$$

with C_2 constant, $F=f/f_s$. As $\sin^2(x) \approx x^2$ for x<<1 → 40dB/dec in the Bode diagram for F<<1.

Fig. 3.5-1 illustrates the 2nd order modulator to be used in this laboratory. Figure part **(a)** sketches the fundamental idea, **(b)** prepares it for modeling and **(c)** details the circuit.

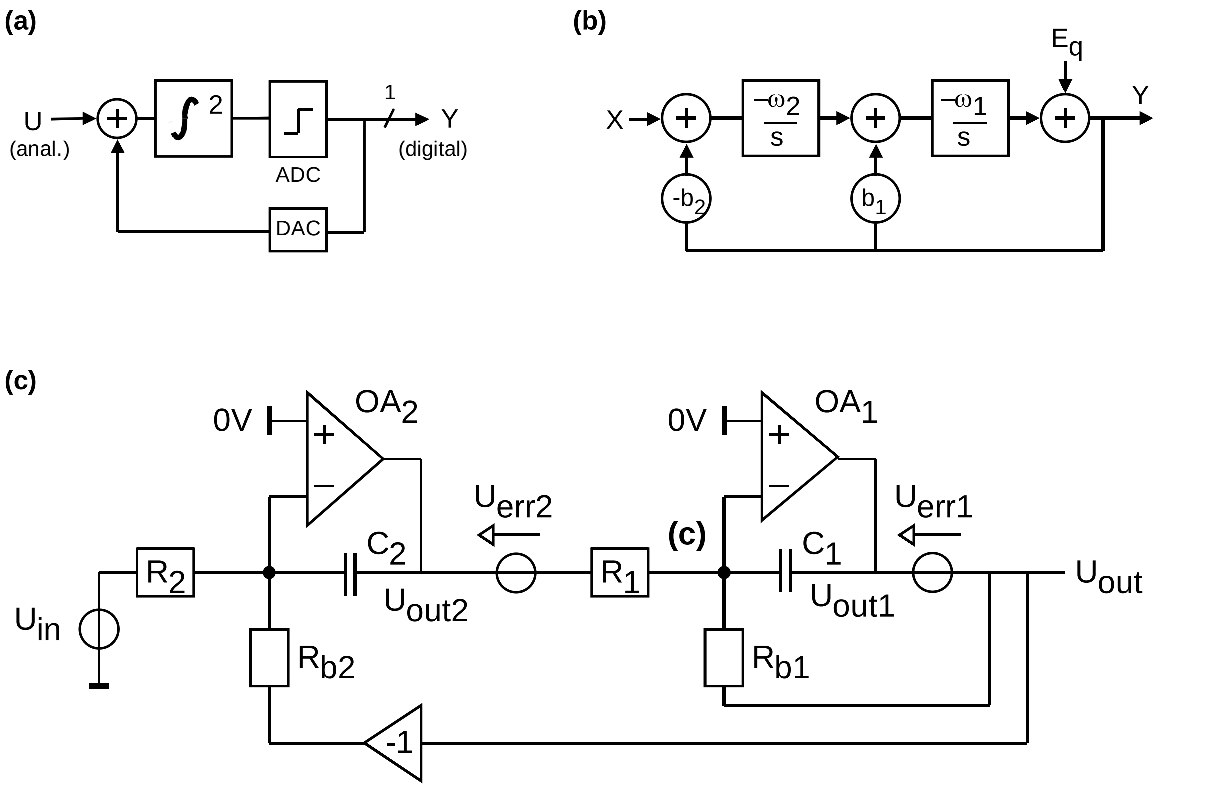


Fig. 3.5-1: **(a)** High-level schematics, **(b)** depicted for modeling, **(c)** realization with $\omega_1=1/R_1C_1$, $\omega_2=1/R_2C_2$, $b_1=R_1/R_{b1}$, $b_2=R_2/R_{b2}$ and error source U_{err1}, which will be replaced by the 2-level quantizer to obtain the ΔΣ modulator.

Fig. 3.5-2(a) sketches how quantization noise within the baseband (i.e. f=0...f_B corresponding to F=0...F_B with $F_B=f_B/f_s$) is reduced compared to a situation without noise shaping.

Fig. 3.5-2 (b) sketches the situation with the inevitable noise floor, because there are several noise sources other than the quantizer, too. The noise in the baseband is increased compared to Fig. part (a) as illustrated with the ideal lowpass.

Fig. 3.5-2 (c) illustrates overloading: A second order ΔΣ modulator will always try to do some two-level jumps. If this is not possible, as for the 2-level quantizer, we say that the modulator is overloaded. Higher order modulators will try to perform N-level jumps (with N increasing with the modulator's order) and tend fail in convergence in case of overloading [1].

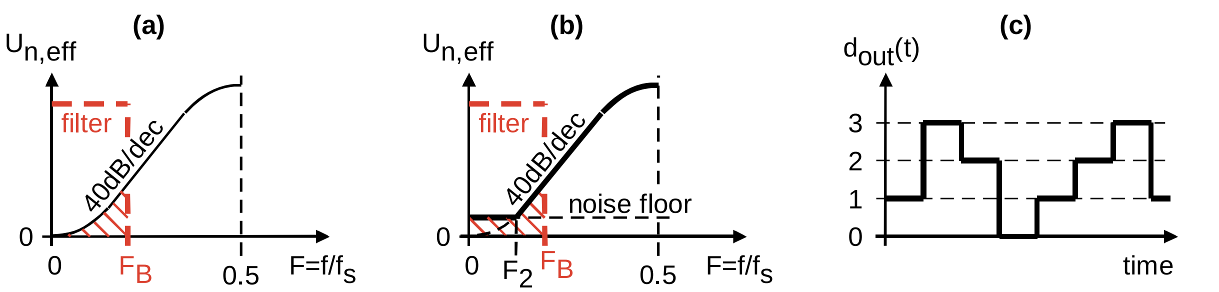
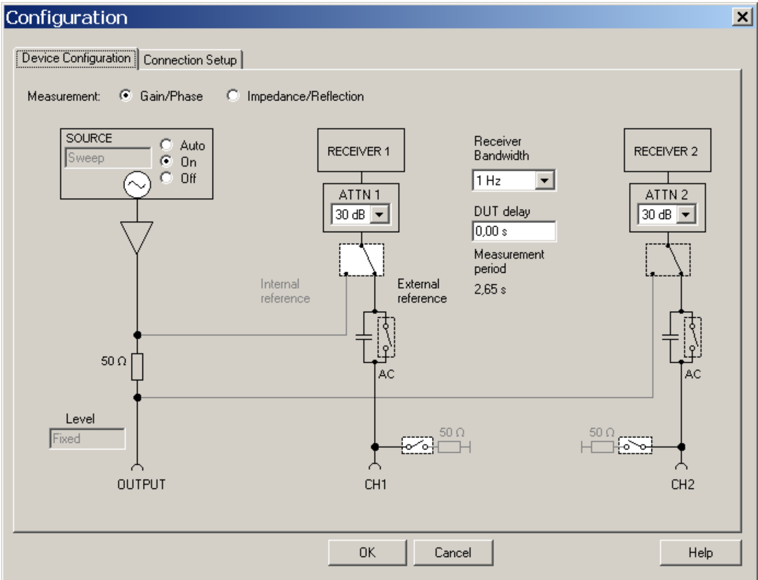


Fig. 3.5-2: **(a)** ideal situation, **(b)** situation with noise floor, **(c)** non-overloaded output behav.

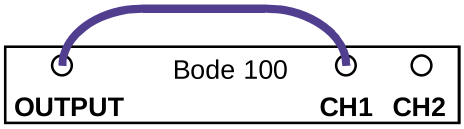
4 Laboratory Part

4.1 Knowing the Instruments

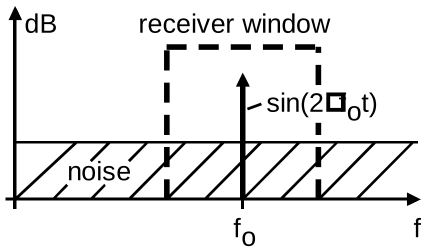
4.1.1 Starting the *BODE100 Vector Network Analyzer* (10min, Σ 0:10h)



(a) Conf. with *OUTPUT* not internally connected to *CH1*



(b) External connection



(c) The receiver band width (RBW) decides about the amount of noise in your measurement.

Fig. 4.1.1: *BODE100* **(a)** setup and **(b)** external setup. **(c)** Getting noise acc. to RBW width.

Switch on power of *BODE100*, connect its USB cable to your PC and start the *Bode Analyzer Suite* (*BAS*) on PC. *BODE100* device No. (lower right corner of *BAS*): CC156C

In the *Configuration* menu set the switch in the middle of Fig. 4.1.1 to the right, so that *CH1* gets its voltage always from "*External reference*". Connect *CH1* externally to *OUTPUT* as shown in Fig. part (b). Part (c) illustrates: More *receiver bandwidth* measures more noise.

Table 4.1.1: Use following ***BODE100*** settings in the ***Frequency Sweep*** mode:

Start Frequency: **10Hz**	Stop Frequency **40 MHz**	Sweep Mode: **Log**arithmic	Number of Points: **201**
Output Level: **0,00 dBm**	Attenuator CH1: **30 dB**	Attenuator CH2: **30 dB**	Receiver Bandwidth: **30 Hz**

Connect *BODE100's OUTPUT* to *CH2*, so that OUTPUT, *CH1* and *CH2* are connected and start a single sweep. ***BODE100* computes U(*CH2*)/U(*CH1*)**. Except from some resonant cable effects at high frequencies you should measure constantly 0dB amplitude and 0° phase.

Frequency-Sweep:

Do you get constantly 0dB and 0° phase from 10Hz ... 1MHz? yes
(If not check your setup and / or ask your supervisor.)

Modify the Attenuator's damping: Strong change in measurement results? no

Set lower Receiver Bandwidths: Does this change the measurement speed? yes
(Measurement duration is approximately 3/RBW per point.)

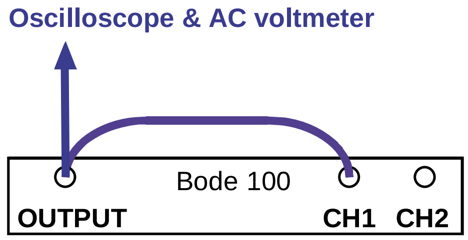
4.1.2 Testing *BODE100*, Oscilloscope and AC Voltmeter (20min, Σ 0:30h)

Fig. 4.1.2:

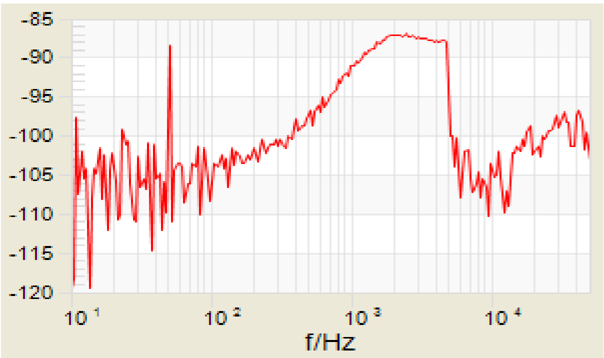
(a) Setup for testing the instrumentation

(b) Measurement with free running *CH2*, receiver bandwidth: 1 Hz, 201 points.

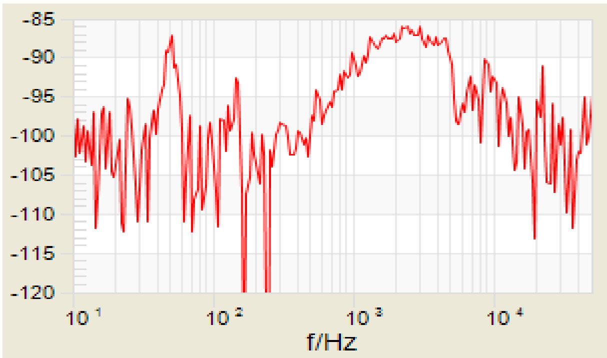
(c) Measurement with free running *CH2*, receiver bandwidth: 10 Hz, 201 points.



(a) *BODE100* test setup



(b) *CH2* disconnected, RBW = 1 Hz



(c) *CH2* disconnected, RBW = 10 Hz

Note: *BODE100* will display the ratio of the voltages U(*CH1*)/U(*CH2*).

BODE100's OUTPUT keeps connected its to *CH1* while *CH2* is disconnected.
Connect *BODE100's OUTPUT* to your oscilloscope and AC voltmeter.

Set *BODE100* to *Gain/Phase* mode,
OUTPUT: *Level*=0dB, *Source Frequency*=1000 Hz, *Attn CH1*: 30dB, *RBW* 1Hz

Oscilloscope: Type: **Tektronix TDS2024B** , Peak-to-peak voltage? **1.3** V_{pp}

AC voltmeter: Type: **VoltCraft VC960** , U_{eff}(f=1KHz)? **445 mV**

Increase *BODE100's OUTPUT Level* to the Overload limit of *CH1* → *MaxLevel* = **17 dB**
(*OUTPUT Level* is limited to ≤13dB, but decreasing *CH1's Attenuation* has same effect as increasing *Level*.)

NoiseLevel at 100Hz should be some –103dB as can be seen from Fig. 4.1.2(b). Compute the dynamic range *MaxLevel-NoiseLevel* at 100Hz (a) in dB, (b) as factor and (c) as **effective number of bits** (**ENOB**). Use authors *NoiseLevel*$|_{100Hz}$ of –103dB (or the slightly different value that you measured yourself).

Range at 100Hz = *MaxLevel–NoiseLevel*$|_{100Hz}$ = **..27dB..- (-103dB) = ..130..dB**

This corresponds to a Factor =**$10^{6.5}$=3162000**, → ENOB = ln(Factor)/ln2 = **21.6** bits

Spurious Free Dynamic Range (**SFDR**) for 0...50 KHz, i.e. *MaxLevel*–max{*NoiseLevel*}, in the range 0...50KHz. Use authors max{*NoiseLevel*}$|_{0...50KHz}$ of –87dB (or the slightly different value that you measured yourself).

SFDR = *MaxLevel*–max{*NoiseLevel*}$|_{0...50KHz}$ = **..27dB..- (-87dB) = ...114... dB**

This corresponds to a Factor = **..501000...**, → ENOB = ln(Factor)/ln2 = **18.9** bits.

Check your AC Voltmeter

To check the valid frequency range of the voltmeter connect its input with *BODE100's OUTPUT* and adjust a measured, effective output voltage of 1V_{eff} (using parameter *Level*). Don't waste time with being too accurate here!

Connect *BODE100's OUTPUT* to the AC voltmeter's input,

control measurement speed with receiver bandwidth, e.g. RBW=10Hz,

Level=7dB →1V_{eff} at 50Hz (or adjust *Level* such that you measure 1V_{eff} with your voltmeter)

Freq. Sweep: 50Hz...1MHz.

The AC voltmeter shows ...

an error of +2% at $f_{+2\%}$ = **....83....** KHz,

a maximum voltage of + **...10...** % at frequency f_{max} = **.....290......** KHz,

an error of -10% at $f_{-10\%}$ = **...454...** KHz

4.1.3 Getting Started with the Board (10min, Σ 0:40h)

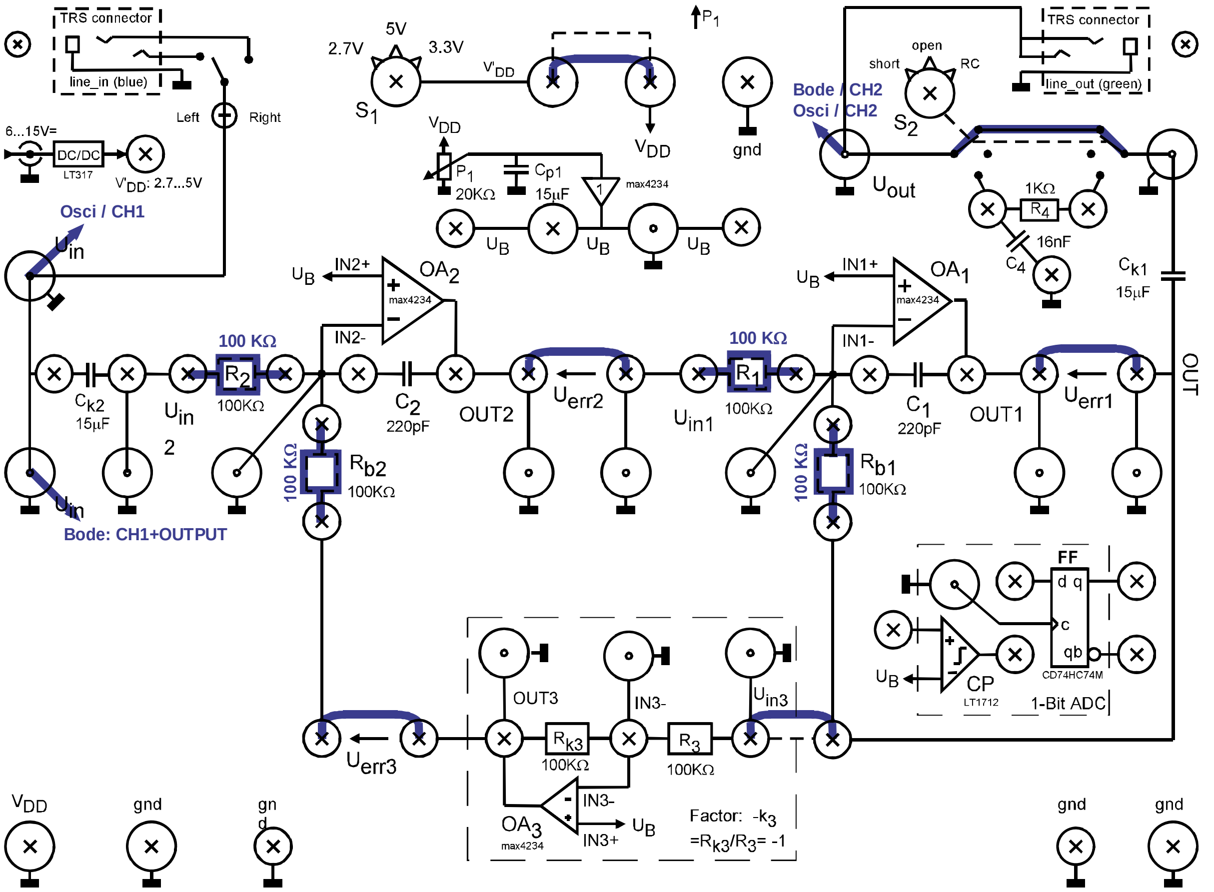


Fig. 4.1.3: Board configuration for first test.

First of all: disconnect clock generator, apply power plug, turn V_{DD} switch to 3.3V.
Then assemble the board as shown in the figure above.

Table 4.1.3: Use following ***BODE100*** settings in the ***Frequency Sweep*** mode:

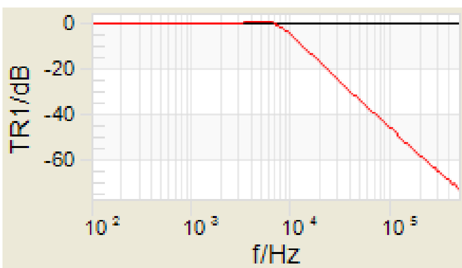
Start Frequency: **10 Hz**	Stop Frequency **500 KHz**	Sweep Mode: **Logarithmic**	Number of Points: **201**
Output Level: **0,00 dBm**	Attenuator CH1: **30 dB**	Attenuator CH2: **30 dB**	Receiver Bandwidth: **10 Hz**

Connect *BODE100's OUTPUT+CH1* to Board's U_{in}, *CH2* to Board's *OUT*. Confirm diagram:

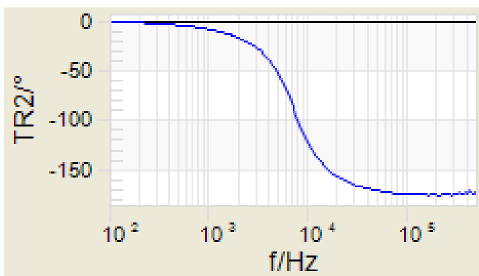
Screen Shot 4.1.3: Bode Diag.

$\frac{U(OUT)}{U(Uin)}$

(a)

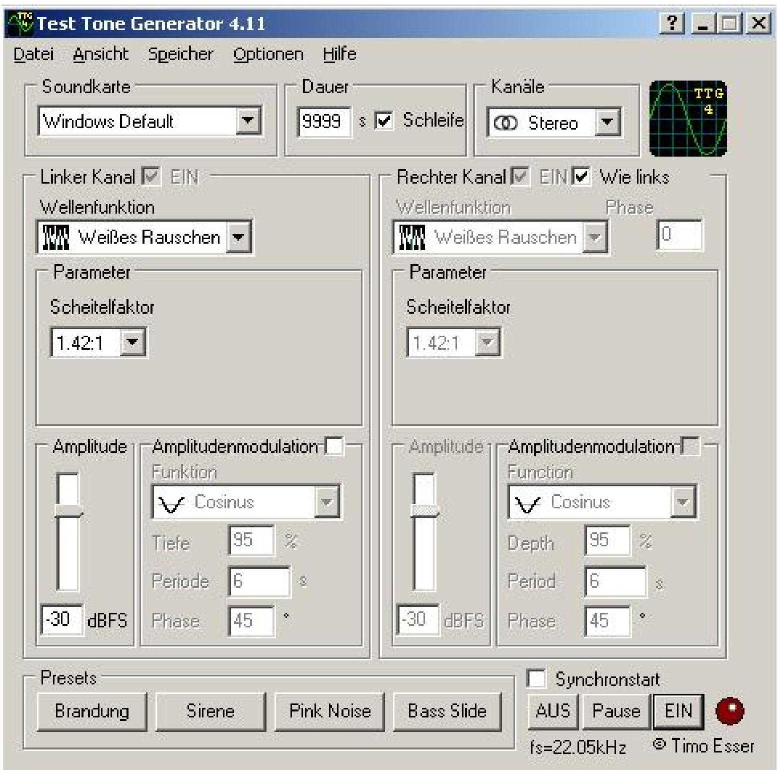


(b)

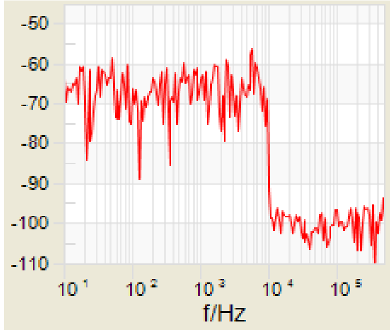


4.1.4 Using the PC's Soundcard as Signal Source (20min, Σ 1:00h)

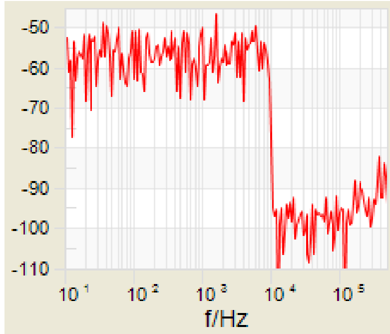
4.1.4.1 White Noise Generation (15min)



(a) Screen window of the *Test Tone Generator 4.11* [10].



(b) *BODE100*: RBW=1Hz



(c) *BODE100*: RBW=10Hz

Fig. 4.1.4.1: Measuring White noise from *Test Tone Generator* 4.11

Connect the PC's soundcard and the external speakers to the board
Disconnect *BODE100's CH1* from the board's U_{in}. (Never connect two low-impedance sources!)
Connect PC-soundcard's "*line out*" (green) with the boards *TRS connector line_in*.
Connect speakers to the board's *TRS connector line_out*.
(*TRS* is an acronym for Tip, Ring, Sleeve, connected to left channel, right channel and ground, respectively)

Start *Test Tone Generator 4.11* [10], [11] as shown in Fig. 4.1.4.1(a) on your PC. Use settings as illustrated above. Set the soundcard of your PC to maximum output amplitude. Use your AC voltmeter to measure the effective output voltage:

Measure u_{eff} of the test-tone generator with settings above: **22.6 mV**

Table 4.1.4.1: Use following ***BODE100*** settings in ***Frequency Sweep*** mode:

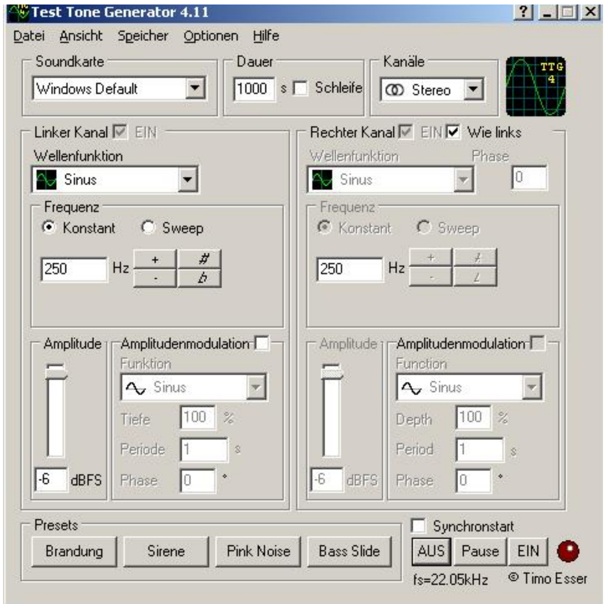
Start Frequency: **10 Hz**	Stop Frequency **500 KHz**	Sweep Mode: **Log**arithmic	Number of Points: **201**
Output Level: **0,00 dBm**	Attenuator CH1: **30 dB**	Attenuator CH2: **30 dB**	Receiver Bandwidth: **10 Hz**

Question: The test-tone generator can generate noise only up to ca. 10KHz. Why? (→ see f_s!)

f_s=22.05KHz → maximum frequency = fs/2 = 11.025 KHz

4.1.4.2 Sin-Wave Generation (read only: 5min)

Fig. 4.1.4.2:
Settings of the *Test Tone Generator 4.11* [10] controlling the PC's soundcard: 220 Hz sinusoidal, 445mV$_{eff}$ (adjusted to 0 dB of *BODE100*)



Start Timo Esser's "*Test Tone Generator 4.11*" on your PC and adjust the settings window according to the figure right. Fig. 4.2.2-2(b) illustrates the spectrum measured for this settings with *BODE100*. The 50Hz peak might come from the 220V/50Hz power supply grid of the laboratory.

BODE100's noise floor for the settings shown in Tab. 4.1.4.2 was checked with the measurement shown in Fig. 4.1.4.3(c) with an assembly according to figure part (a).

Then the sin-wave generator illustrated in Fig. 4.1.4.2 drove the PC's soundcard. The assembly according to Fig. 4.1.4.3(b) yielded the measurement shown in Fig. part (d).

Fig. 4.1.4.3:

(a) Config. for (c)

(b) Config. for (d)

(c) *Bode100* measurement with config. (a).

(d) *Bode100* measurement with config. (b).

Each measurement acc. to Tab. 4.1.4.2 took some 70min.

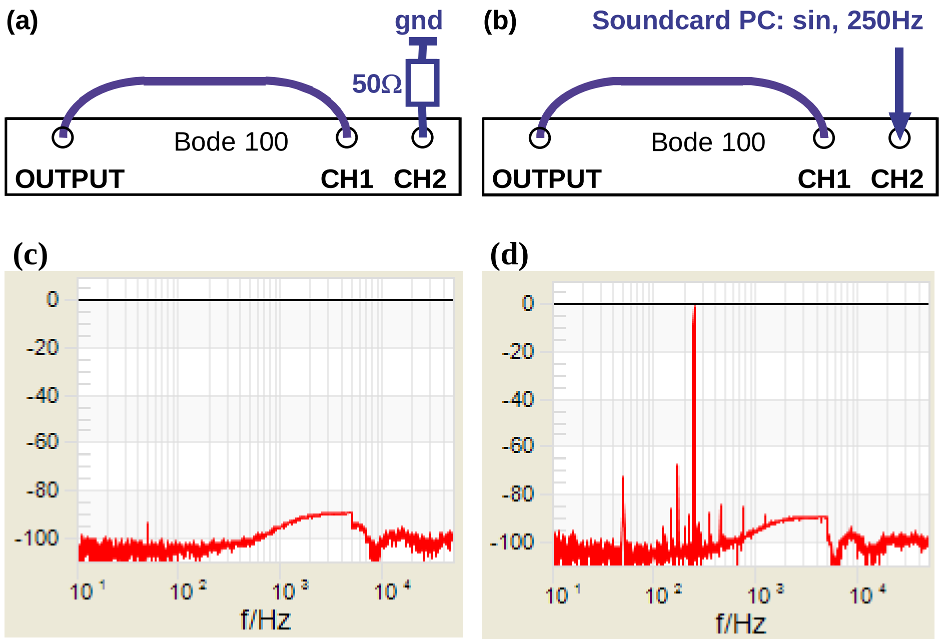


Table 4.1.4.2: Use following ***BODE100*** settings in ***Frequency Sweep*** mode:

Start Frequency: **10 Hz**	Stop Frequency **50 KHz**	Sweep Mode: **Logarithmic**	Number of Points: **1601**
Output Level: **0,00 dBm**	Attenuator CH1: **30 dB**	Attenuator CH2: **30 dB**	Receiver Bandwidth: **1 Hz**

4.1.5 Comments on Using *Bode100* as Spectrum Analyzer (10min, Σ 1:10h)

If you have spent more than 1 hour for chapter 4.1 of this laboratory, then you'd better continue with chapter 4.2 and read the following statements at home.

The *BODE100* is a network analyzer, not a spectrum analyzer. In the text above and below we use it in both modes: for network and spectrum analysis. The latter mode was chosen when *BODE100's OUTPUT* is not used as input signal to the circuit under test. As *BODE100* is not intended for spectrum analysis some drawbacks have to be accepted in this mode:

1. For frequencies above 6KHz we have a superposition of the spectrum to be measured and a slightly shifted image of it. The shift varies from some 100Hz to some KHz depending on frequency and receiver bandwidth. This could result in indicated +3dB within the diagrams of this chapter. Fortunately, this effect is not very obvious in this lab.
2. Measurement results of spectral power densities like white noise depend on the receiver bandwidth (RBW), as will be shown and explained in more detail below.

As illustrated in Fig. 4.1.1(c) a sinusoidal signal corresponds to a Dirac function on the frequency axis so that the width of the receiver window has few impact on the measured signal power. However, white noise – being constant over the frequency axis – delivers 10 times more power within a 10 times wider bandwidth. A factor 10 in power corresponds to 10dB in the Bode diagram as visible for example in Figs. 4.1.4.1 (b) and (c).

Signals with different frequencies are always uncorrelated. Uncorrelated signals sum in power. Therefore, white-noise power grows linear with bandwidth and consequently the effective noise voltage grows with the square-root of the bandwidth. It is measured in V/$\sqrt{Hz}$. (Frequently **`rtHz`** is used as acronym for $\sqrt{Hz}$.)

In chapter 4.1.4.1 we measured $u_{n,eff}$ = 22.6 mV effective noise voltage generated by the sound card. It is equidistributed over a frequency range of 10 KHz. Consequently, the effective noise voltage caused by 1 Hz bandwidth is 22.6 mV/$\sqrt{10000 Hz}$ = 22.6 µV/$\sqrt{Hz}$.

For *Bode100* we measured an effective voltage of 445mV as 0dB. Consequently, the –60dB effective noise voltage measured in Fig. 4.1.4.1(b) with RBW=1Hz correspond to a factor 1/1000 resulting in 445µV/$\sqrt{Hz}$. This is about twice the value measured with the AC voltmeter and illustrates, that *BODE100* is not intended to perform accurate spectrum analysis. Nevertheless, it is useful to deliver valuable insight into noise shaping, the subject of this laboratory.

4.2 ΔΣ Modulation

4.2.1 Audiometric Test of 0^{th} , 1^{st} , 2^{nd} Order Modulators. (20min, Σ 1:30h)

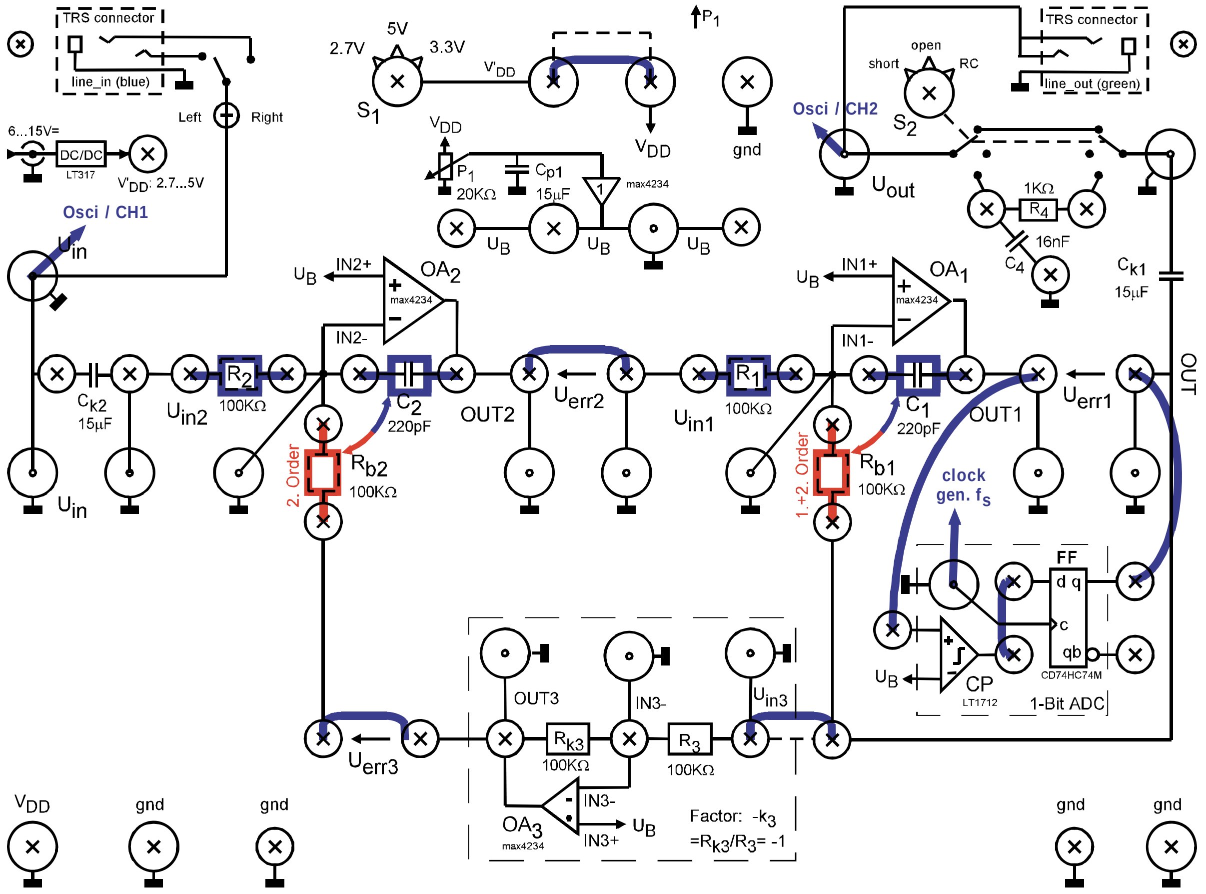


Fig. 4.2.1: Board configuration for modulators of 0^{th}, 1^{st} and 2^{nd} order.

Goal of this subsection is to get an impression of the power of ΔΣ modulation by simply **hearing** the differences in sound. In next subsections we will measure the differences.

Do not destroy the flipflop's clock input! Voltages < **0V** and >**V_{DD}** can destroy the flipflop. **Before** connecting the clock generator to the FF's clock input make sure that

(a) $V_{DD} \geq 2.7V$ **AND** **(b)** the clock signal is within the range 0.1...2.7V

For clocking with Altera DE2 board and monitoring with DA2 board's DAC see chapter 5.

Assemble the circuit drawn in Fig. 4.2.1 with clock swing 0...V_{CC}.

First test:

Connect oscilloscope's *CH1* and *CH2* to the board's U_{in} and U_{out}, respectively.

R_{b1}, R_{b2} are in vertical positions (red). The board's *FF* should toggle now!

Connect PC's soundcard and speakers to the board

Disconnect *BODE100's CH1* from the board's U_{in}. (Never connect two low-impedance sources!)

Connect your PC-soundcard's "*line out*" (green) with the boards *TRS connector line_in*.

Connect your speakers to the board's *TRS connector line_out*.

Get sound from the internet (e.g. www.youtube.com or www.antenne.de)
Set switch S_2 to "short". (It is located before the output to the speakers.)

0th order modulator
Set resistor R_{b1} horizontal (blue) and R_{b2} horizontal (blue) and rate the sound:

	sound not identified									sound excellent
f_s	**0**	**1**	**2**	**3**	**4**	**5**	**6**	**7**	**8**	**9**	**10**
10 KHz	O	X	O	O	O	O	O	O	O	O	O
100 KHz	O	O	X	O	O	O	O	O	O	O	O
1 MHz	O	O	X	O	O	O	O	O	O	O	O
10 MHz	O	O	X	O	O	O	O	O	O	O	O

1st order modulator
Set resistor R_{b1} horizontal (blue) and R_{b2} vertical (red) and rate the sound:

	sound not identified									sound excellent
f_s	**0**	**1**	**2**	**3**	**4**	**5**	**6**	**7**	**8**	**9**	**10**
10 KHz	X	O	O	O	O	O	O	O	O	O	O
100 KHz	O	O	O	O	O	X	O	O	O	O	O
1 MHz	O	O	O	O	O	O	O	X	O	O	O
10 MHz	O	O	O	O	O	O	O	O	X	O	O

2nd order modulator
Set resistor R_{b1} vertical (red) and R_{b2} vertical (red) and rate the sound:

	sound not identified									sound excellent
f_s	**0**	**1**	**2**	**3**	**4**	**5**	**6**	**7**	**8**	**9**	**10**
10 KHz	X	O	O	O	O	O	O	O	O	O	O
100 KHz	O	O	O	O	O	O	X	O	O	O	O
1 MHz	O	O	O	O	O	O	O	O	X	O	O
10 MHz	O	O	O	O	O	O	O	O	O	X	O

Draw conclusions:
For the stand-alone quantizer ...
increasing *OSR* has O no **X** moderate O strong effect on sound quality

For a quantizer properly operated by a ΔΣ modulator ...
increasing *OSR* has O no O moderate **X** strong effect on sound quality
increasing modulator order has O no O moderate **X** strong effect on sound quality

Further options:
- You can switch S_2 to include an RC lowpass into the signal path to the speakers.
- Or set S_2 to "open" and bridge it with a 4th order lowpass, f_g=5KHz. (Don't short-circuit lowpass!)
- You may use a digital lowpass as ΔΣ demodulator, available e.g. at HS.R [13].

4.2.2 Investigating Unshaped White Noise (0th Order) (20min, Σ 2:00h)

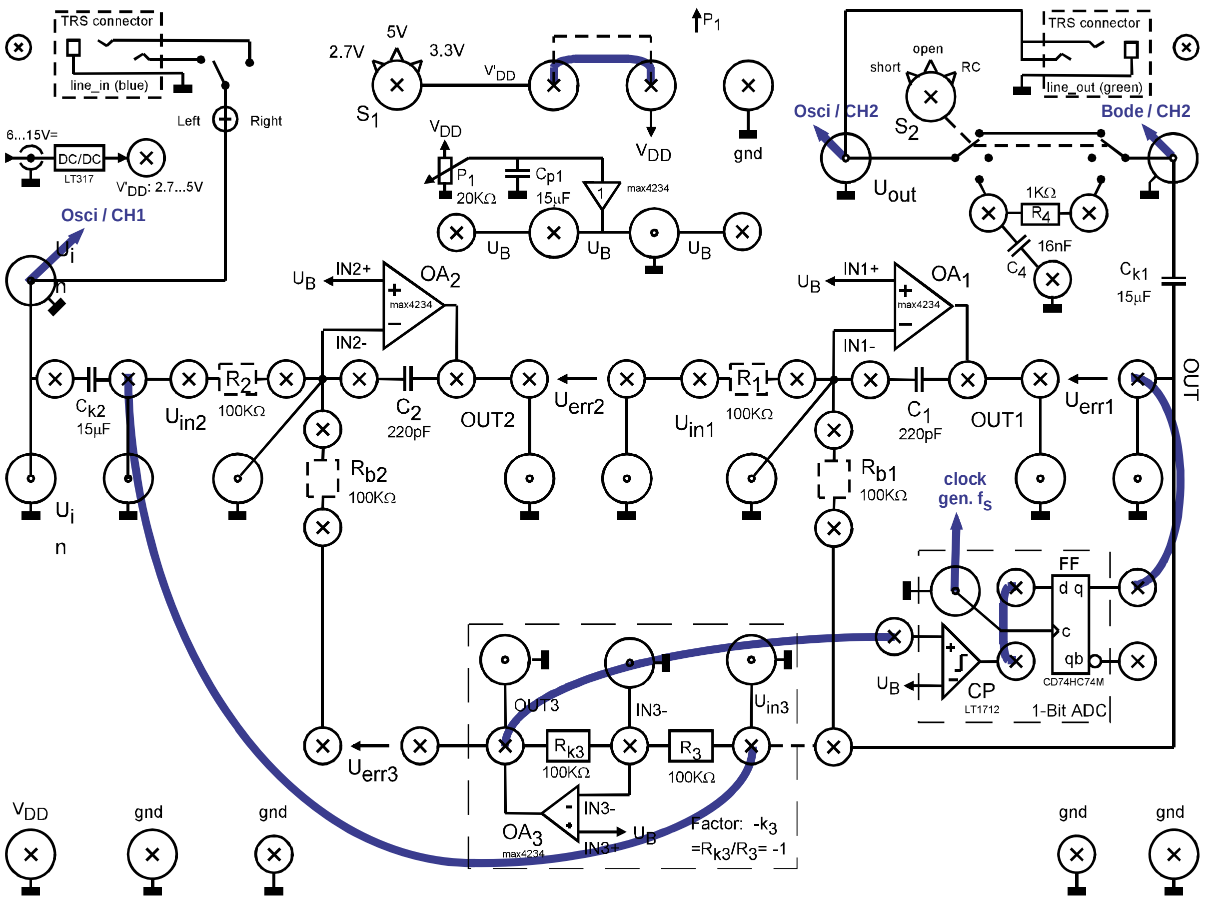


Fig. 4.2.2-1: Board configuration for unshaped noise (0th order).

Goal: Demonstrate that spectral power density of a random-bit stream is flat over frequency and proportional to Δ^2/f_s in the frequency range 0...f_s/2., while eff. voltage density is ≈ $\Delta/\sqrt{f_s}$.

Quantization noise according to (2.10) is not easy to measure. We need to obtain the difference between unquantized and quantized signal, and often the first is analog and the latter digital. The way we go here is to generate a random bit-stream from a white-noise input signal $u_{in}(t)$, with $|u_{in}(t)| << \Delta/2$.

Choose *Test Tone Generator* settings according to Fig. 4.1.4.1(a) and feed the noise signal from your PC's sound card into the board's *TRS connector line_in*. Assemble the board as illustrated in Fig. 4.2.2-1 above.

Table 4.2.2: Used ***BODE100*** settings in the ***Frequency Sweep*** mode:

Start Frequency: **10 Hz**	Stop Frequency: $f_s/2$	Sweep Mode: **Log**arithmic	Number of Points: **201**
Output Level: **0,00 dBm**	Attenuator CH1: **30 dB**	Attenuator CH2: **30 dB**	Receiver Bandwidth: **1 Hz**

As *Bode100*'s measurement time per point is ca. $T_{meas} \approx 3/RBW$ (4.2),
the 201-point measurements shown in Fig. 4.2.2-2 take about 10 minutes each.

Settings: V_{DD}=2.7, 3.3, 5V, Sampling clock: f_s=100Hz, 1KHz, 100KHz with swing 0...3.3V.

Run the measurement for missing Fig. 4.2.2-2(c) while answering the questions below. Sketch the result in (C). (101 points are enough. Begin with a fast test using RBW=10Hz.)

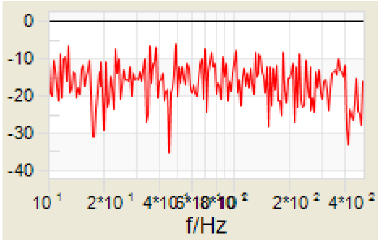
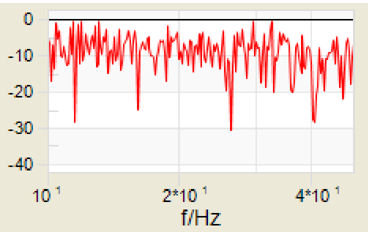
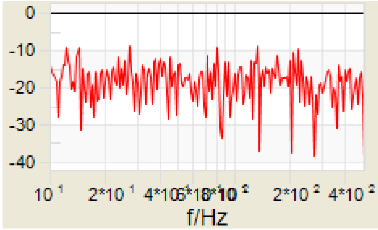
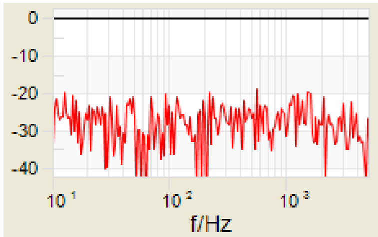
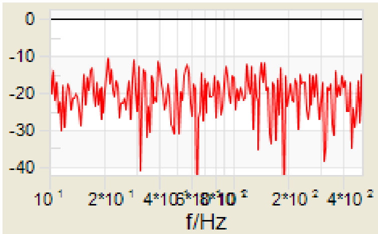
	 **(a)** f_s=1KHz, Δ=5V	
 **(b)** f_s=100Hz, Δ=3.3V	 **(c)** f_s=1KHz, Δ=3.3V	 **(d)** f_s=10KHz, Δ=3.3V
	 **(e)** f_s=1KHz, Δ=2.7V	

Fig. 4.2.2-2: Noise amplitude densities measured in the frequency band 10...50Hz

Fill the following statements on the $1/\sqrt{f_s}$ -relationship of spectral noise voltage density:

From Fig. 4.2.2-2(b) to (d) the frequency range 0...f_s/2 is extended by a factor 100.

Consequently, the noise <u>power</u> density, which is ≈ V^2/Hz, decreases by a factor ...100...

and the noise <u>voltage</u> density measured in $V/\sqrt{Hz}$ decreases by a factor .. $\sqrt{100}$ = 10.

Both corresponds to change on the logarithmic scale of -20 dB.

From Fig. 4.2.2-2(e) to (a) Δ=V_{DD} increases by factor 5 / 2.7 = 1.85 = 5.35 dB

Consequently, the noise voltage density increases by factor 5/2.7=1.85 = 5.35 dB

and the noise power density per *Hz* decreases by a factor ..$1,85^2$=3.43. = 5.35 dB

4.2.3 Investigating the 1st Order Delta-Sigma Modulator (20min, Σ 2:20h)

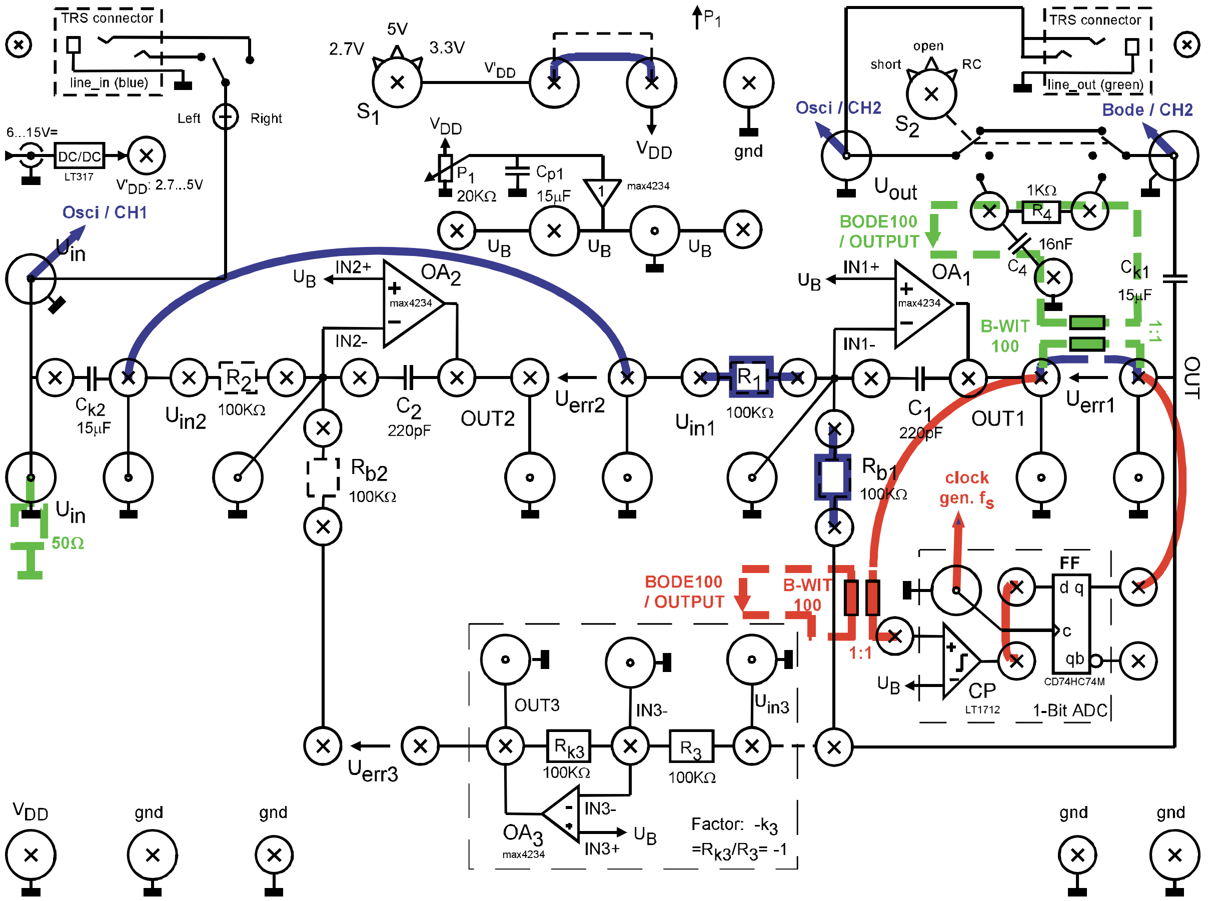


Fig. 4.2.3: Board configuration for 1st order modulator.

Goal: Run a 1st order ΔΣ modulator, measure STF, NTF and noise shaping.

Table 4.2.3: Used following ***BODE100*** settings in ***Frequency Sweep*** mode:

Start Frequency: **10 Hz**	Stop Frequency **50 KHz**	Sweep Mode: **Logarithmic**	Number of Points: **201**
Output Level: **0,00 dBm**	Attenuator CH1: **30 dB**	Attenuator CH2: **30 dB**	Receiver Bandwidth: **10 Hz**

Time-continuous transfer function (TF) measurements (V_{DD}=3.3V):

STF, time-continuous: Assemble the circuit as illustrated in Fig. 4.2.3 with blue solid and dashed lines, i.e. U_{err1} is a short circuit, comparator and flipflop not connected. Measure the signal transfer function (discussed detailed in [12]) with *BODE100* to confirm Fig. 4.2.5(a).

NTF, time-continuous: As indicated with dashed green lines in the figure above connect the Board's U_{in} to ground using a short circuit or 50Ω (or any resistor small compared to R_2=100KΩ). Remove the short-circuit over U_{err1} and generate U_{err1} by *BODE100's OUTPUT*, which is made independent from ground by the *B-WIT-100* isolating transformer. Keep *BODE100's CH1* connected to its *OUTPUT* and *CH2* to the board's U_{out}. Measure the noise transfer function (NTF) with *BODE100*. The result should comply with Fig. 4.2.5(c) [12].

Time-Discrete TF measurements with 2-level quantizer:

Settings: V_{DD}=3.3V, Sampling clock: f_s=100KHz with swing 0...3.3V.
Remove *B-WIT100* transformer.
Connect the flipflop's clock input to the clock generator (red line).
Connect the board's *OUT1* to *CP's* IN^+ (red line without *B-WIT 100*).
Connect and *CP's* output to the *FF's D*-input (red line)
Connect and *FF's* output to the board's *OUT* (red line).

STF, time-discrete:
Remove the 50Ω short circuit from U_{in} and connect U_{in} with *BODE100*'s *OUTPUT*.
Connect the board's *Uout* with *BODE100*'s *CH2*. (*CH1* remains connected to *OUTPUT*.)
Measure the *STF* with *BODE100* and sketch it in Fig. 4.2.5(b).

NTF, time-discrete:
Connect the board's U_{in} with low impedance (≤50Ω) to ground.
Connect *B-WIT 100* transformer on the board between *OUT1* and the *CP*'s IN^+ (red dashed).
Measure the NTF with *BODE100* and sketch the result in Fig. 4.2.5(d).

Shaped noise measurements (V_{DD}=3.3V)**:**

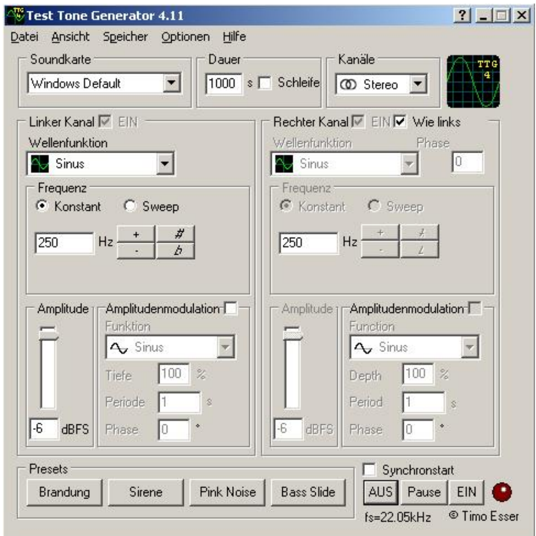
Replace *B-WIT100* (red dashed) by a short.

Disconnect *BODE100's OUTPUT* from U_{in}. Connect U_{in} to your PC's soundcard.

Generate a 250Hz signal at *BODE100's* 0dB-level with your soundcard (settings see right).

Perform a frequency sweep and confirm the measurement observed in Fig. 4.2.5(e). Does the measurement comply with your expectations of Fig. 3.4-2?

yes
. . . .



Test tome generator: 250Hz sinus

Complete the following sentences:

We observe (1) the **250Hz signal** as Dirac function, (2) the noise floor at ca. **-90** dB and (3) the **sin(πF)** behavior for $f > f_{NCF1}$ (noise corner frequency).

The f_{NCF1} is observed at about **20** Hz.

The maximum slope of the noise shape for this 1st order modulator is **20 db/dec**

4.2.4 Investigating the 2nd Order Delta-Sigma Modulator (20min, Σ 2:40h)

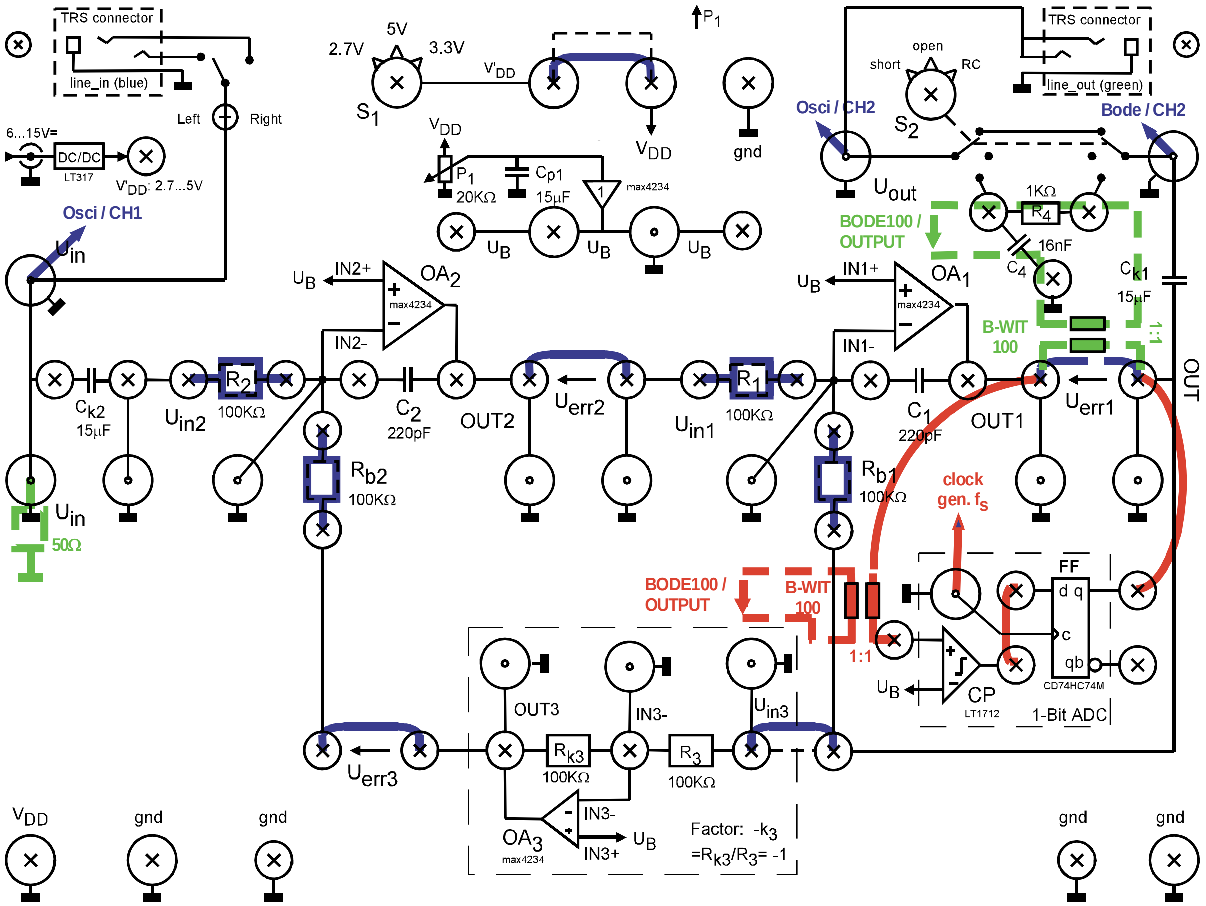


Fig. 4.2.4: Board configuration for 2nd order modulator.

Goal: Run a 2nd order ΔΣ modulator, measure STF, NTF and noise shaping.

Table 4.2.4: Used following ***BODE100*** settings in ***Frequency Sweep*** mode:

Start Frequency: **10 Hz**	Stop Frequency **50 KHz**	Sweep Mode: **Logarithmic**	Number of Points: **201**
Output Level: **0,00 dBm**	Attenuator CH1: **30 dB**	Attenuator CH2: **30 dB**	Receiver Bandwidth: **10 Hz**

Time-continuous transfer function (TF) measurements (V_{DD}=3.3V):

STF, time-continuous: Assemble the circuit as illustrated in Fig. 4.2.4 with blue solid and dashed lines, i.e. U_{err1} is a short circuit, comparator and flipflop not connected. Measure the signal transfer function (discussed detailed in [12]) with *BODE100* to confirm Fig. 4.2.5(a).

NTF, time-continuous: As indicated with dashed green lines in the figure above connect the Board's U_{in} to ground using a short circuit or 50Ω (or any resistor small compared to R_2=100KΩ). Remove the short-circuit over U_{err1} and generate U_{err1} by *BODE100's OUTPUT*, which is made independent from ground by the *B-WIT-100* isolating transformer. Keep *BODE100's CH1* connected to its *OUTPUT* and *CH2* to the board's U_{out}. Measure the noise transfer function (NTF) with *BODE100*. The result should comply with Fig. 4.2.5(c) [12].

Time-Discrete TF measurements with 2-level quantizer:

Settings: V_{DD}=3.3V, Sampling clock: f_s=100KHz with swing 0...3.3V.
Remove *B-WIT100* transformer.
Connect the flipflop's clock input to the clock generator (red line).
Connect the board's *OUT1* to *CP's IN*$^+$ (red line without *B-WIT 100*).
Connect and *CP's* output to the *FF's D*-input (red line)
Connect and *FF's* output to the board's *OUT* (red line).

STF, time-discrete:
Remove the 50Ω short circuit from U_{in} and connect U_{in} with *BODE100*'s *OUTPUT*.
Connect the board's *Uout* with *BODE100*'s *CH2*. (*CH1* remains connected to *OUTPUT*.)
Measure the *STF* with *BODE100* and sketch it in Fig. 4.2.5(b).

NTF, time-discrete:
Connect the board's U_{in} with low impedance (≤50Ω) to ground.
Connect *B-WIT 100* transformer on the board between *OUT1* and the *CP*'s *IN*$^+$ (red dashed).
Measure the NTF with *BODE100* and sketch the result in Fig. 4.2.5(d).

Shaped noise measurements (V_{DD}=3.3V):

Replace *B-WIT100* (red dashed) by a short.

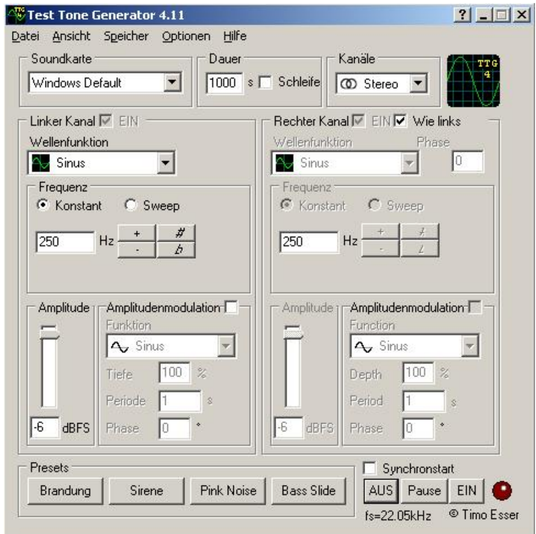
Disconnect *BODE100's OUTPUT* from U_{in}. Connect U_{in} to your PC's soundcard.

Generate a 250Hz signal at *BODE100's* 0dB-level with your soundcard (settings see right).

Perform a frequency sweep and confirm the measurement observed in Fig. 4.2.5(e). Does the measurement comply with your expectations of Fig. 3.5-2?

yes

....



Test tome generator: 250Hz sinus

Complete the following sentences:

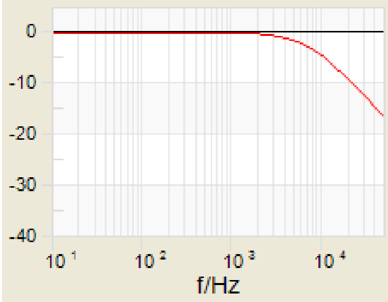
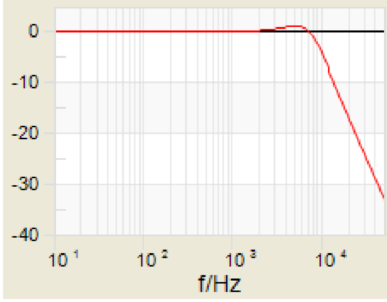
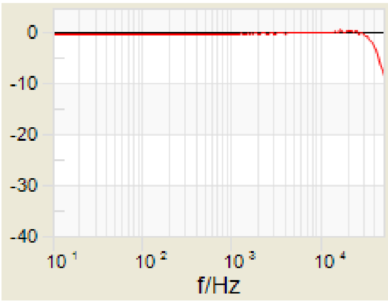
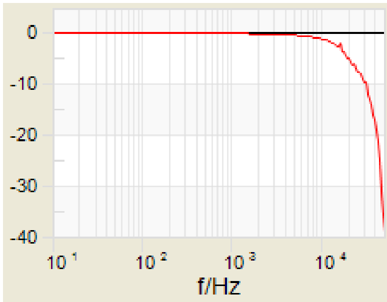
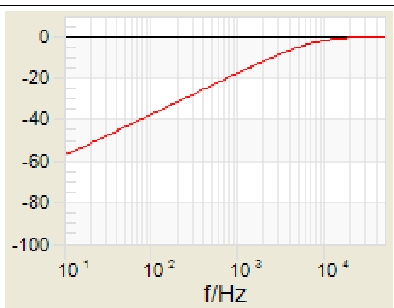
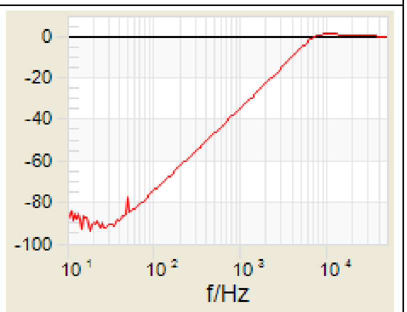
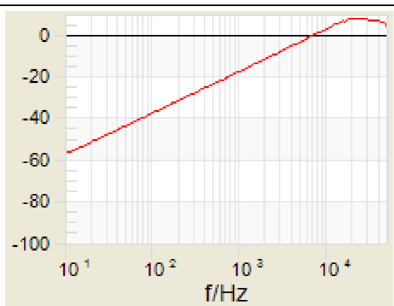
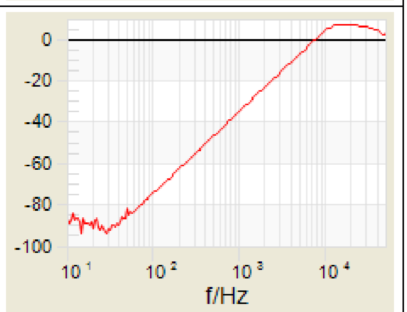
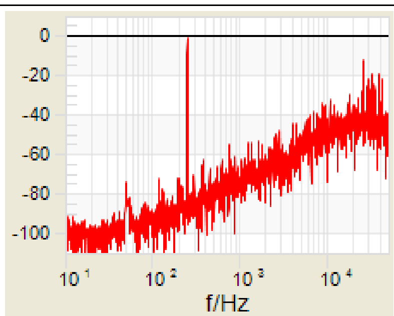
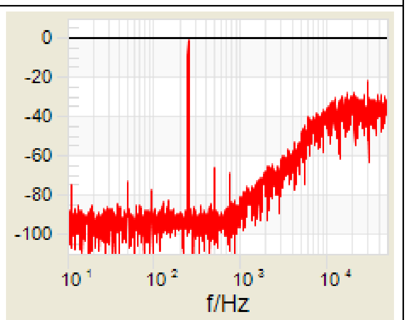
We observe (1) the **250Hz signal** as Dirac function, (2) the noise floor at

ca. **-90** dB and (3) the **$\sin^2(\pi F)$** behavior for $f > f_{NCF2}$ (noise corner frequency).

The f_{NCF2} is observed at about **1000** Hz.

The maximum slope of the noise shape for this 2nd order modulator is **40 db/dec**

4.2.5 Characteristics Summary of 1st and 2nd Order ΔΣ Modulators

Fig. 4.2.6. V_{DD}=3.3V	**1st Order Modulator**	**2nd Order Modulator**
(a) STF no quantizer (time and value continuous), RBW = 1 Hz, 201 points.		
(b) STF 2-level quantizer, f_s=100KHz (time and value discrete), RBW = 1 Hz, 201 points.		
(c) NTF no quantizer (time and value continuous), RBW = 1 Hz, 201 points.		
(d) NTF 2-level quantizer, f_s=100KHz (time and value discrete), RBW = 1 Hz, 201 points.		
(e) 250 Hz input signal, f_s=100KHz, 2-level quantizer, (time and value discrete) RBW = 1 Hz, 1601 points.		

4.3 Check your Knowledge

What is *OSR*? How does quantization-noise power in the baseband depend on the *OSR*? How does the effective quantization-noise voltage in the baseband depend on the *OSR*? Give both as functions of modulator order *L*. What 3 sections can you identify in the spectral noise voltage densities of Fig. 4.2.6(e)? Can you identify the noise corner frequencies? What function describes the rising section of the power-spectra for L=1 and L=2? What is the maximum slope of the rising spectra for L=1 and L=2 in dB/dec?

5 Conclusions

After presenting some fundamentals of signal processing in section 2, theory of ΔΣ modulation was presented in section 3 and verified in section 4. Unshaped quantization noise behaves according to $\Delta/\sqrt{f_s}$ with Δ being the smallest step of the quantizer and f_s the sampling frequency. ΔΣ modulated quantization noise behaves according to $\Delta/OSR^{L+½}$, where L is the order of the modulator and $OSR=f_s/2f_B$ the oversampling ratio over baseband f_B. This formula includes for L=0 the case of unshaped noise.

6 References

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[13] M. Schubert, [4], Document → Digital Filter for ΔΣ Demodulation (Lab 2 at HS.R).