The CERN Accelerator School
"*Magnets*"
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Electronics for Measurement Systems

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Outline

- Design hints
  - Integration effects (10 min)
  - Oversampling effects (10 min)
  - Noise effects (20 min)
  - Architecture evolution (20 min)

- Characterization hints
  - Metrological specs (30 min)
  - Metrological tests (30 min)

Outline

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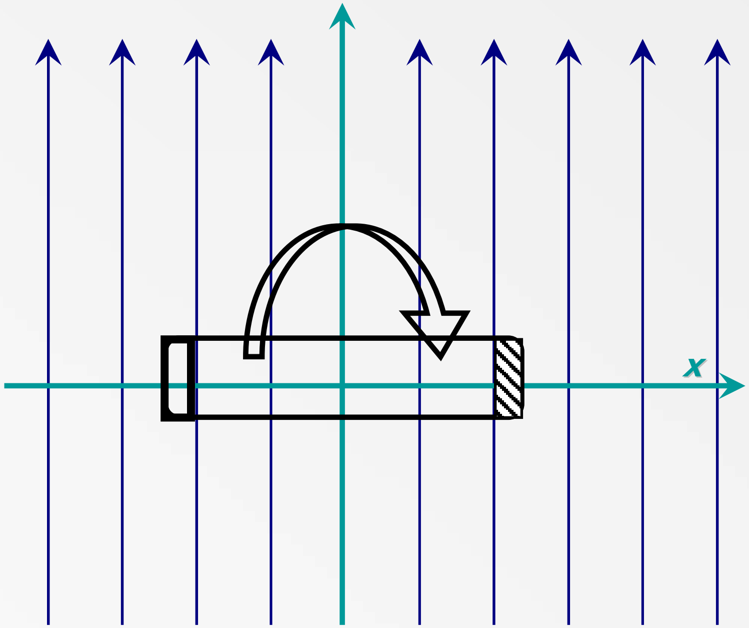
Design Hints:
Integration Effects

- Integration background
- Integration effect on the noise
    - normal-mode noise
    - white noise

Integration Background

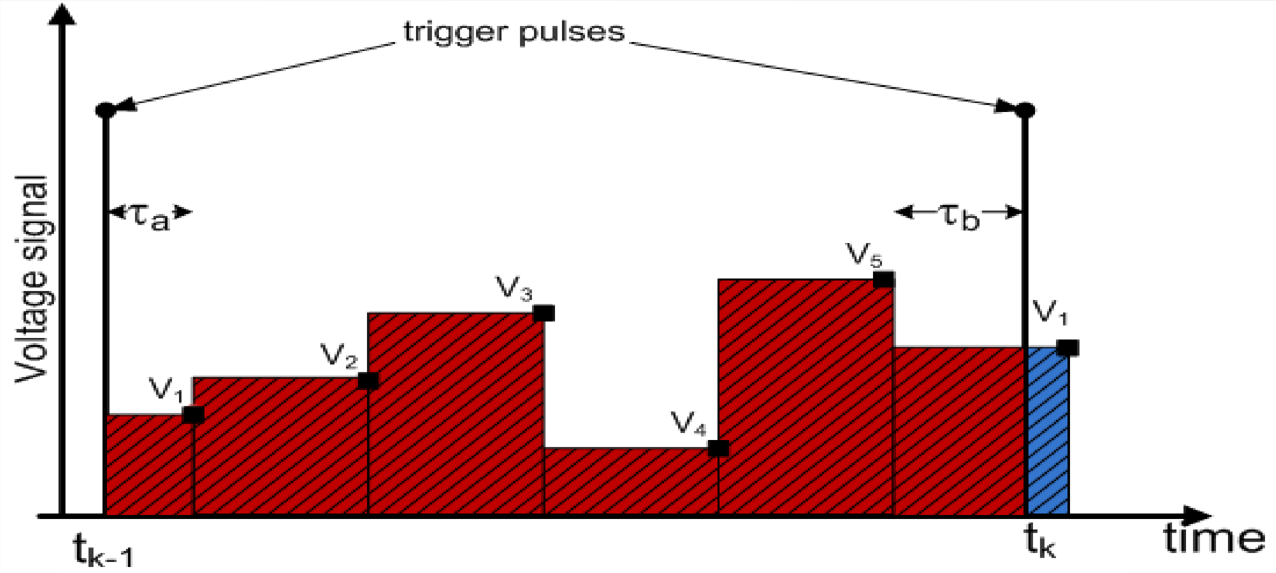
- Measurement by direct application of the Faraday law.
- Used for mapping the magnetic field intensity.
- By moving from a null field region, the voltage integral is equal to the magnetic flux variation, ΔΦ.
- By dividing the flux by the coil area the magnetic flux density By is obtained.
- <u>Typical accuracy 0.01 %.</u>

$$\Phi(\pi) - \Phi(0) = 2 \cdot \int_0^L B_y(x) \cdot dl$$



Integration Background

- $\Delta\Phi$ is computed between two trigger pulses t_{k-1} and t_k.
- A correspondence between space and time domains is created.
- The flux is sampled asynchronously at the trigger frequency.



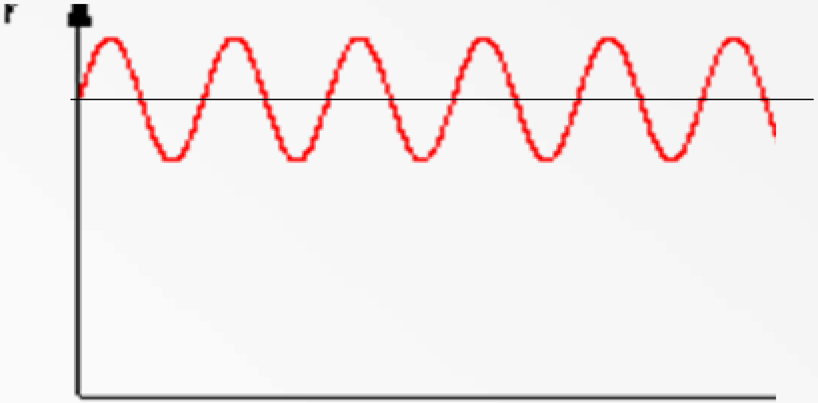
$$\Delta\varphi_k = V_{k_1}\cdot\tau_{a_k} + \sum_{i=2}^{N} V_{k_i}\cdot\tau_i + V_{(k+1)_1}\cdot\tau_{b_k}$$

Integration effects on the noise

Normal mode noise

A *alternative* noise superimposed to the measurand, i.e. with

- null average over a period,
- odd function in the half period (symmetrical with respect to the origin of a reference placed on the half period).



Integration effects on the noise

Normal-Mode Rejection (NMR)

Capability of an electronic circuit of attenuating the normal-mode noise.

Integration effects on the normal-mode noise

The integration over an integer number of periods suppresses the normal-mode noise by definition, i.e. intrinsically.

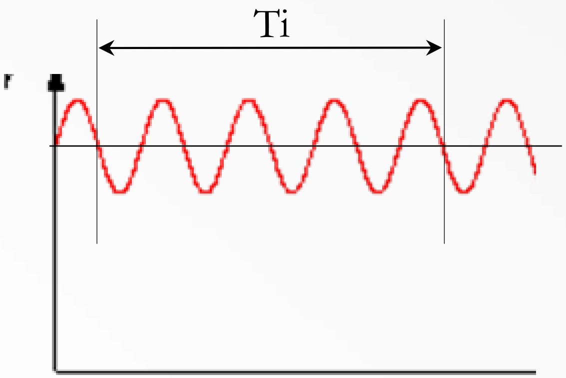
Thus, the integration *rejects* the normal-mode noise intrinsically.

Integration effects on the normal-mode noise

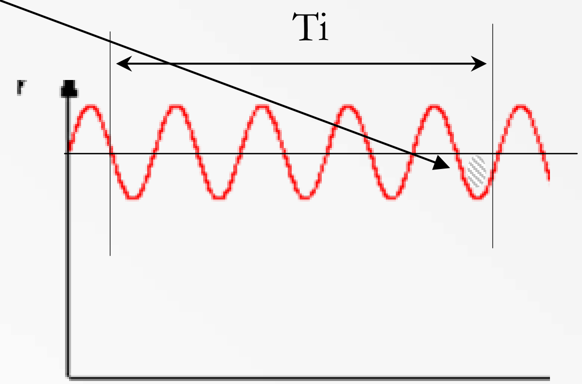
The NMR of the integration depends on:

- integer number of noise periods inside the integration time Ti,
- residual amount of nonintegrated normal-mode noise.

In synthesis, the NMR of the integration in a given Ti is related to the <u>relative residual</u>:



Maximum rejection



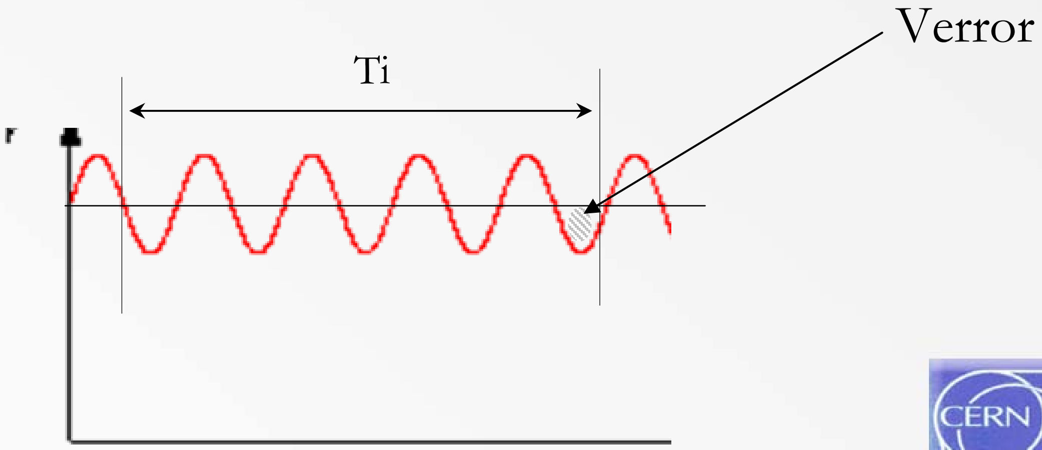
Minimum rejection

Integration effects on the normal-mode noise

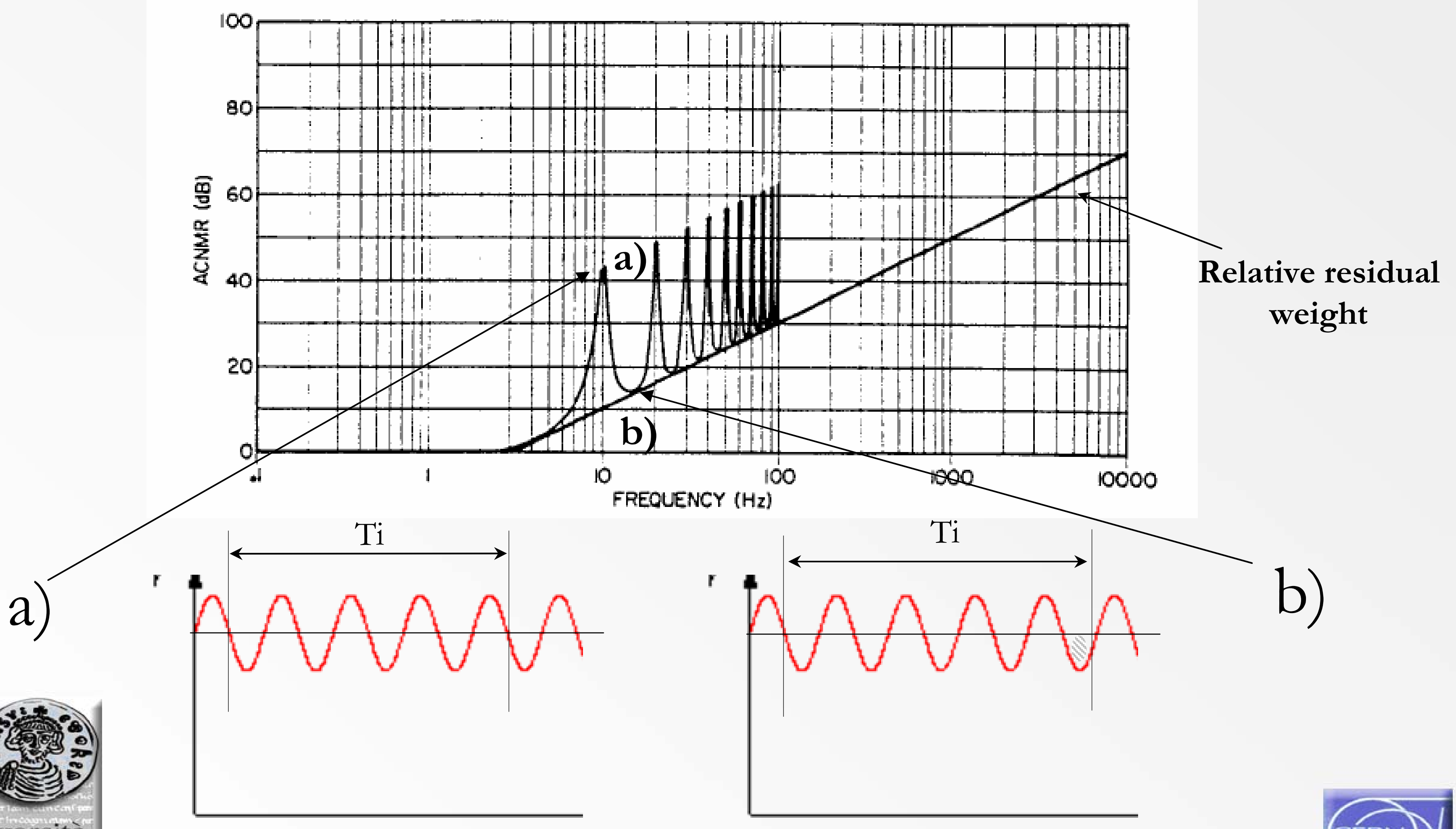
Normal Mode Rejection Ratio (NMRR) :

$NMRR = 20*\log(V_{in}/V_{error})$

where V_{error} is the value returned by the integration for an applied normal-mode voltage V_{in}.



NMRR vs noise frequency (for a given Ti)



Relative residual weight

Maximum rejection

Minimum rejection

The white noise

- Random signal with a flat power spectral density
- Name from *white light*: the power spectral density of the light is distributed over the visible band so that the eye's three color receptors (cones) are approximately equally stimulated

The white noise

- An infinite-bandwidth, white noise signal is only theoretical
- By having power at all frequencies, the total power of such a signal is infinite and therefore impossible to generate.
- In practice, however, a signal can be "white" with a flat spectrum over a defined frequency band.

Integration effects on the white noise

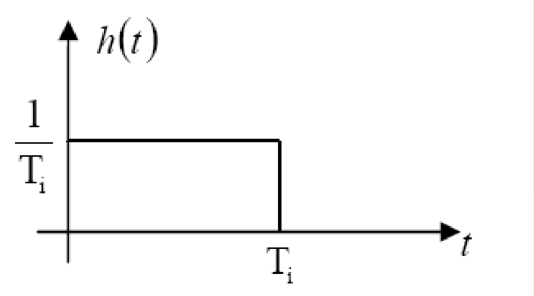
The white noise is low-pass filtered by the integration.

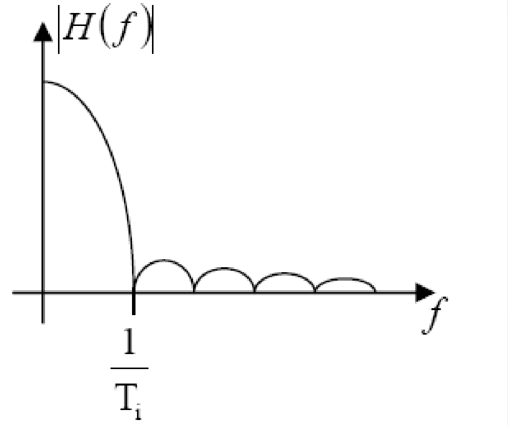
$$v_y(t) = \frac{1}{T_i} \int_{t-T_i}^{t} v_x(\tau)\, d\tau$$

Convolution integral

$$H(f) = e^{-j2\pi f \frac{T_i}{2}} \operatorname{sinc}(fT_i)$$

Frequency response





Integration effects on the white noise

- The filter is linear, thus the superimposed noise is not modified in frequency but attenuated in amplitude by the module of *H(f)*.

- At low frequency. the attenuation is not significant, thus the effect is significant for

$$f \geq {}^{1}\!/_{\mathrm{T_i}} \; .$$

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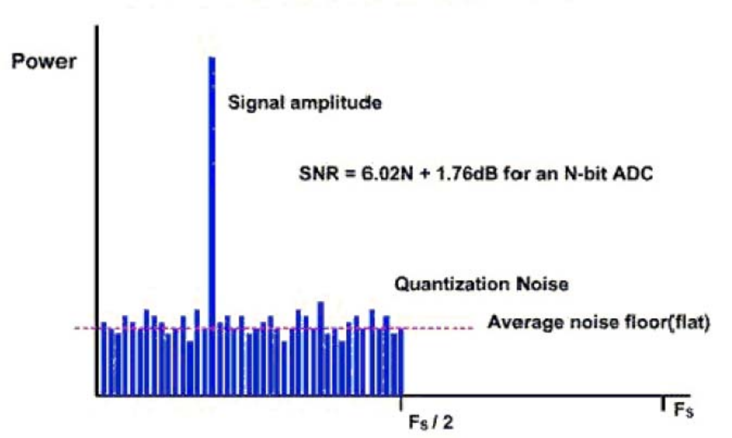
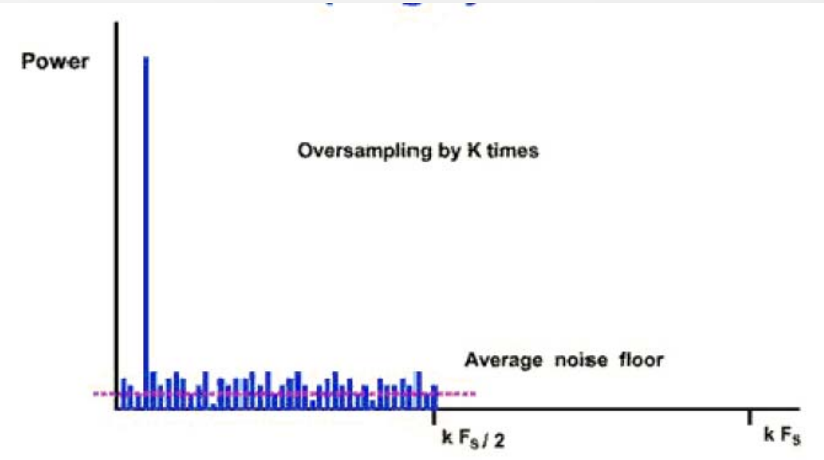
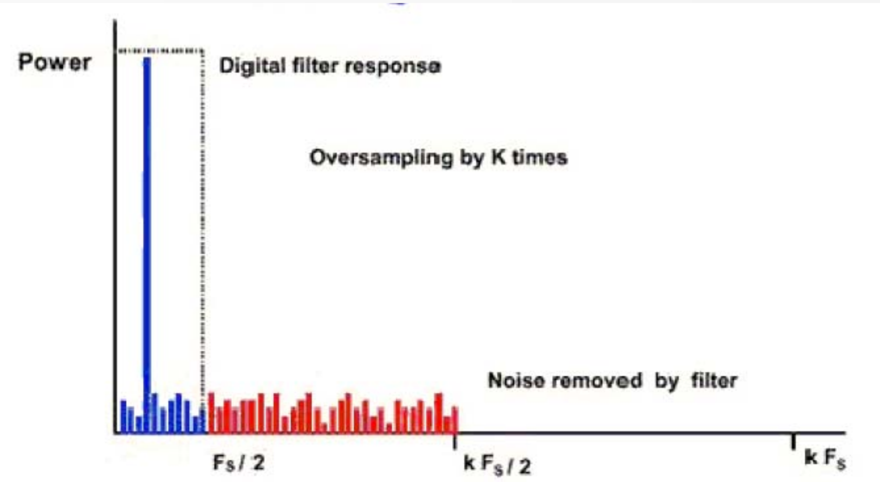
Design Hints: Oversampling Effects

- Oversampling background
- Oversampling effects:
    - in cheaper higher-resolution integrators
    - in anti-aliasing
    - in noise reduction

Oversampling Background







Oversampling Background

Oversampling: sampling with a rate significantly higher than twice the bandwidth or highest frequency of the signal

A signal is said to be oversampled by a factor:

$$OSR = fs/2B$$

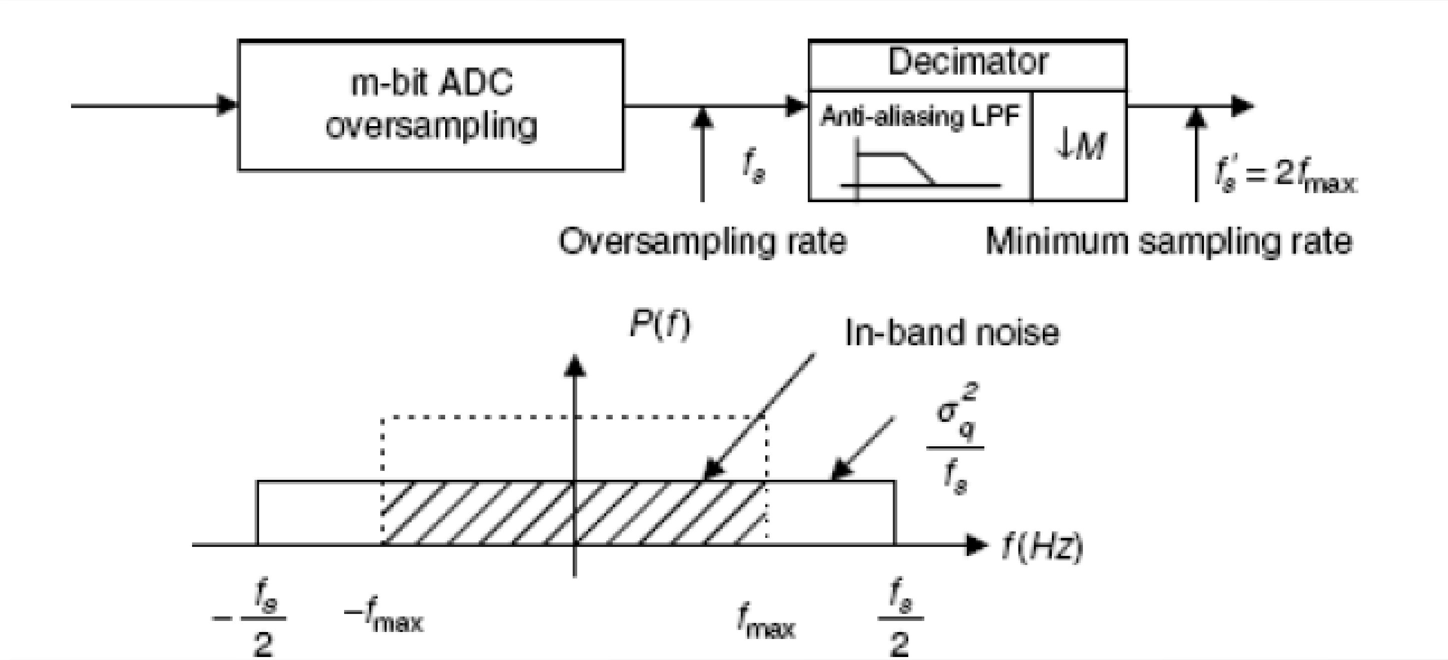
fs : sampling rate

B: bandwidth or highest frequency of the signal

(Nyquist rate: 2*B*).

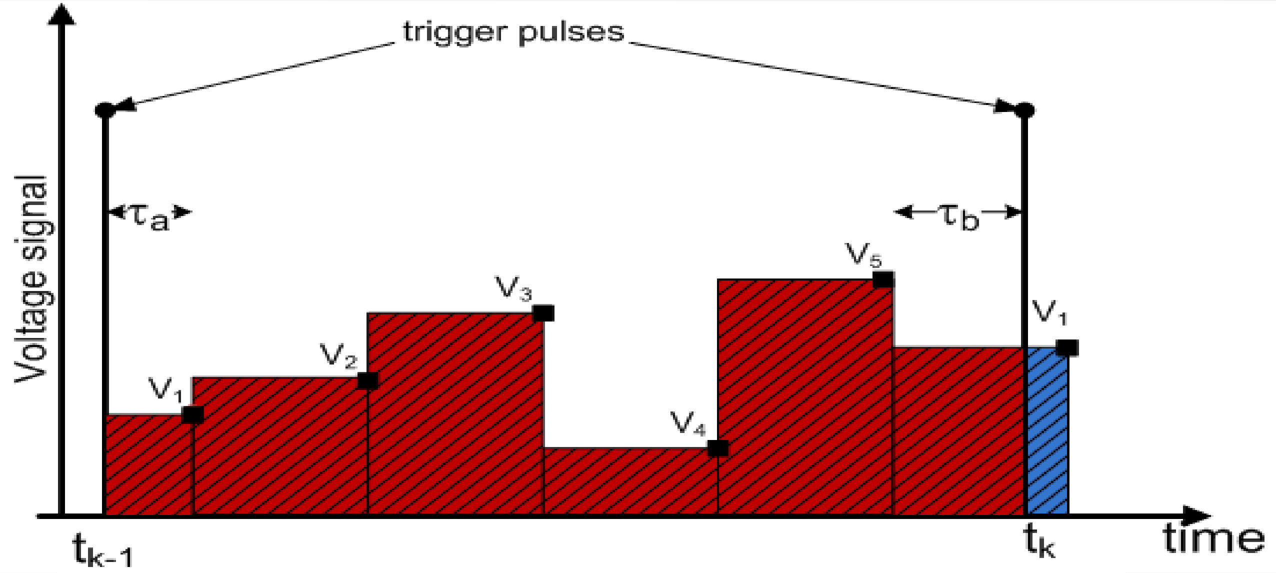
Oversampling Background

Oversampler architecture



Oversampling Background

Digital integration in magnetic measurement as oversampling



$$\Delta\varphi_k = V_{k_1} \cdot \tau_{a_k} + \sum_{i=2}^{N} V_{k_i} \cdot \tau_i + V_{(k+1)_1} \cdot \tau_{b_k}$$

Oversampling benefits

Oversampling aids in:

- Cheaper higher-resolution integrators
- Anti-aliasing
- Noise reduction

Oversampling for cheaper integrators

- In digital magnetic measurements, oversampling is exploited in order to achieve cheaper higher-resolution integrators
- The quantization noise is reduced by OSR
- To implement a (n+m)-bit integrator, a n-bit ADC can run at 2^{2m} times the flux trigger frequency (OSR= 2^{2m}), $m=0.5\log_2(OSR)$

Oversampling for cheaper integrators

- In digital magnetic measurements, oversampling is exploited in order to achieve cheaper higher-resolution integrators
- For instance, to implement a 24-bit integrator, a 20-bit ADC can run at 256 times the flux trigger frequency
- Computing the flux on 256 20-bit voltage samples adds 4 bits to the resolution, producing a single flux increment with 24-bit resolution.
- Note: only if the signal contains perfect equally distributed noise

Oversampling in anti-aliasing

- Oversampling aids in anti-aliasing because realizable analog filters are very difficult to implement with the sharp cutoff necessary to maximize use of the available bandwidth

- By increasing the bandwidth of the sampled signal, the anti-aliasing filter has less complexity and can be made less expensively by relaxing the requirements of the filter at the cost of a faster sampler.

Oversampling in anti-aliasing

- Once sampled, the signal can be digitally filtered and downsampled to the desired sampling frequency.

- In modern integrated circuit technology, digital filters are much easier to implement than comparable analog filters of high order.

Oversampling in noise reduction

- If multiple samples are taken of the same quantity with a different (and uncorrelated) random noise added to each sample, then averaging N samples reduces the noise variance (or noise power) by a factor of $1/N$.

- This means that the SNR improves by a factor of 4 (6 dB or one additional meaningful bit) with an oversample factor of 4.

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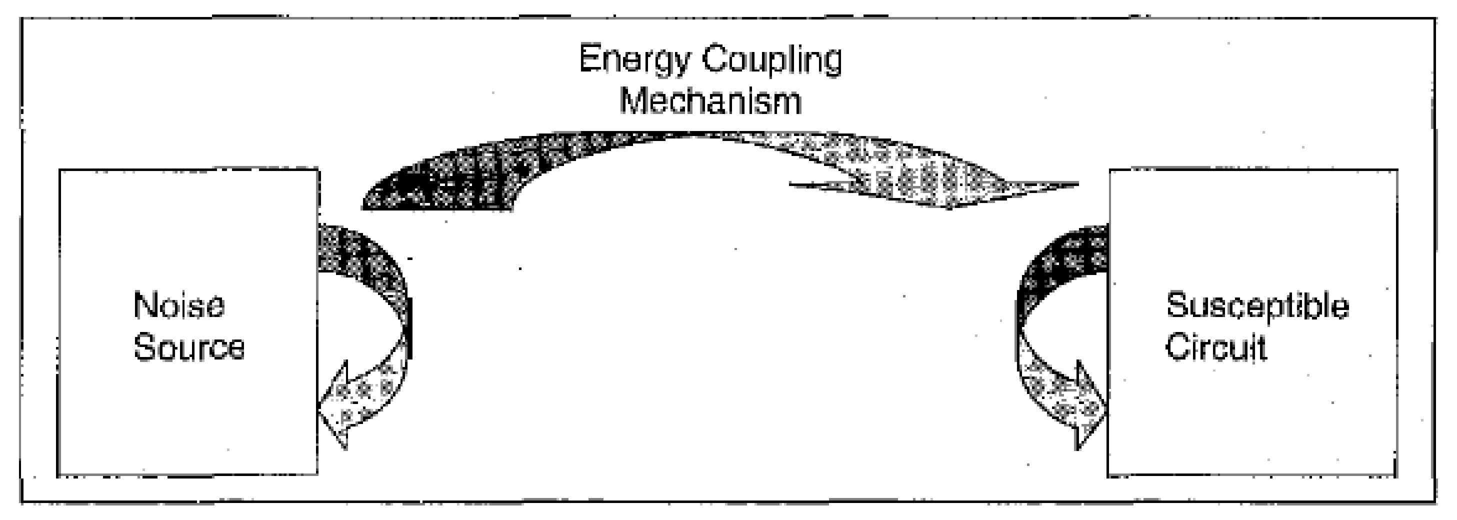
Design Hints:
Noise Effects

- Noise background
- Noise effects:
    - Coupling mechanisms
    - Shielding and Grounding
    - Ground loops

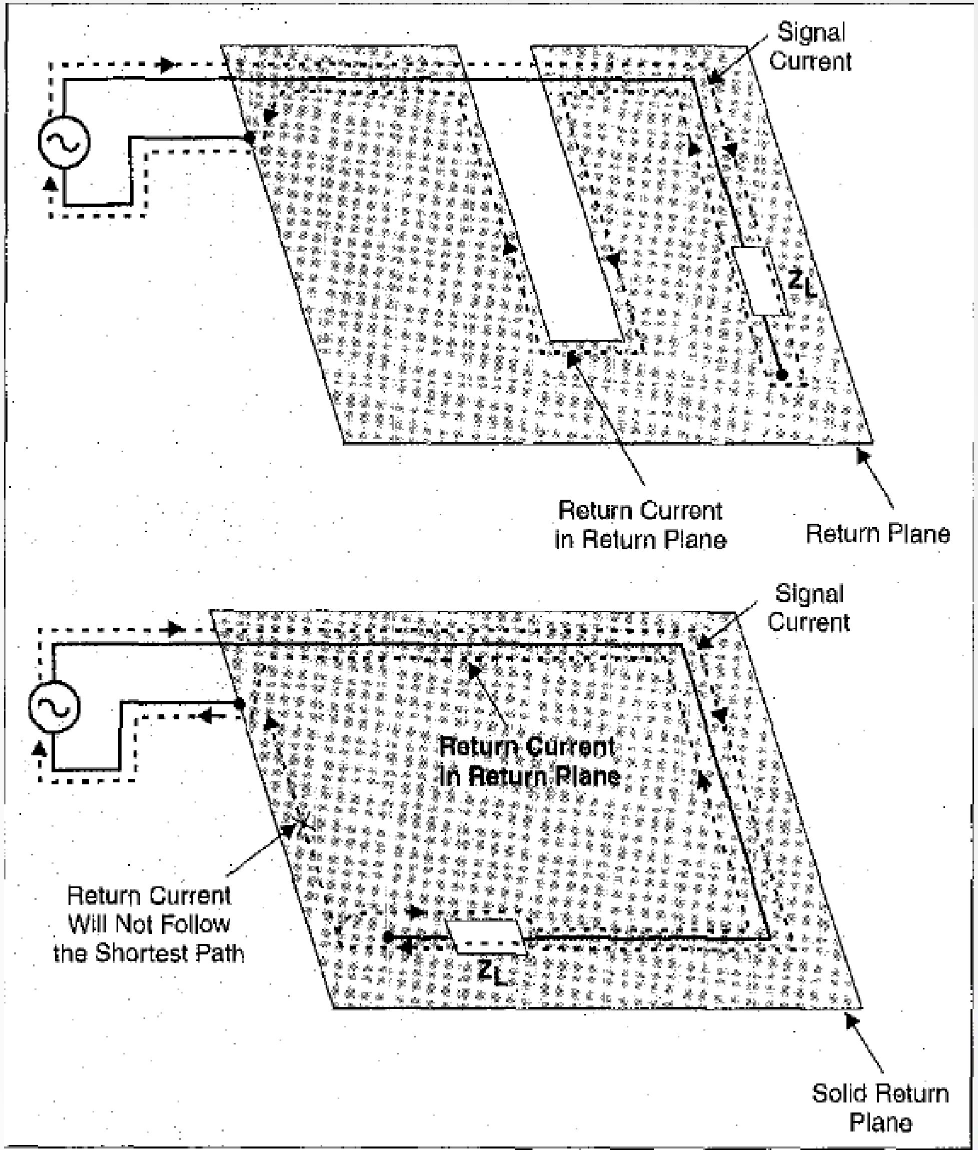
Noise Background

Noise components



Basic concepts in noise energy transfer

"Current follows the path of lowest impedance"



Basic concepts in noise energy transfer

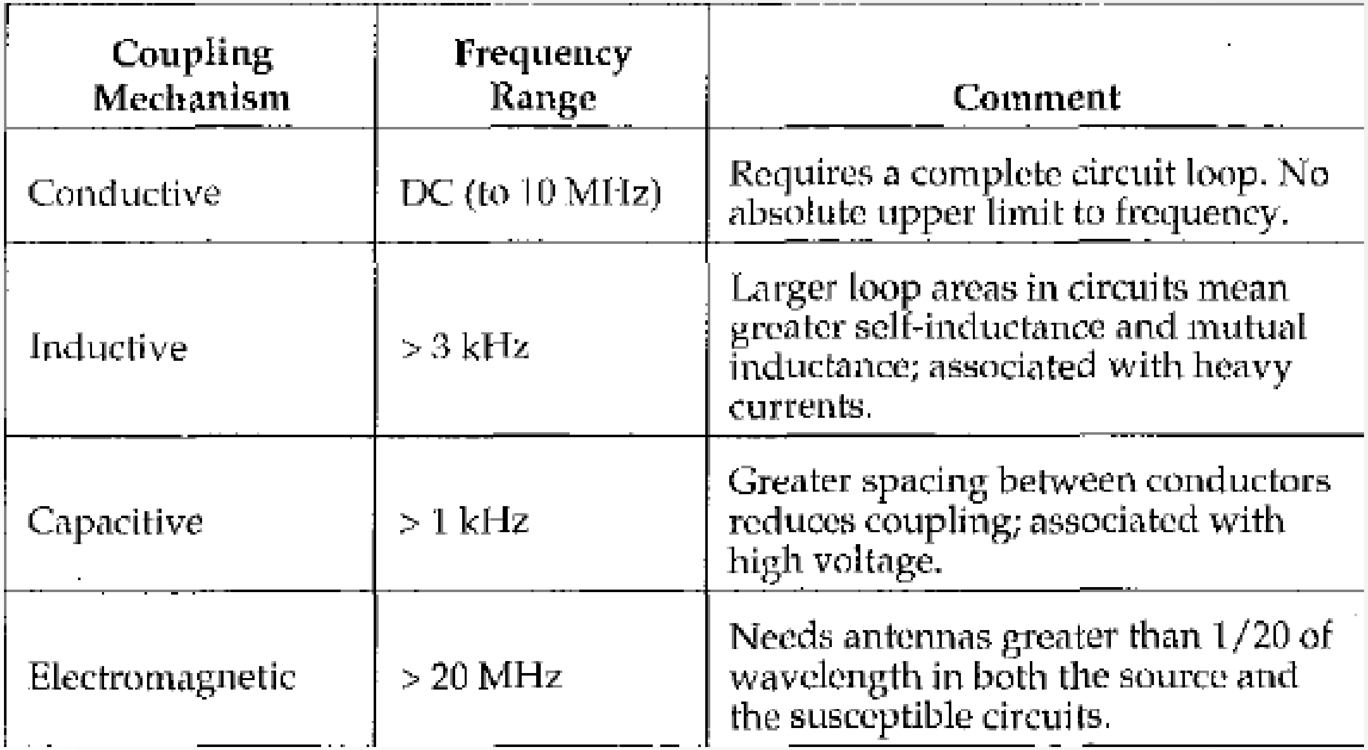
"Pseudo-impedance to diagnose noise coupling"

$$Zps = (dv/dt)/(di/dt)$$

- Useful for f > 5 kHz
- In general, $Zps \ll 377\ \Omega$

Noise Background

Mechanisms of noise coupling

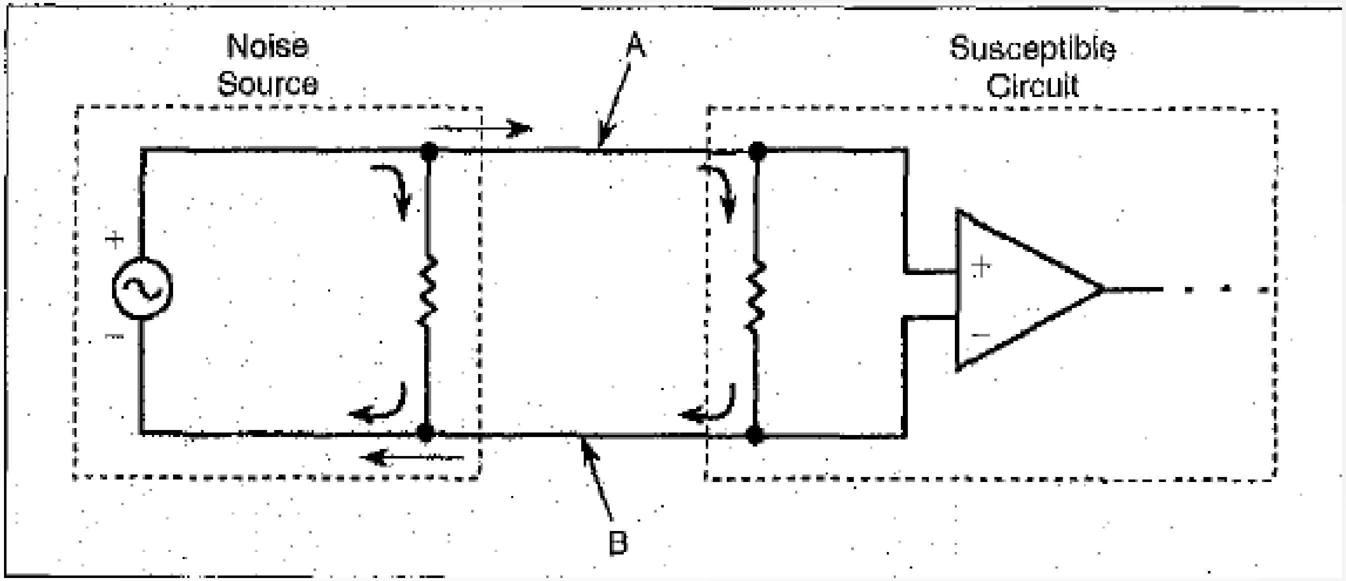


Coupling Mechanism	Frequency Range	Comment
Conductive	DC (to 10 MHz)	Requires a complete circuit loop. No absolute upper limit to frequency.
Inductive	> 3 kHz	Larger loop areas in circuits mean greater self-inductance and mutual inductance; associated with heavy currents.
Capacitive	> 1 kHz	Greater spacing between conductors reduces coupling; associated with high voltage.
Electromagnetic	> 20 MHz	Needs antennas greater than 1/20 of wavelength in both the source and the susceptible circuits.

Conductive coupling

A,B

- unintentional: to be break
- intentional: to be balanced to avoid ground loop (see later)

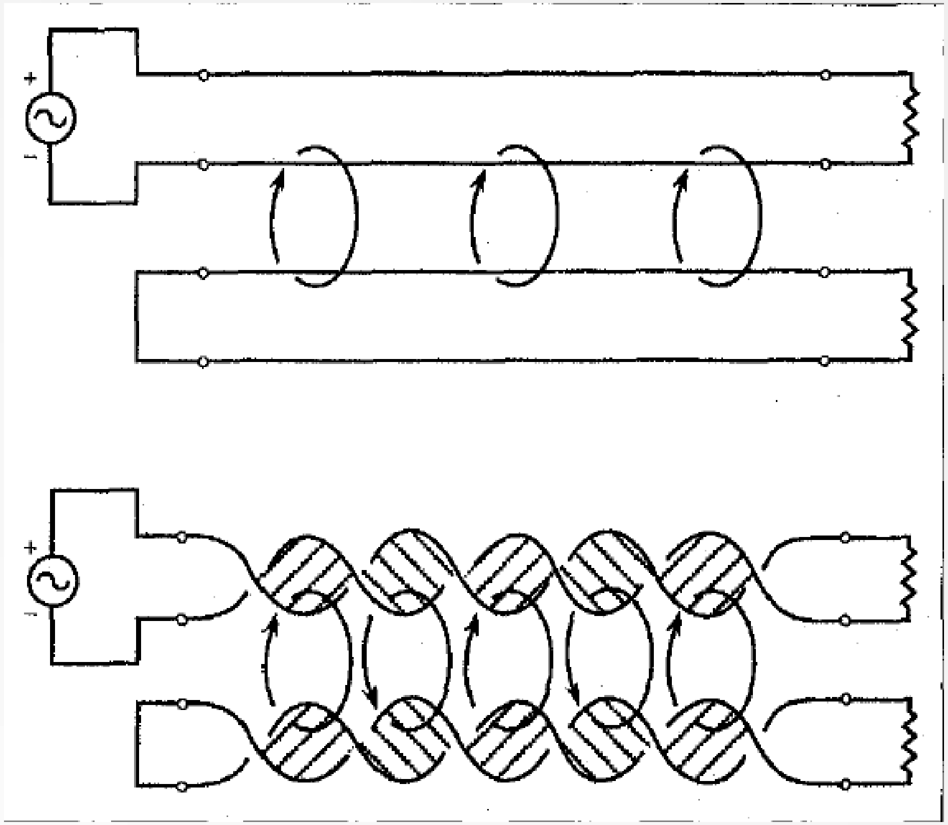


Inductive coupling

- Problem

- Solution:

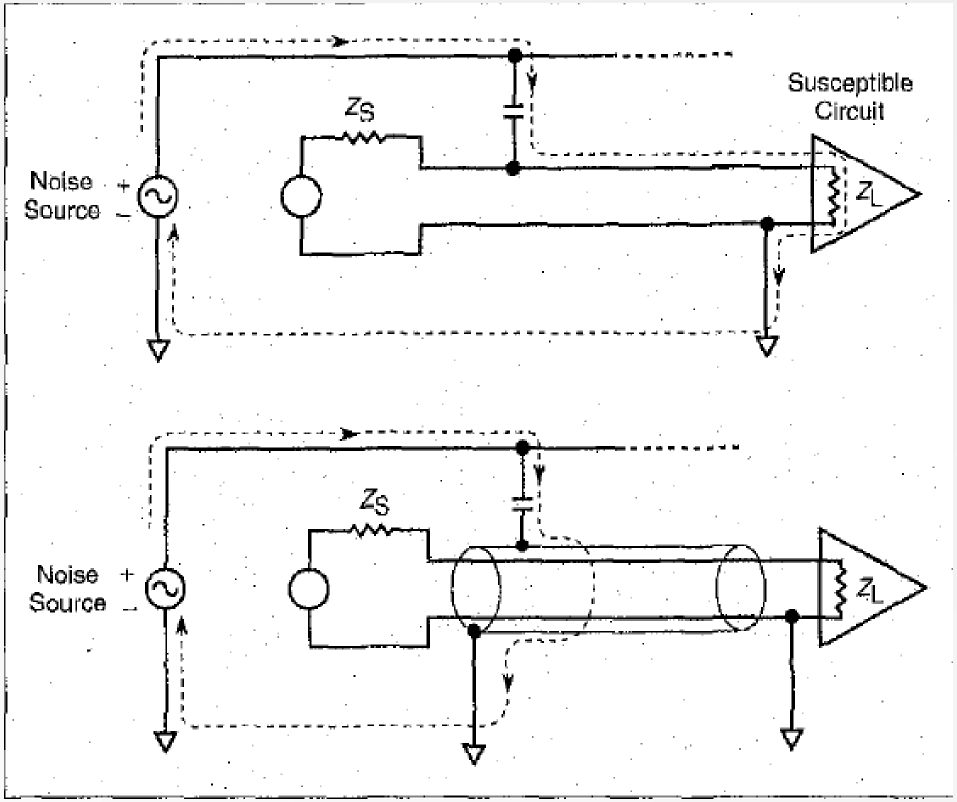
  Twisting



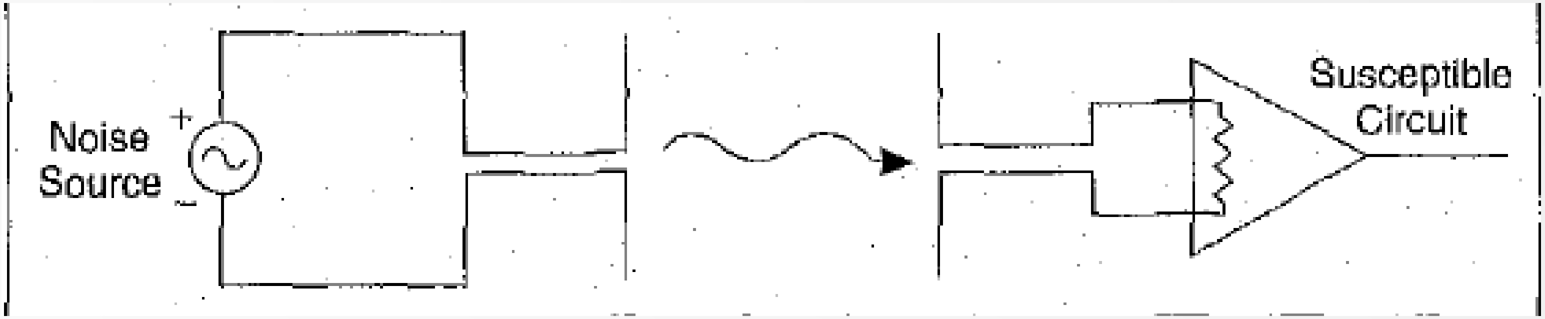
Capacitive coupling

- Problem

- Solution:
  Shielding
  +
  Grounding



Electromagnetic coupling – EMC -
(f > 20 MHz)



- Signal conductors > $\lambda/20$
- Zps = 377 Ω (actually between 100 and 500)
- Radiated EM energy requires antennas with an appreciable wavelength portion
- EMI (EM Interference) begins and ends as conductive

Reducing EMC

- Reduce bandwidth
- Use good signal routing
- Shield enclosures (completely closed condcting surfaces, openings leak EMI)
- Study a good book of EM compatibility (Clayton R. Paul, “Introduction to Electromagnetic Compatibility“, Wiley)

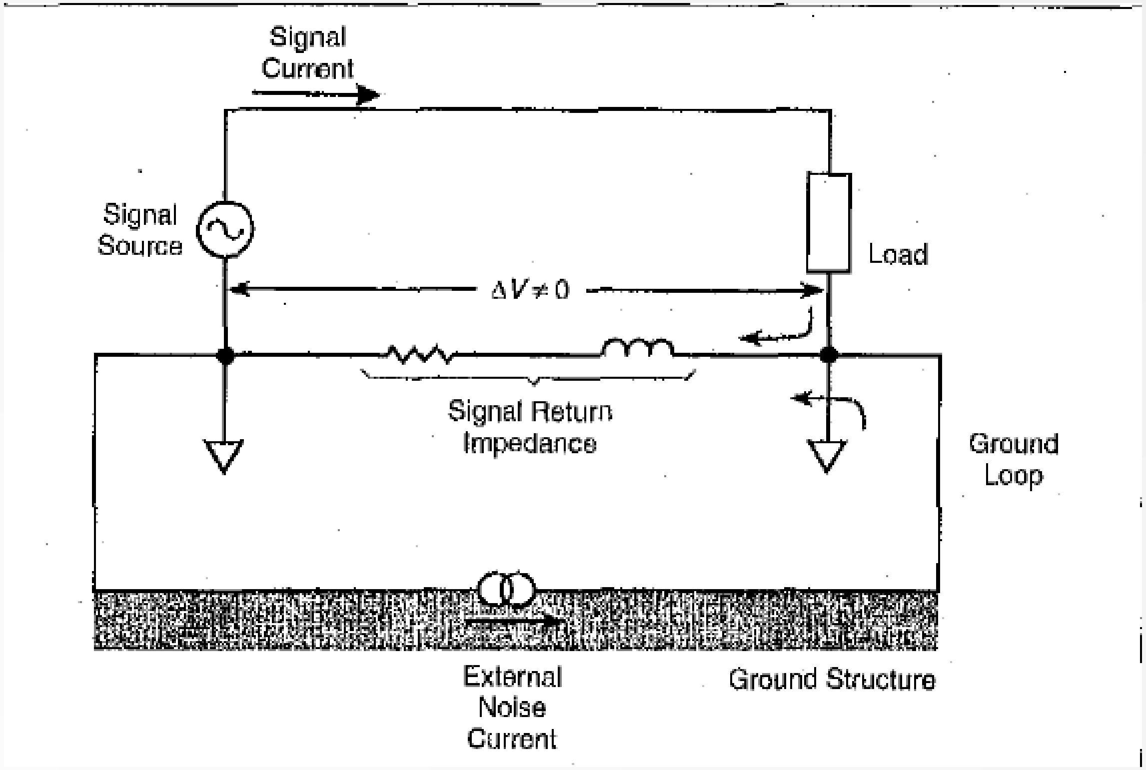
Ground loops (f< 10 MHz)

A *ground loop* is a complete circuit comprising the signal path and part of the ground structure.

Ground loops allow external currents in the ground structure:

- to generate potential differences between ground connections
- and to introduce noise in the signal circuit.

Ground loops



Eliminating ground loops

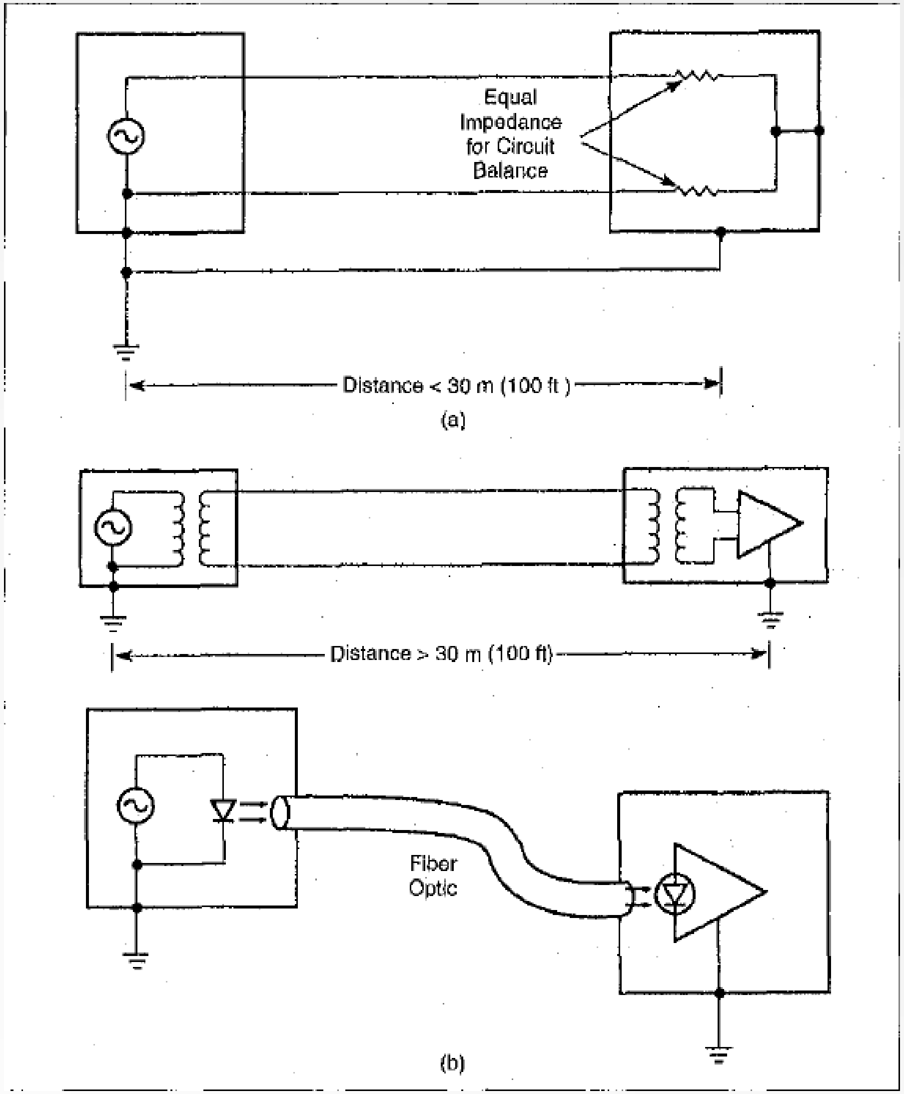
- < 30 m

Branches balance

- > 30 m

Decoupling:

  - transformers
  - optic fiber



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  - **Architecture evolution (20 min)**
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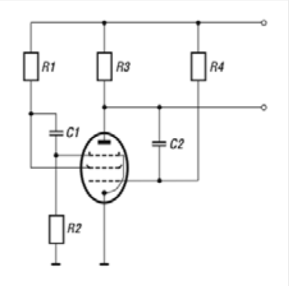
Architecture Evolution

- Analog integrator
- Voltage-to-frequency converter (VFC):
    - Performance analysis
    - VFC as 1st-order delta-sigma (Δ–Σ) ADC
    - The Δ–Σ performance
- CERN Fast Digita Integrator (FDI)
    - New SAR ADC + digital integration
    - CERN FDI: basic ideas
    - FDI architecture
    - FDI vs state-of-the-art VFC

Analog integrator (Miller)



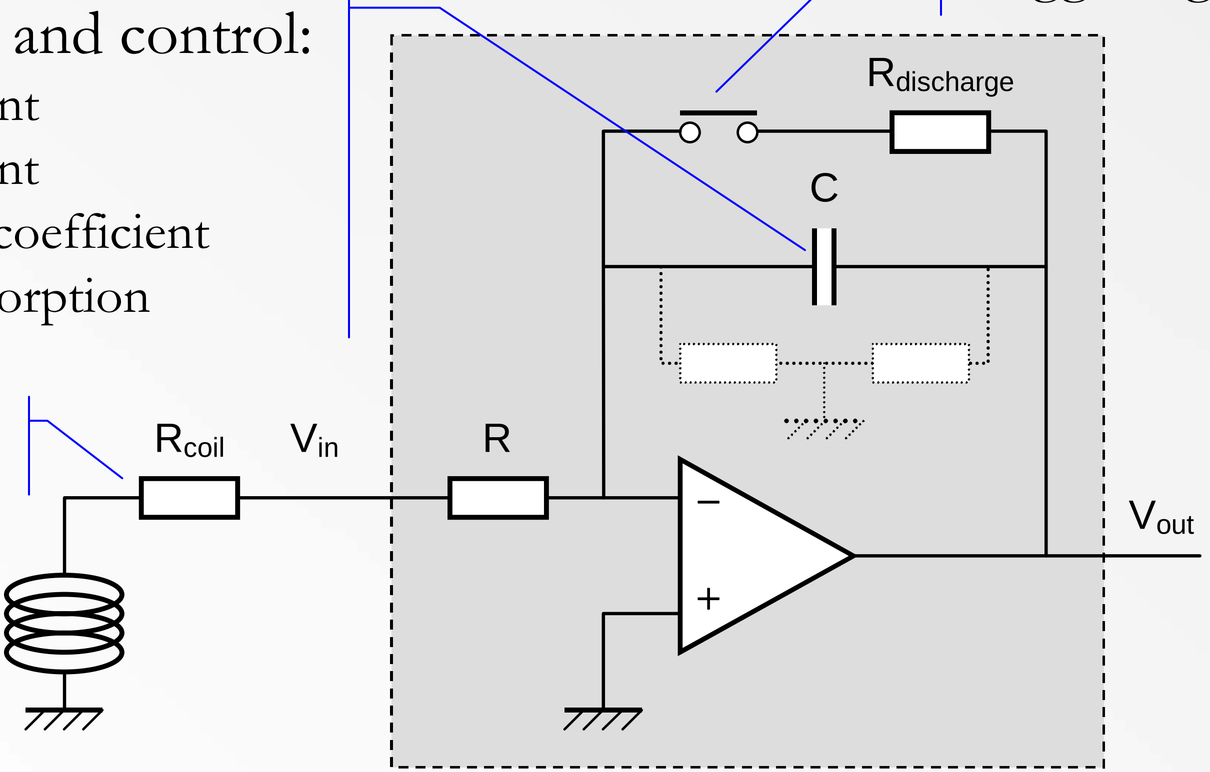
$$V_{out} = -\frac{1}{RC}\int_{-\infty}^{t} V_{in}\,dt$$

triggering

shielding and control:

- leakage current
- ground current
- temperature coefficient
- dielectric absorption

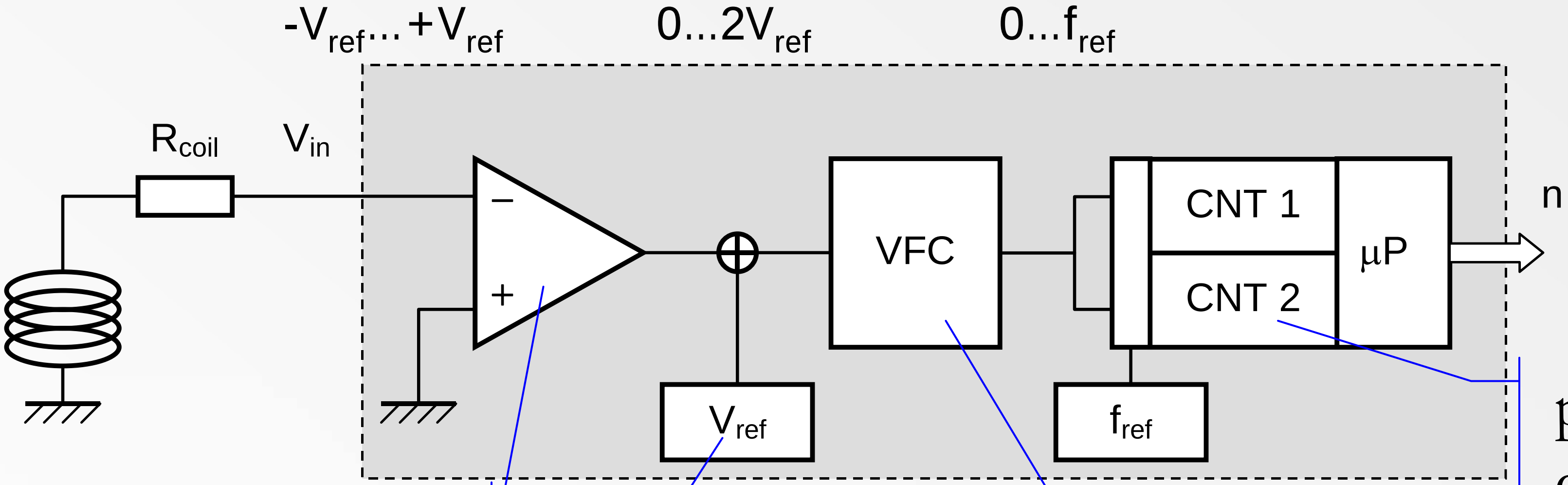
$R+R_{coil}$



simple, inexpensive, effective

accuracy limited by analog electronics

Voltage-to-Frequency Converter (VFC)



parallel counters

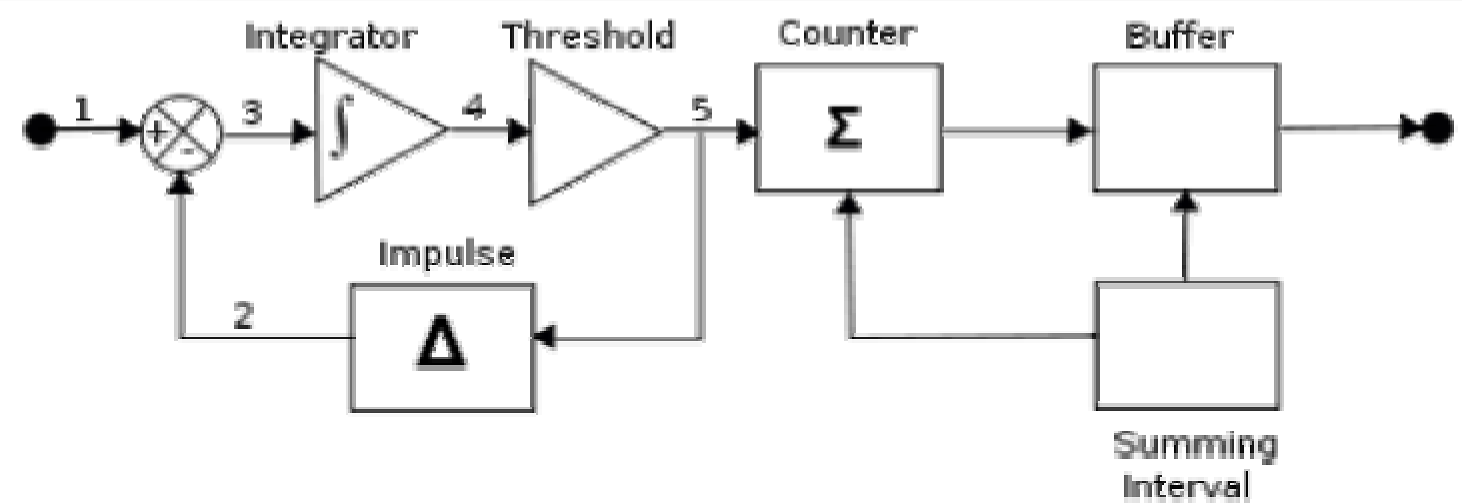
instrumentation amplifier (Z_{input})

Voltage to Frequency Converter

voltage offset

$$n = \int_{t_{start}}^{t_{end}} f dt = K_{VFC} \int_{t_{start}}^{t_{end}} V_{in} dt$$

VFC as 1^{st}-order Δ–Σ ADC

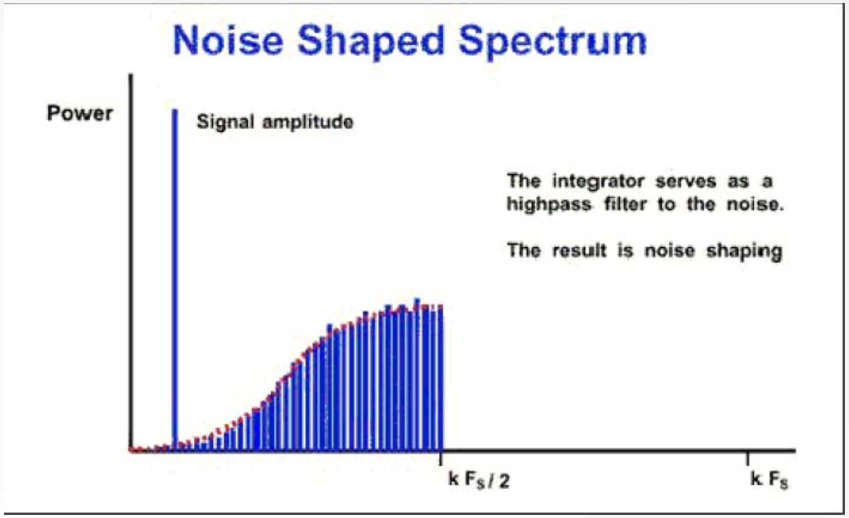


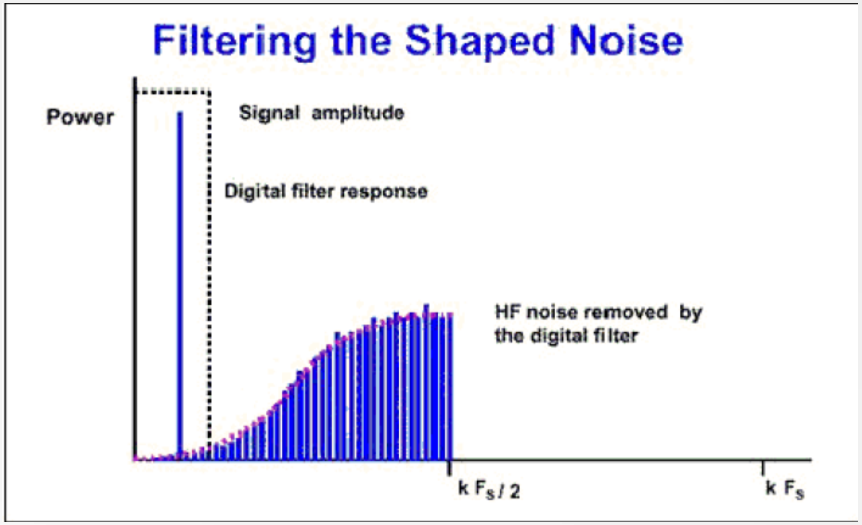
Delta-Sigma ADCs in noise reduction

- **Delta-Sigma (Δ–Σ)** converters produce disproportionately more quantization noise in the upper portion of their output spectrum

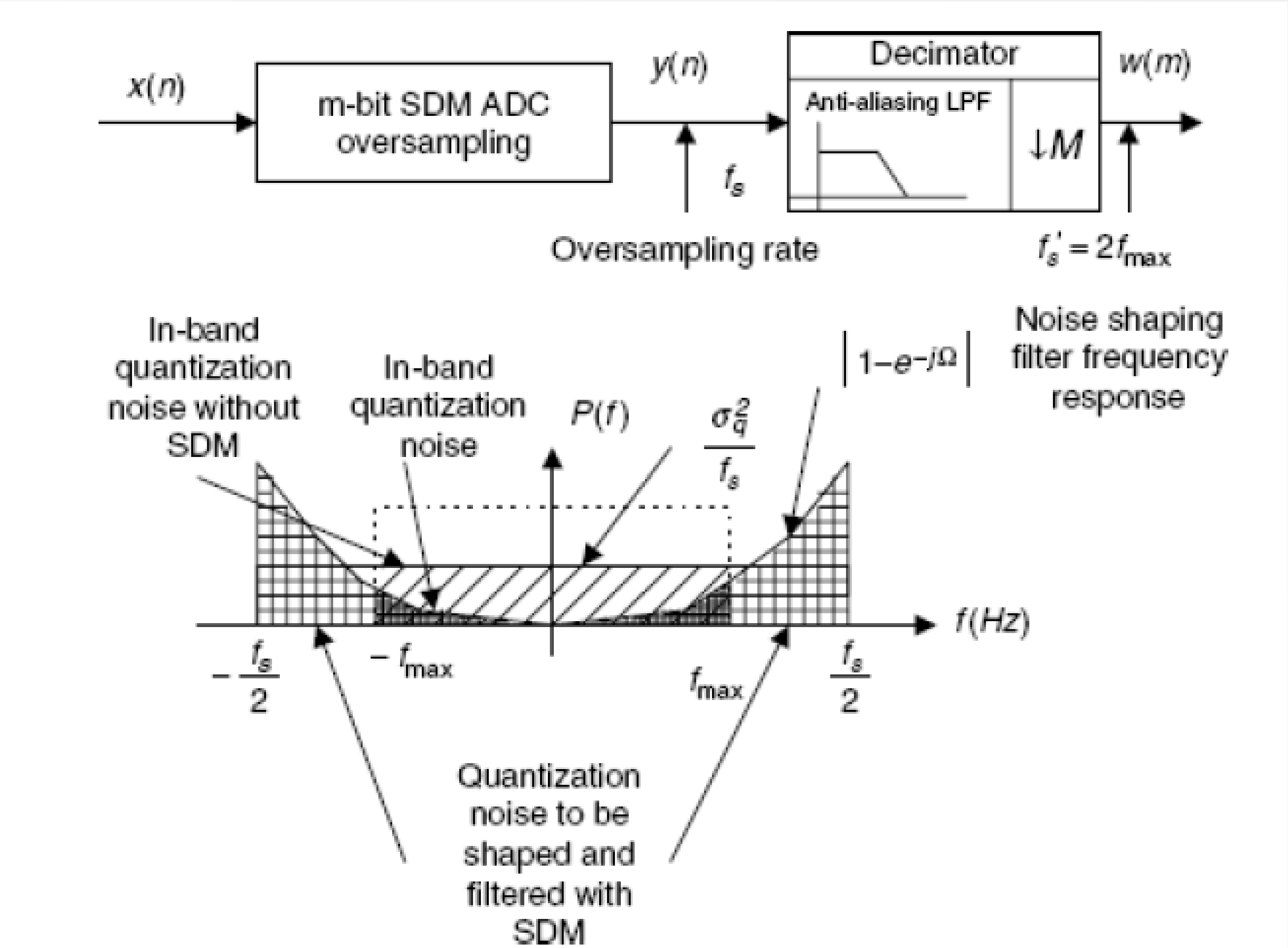
- The Δ–Σ use **noise shaping** to move the quantization noise to the higher frequencies

Delta-sigma noise shaping

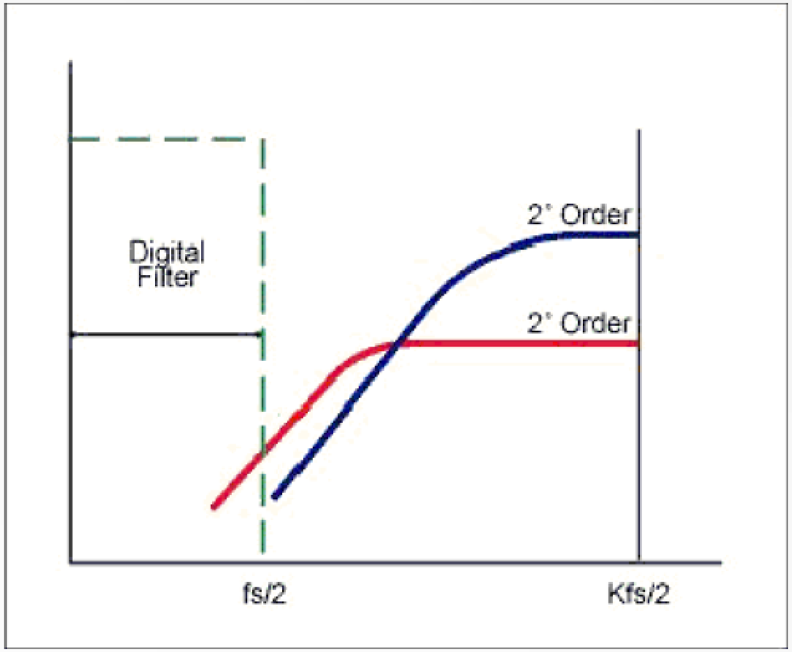


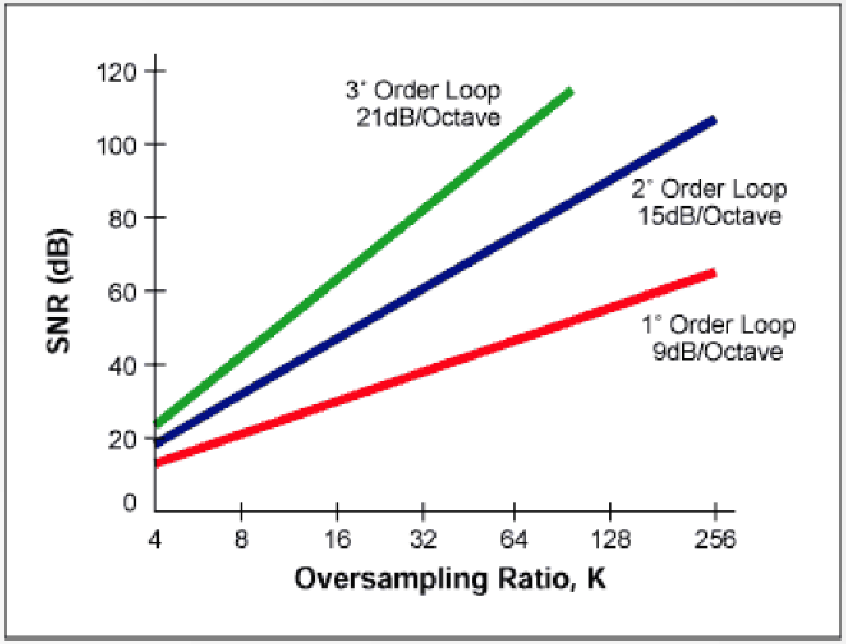


Delta-sigma noise shaping

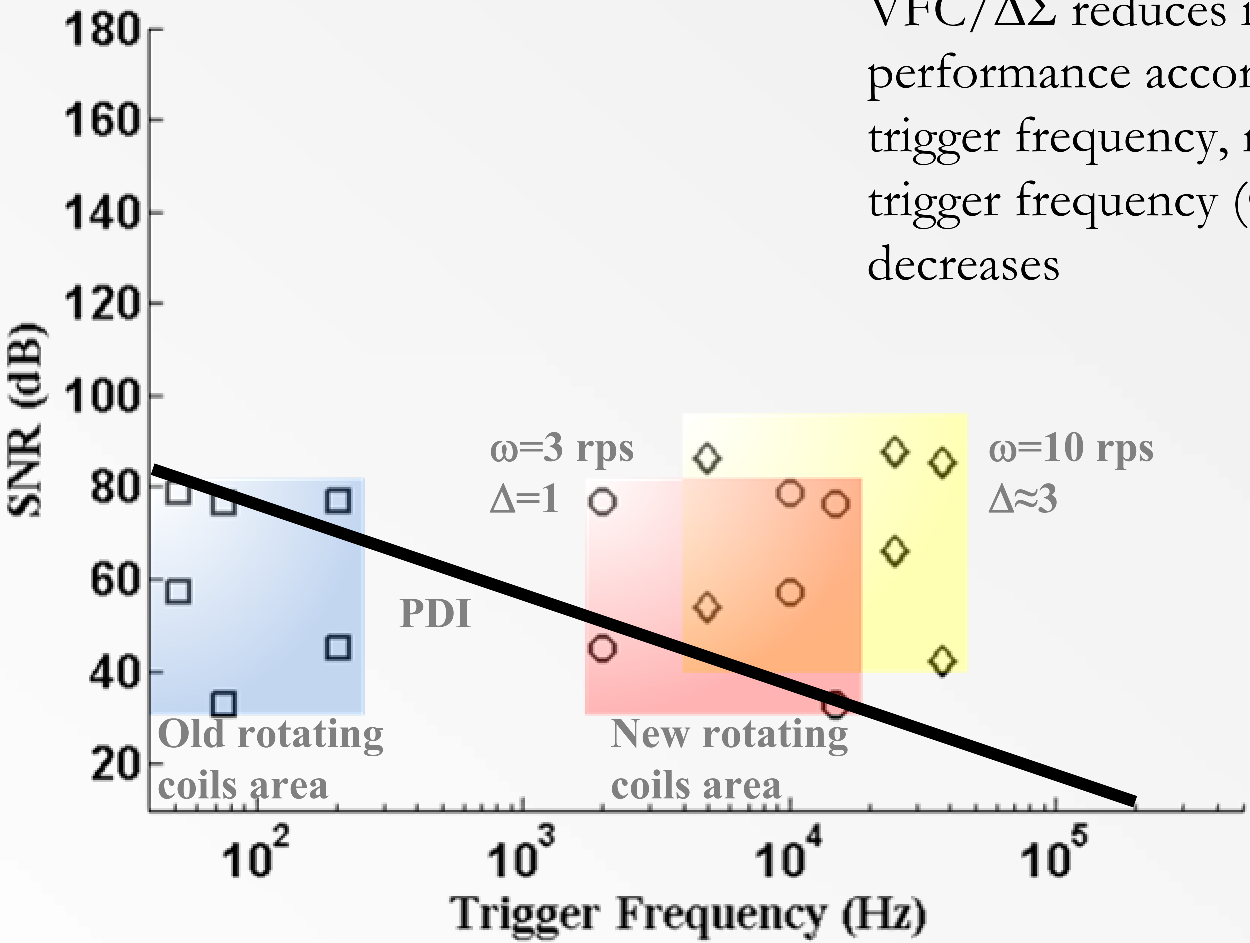


Δ–Σ order in noise shaping



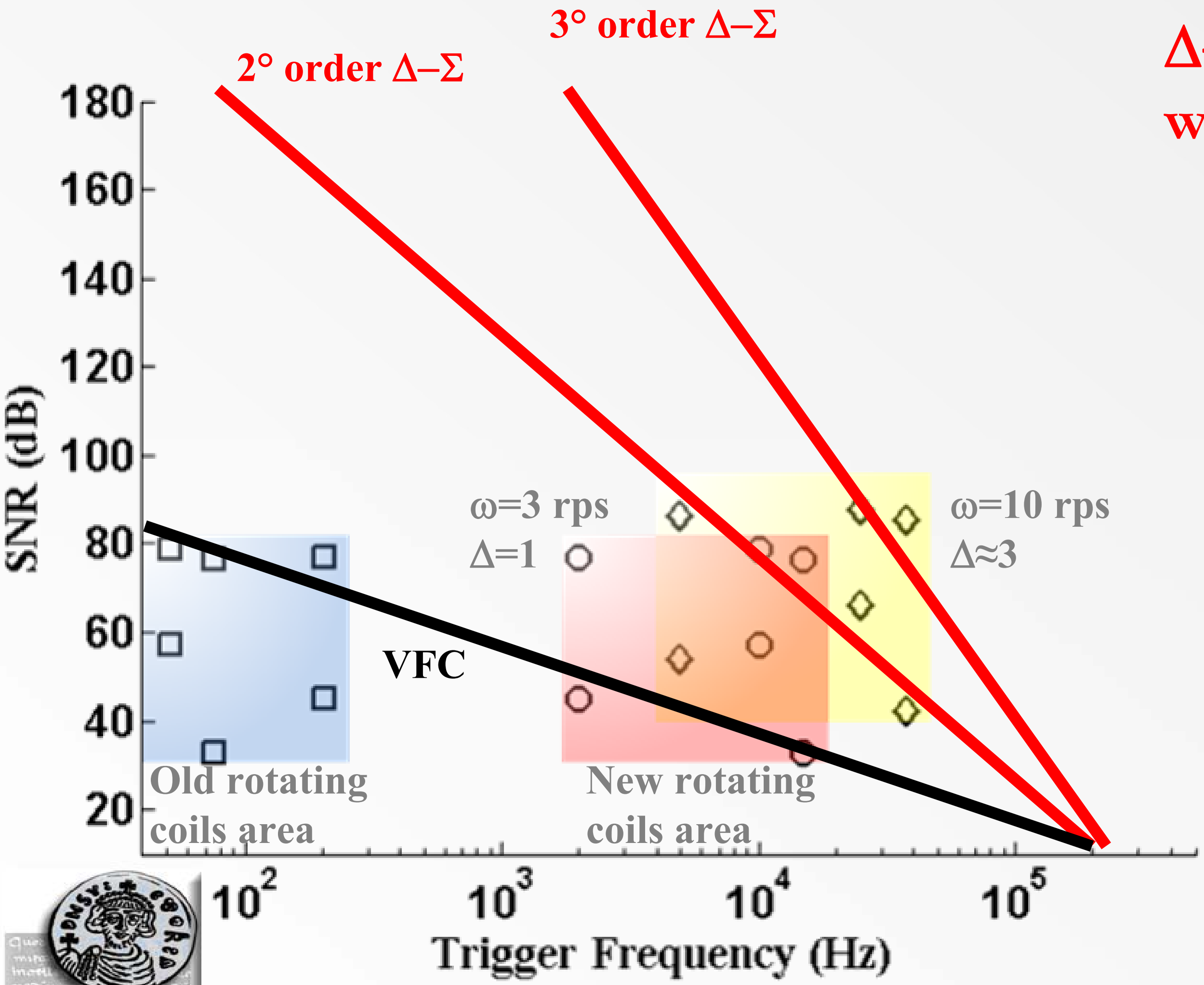


VFC/ΔΣ Performance Analysis



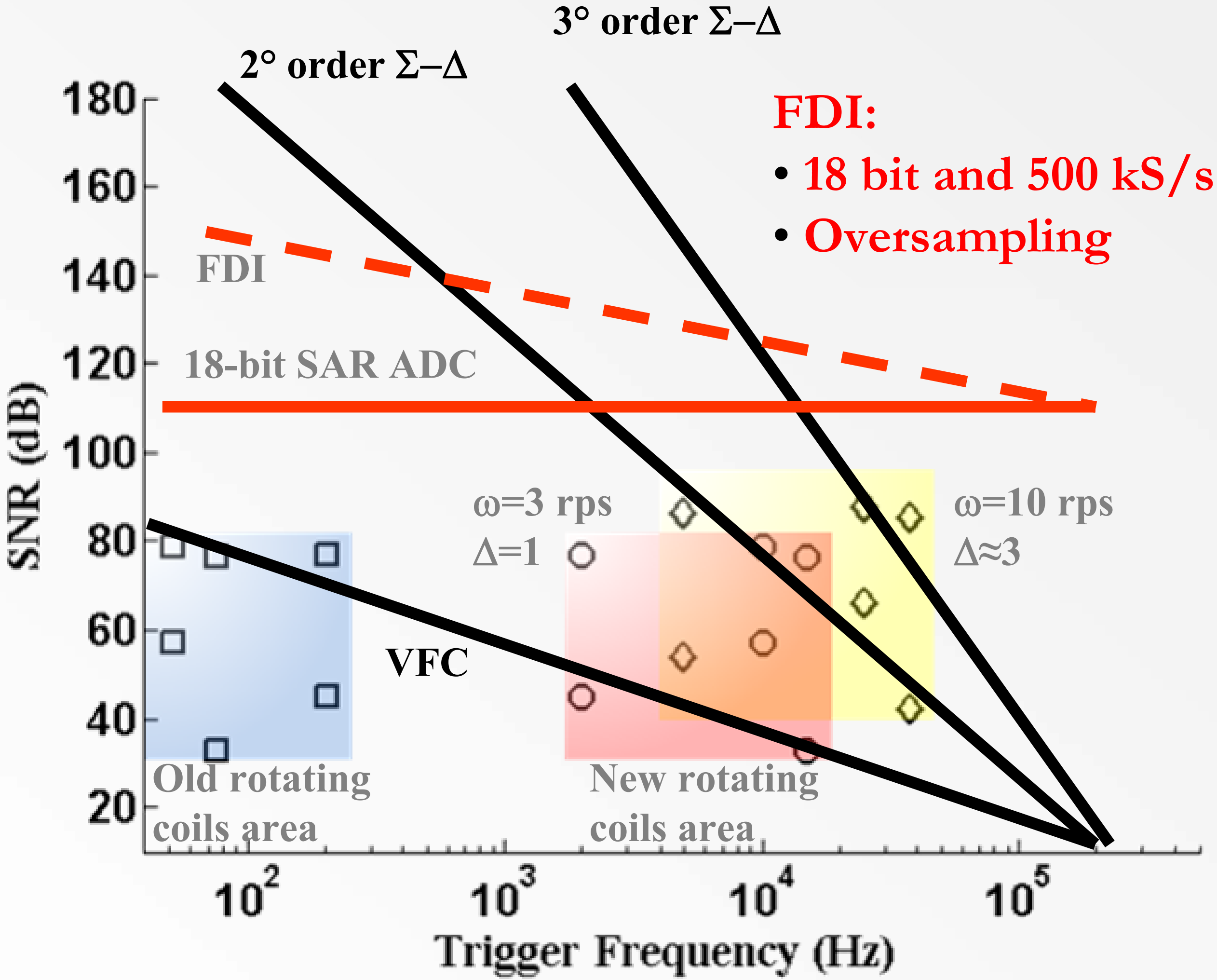
VFC/ΔΣ reduces its performance according to trigger frequency, namely as trigger frequency (OSR) decreases

VFC vs high-order Δ–Σ



Δ–Σ ADCs performs well at low frequency

Idea: High-resolution SAR ADC + Digital Integration



FDI:

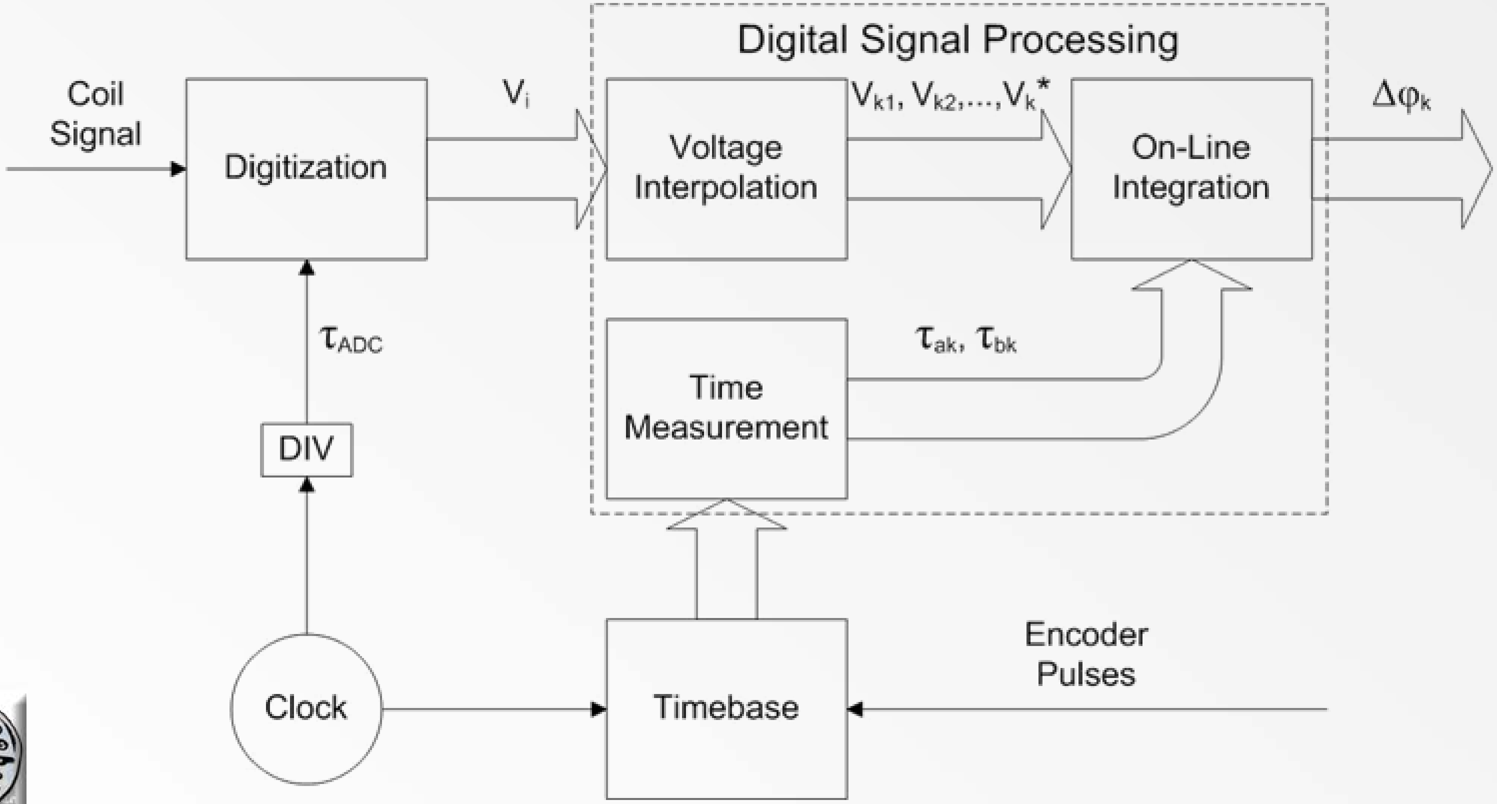
- 18 bit and 500 kS/s
- Oversampling

CERN Fast Digital Integrator

Basic ideas

- Oversampling conversion
- On-line integration
- Capabilities of general-purpose acquisition card with on-line signal processing
- High-resolution time measurement
- Differential measurement analog chain
- Self-calibration

Idea: CERN Fast Digital Integrator



Digital Signal Processing

Coil Signal

Digitization

V_i

Voltage Interpolation

$V_{k1}, V_{k2}, \ldots, V_k^*$

On-Line Integration

$\Delta\varphi_k$

τ_{ADC}

DIV

Time Measurement

τ_{ak}, τ_{bk}

Clock

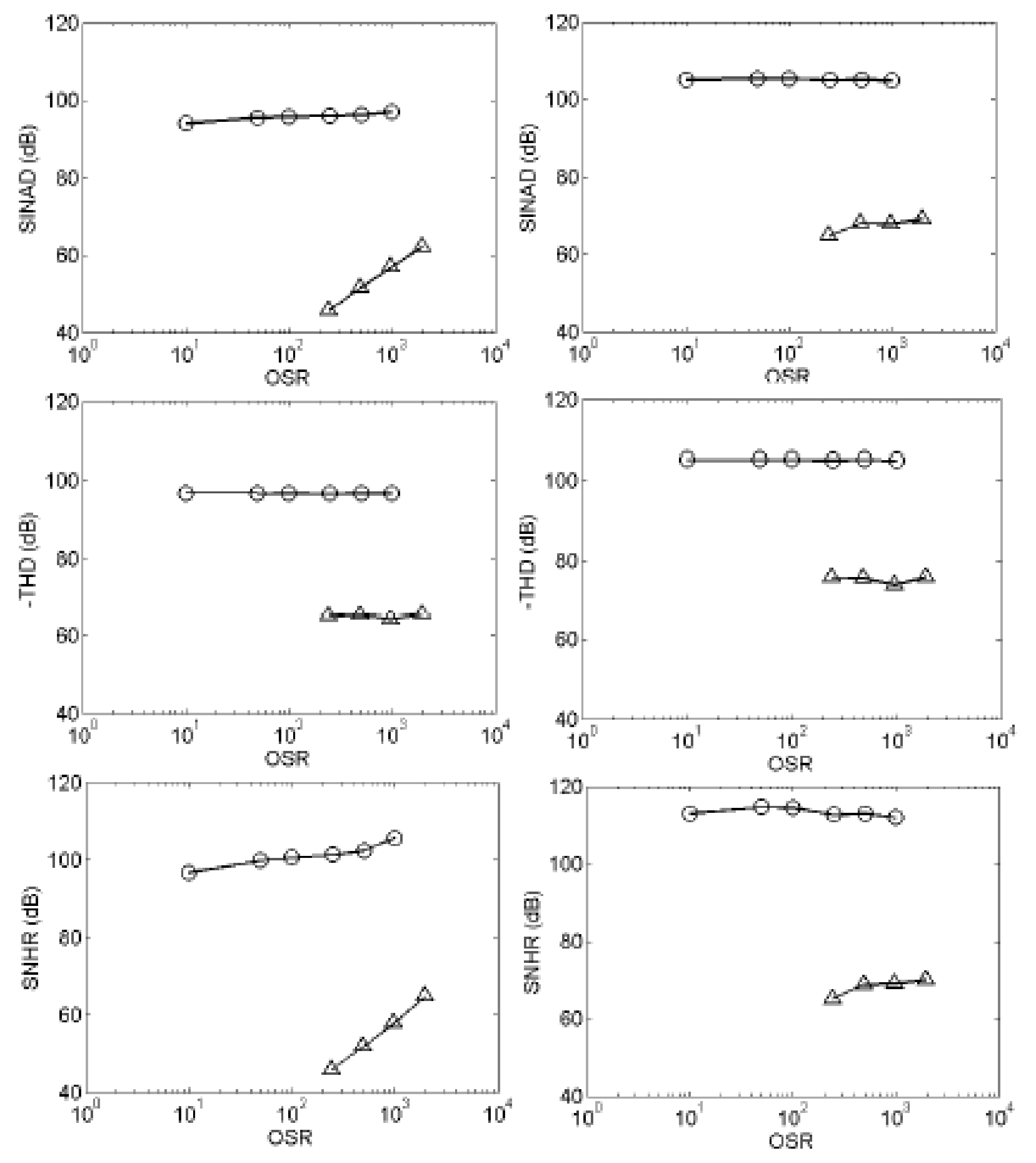
Timebase

Encoder Pulses

FDI vs state-of-the-art VFC

FDI: o

VFC: Δ



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  - **Static specs (10 min)**
  - Dynamic specs

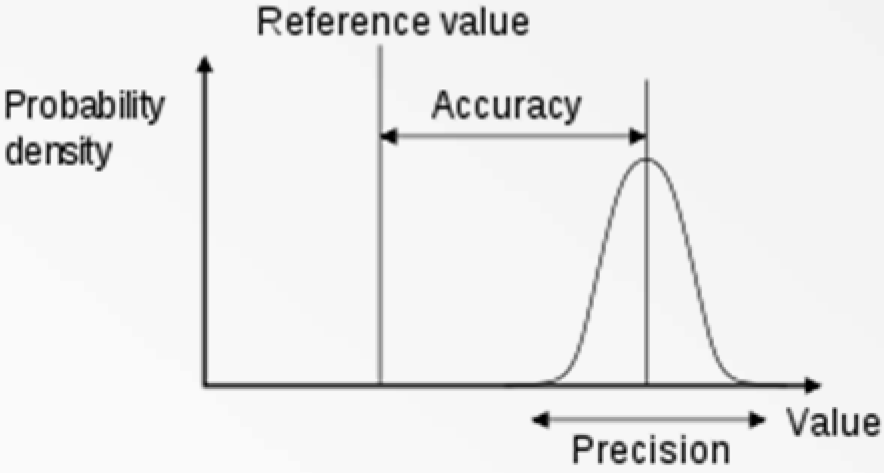
Metrological static specs

- Resolution
  - Analog:
    - 0.5 dev (smaller notches)
    - 0.2 devs (bigger notches)
  - Digital:
    - LSB= Vspan/2^n-1 (V)

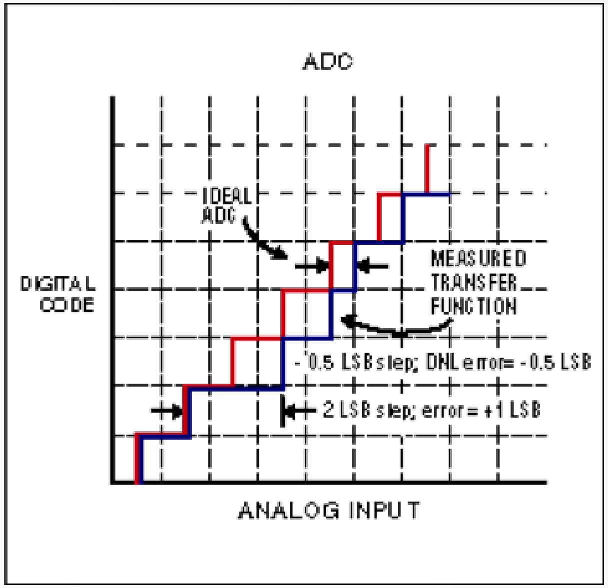
Accuracy vs Precision (VIM 2007)

High accuracy
Low precision

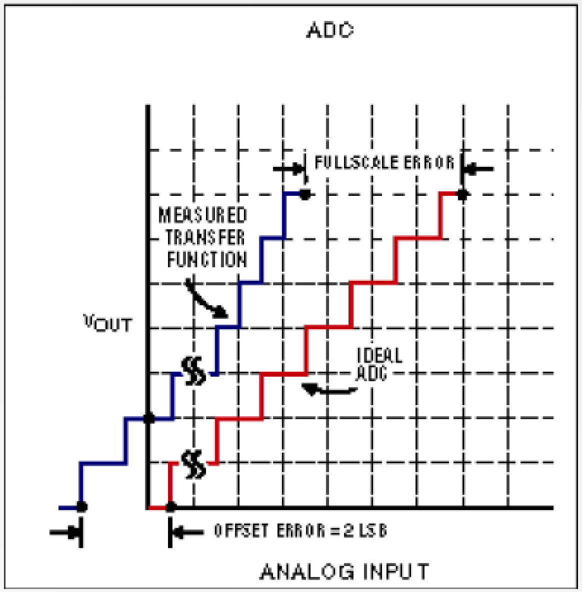
High precision
Low accuracy



Differential and Integral Nonlinearity



Differential NonLinearity DNL



Integral NonLinearity INL

$$DNL = \frac{\Delta_{DNL} - \Delta}{\Delta}\,(LSB),$$

$$INL = \frac{\Delta_{INL} - \Delta}{\Delta}\,(LSB)$$

Precision vs Repeatability

- Precision
  - Random variability in repeated measurements of the same measurand under **controlled measurement conditions** (due to **free** variation of influence parameters)
- Repeatability
  - Random variability in repeated measurements of the same measurand under **the same measurement conditions** (due to **residual** variation of influence parameters)

Repeatability vs Reproducibility

- Repeatability
  - Random variability in repeated measurements of the same measurand under **the same measurement conditions (**due to **residual** variation of influence parameters)
- Reproducibility
  - Random variability in repeated measurements of the same measurand under **different measurement conditions** (due to different lab, instrument, and so on)

Repeatability vs Stability

- Repeatability
  - Random variability in repeated measurements of the same measurand under the same measurement conditions
- Stability (in a given time interval)
  - Random variability in repeated measurements of the same measurand under the same measurement conditions **in a given time interval.**
  - Repeatability in the given time interval.

Traceability

- Traceability of a reference
  - The peculiar attribute of a reference of having been calibrated vs a certified reference
- Metrological chain
  - The uninterrupted chain of certified calibrations from our instrument up to the international unit of measure (about 6).

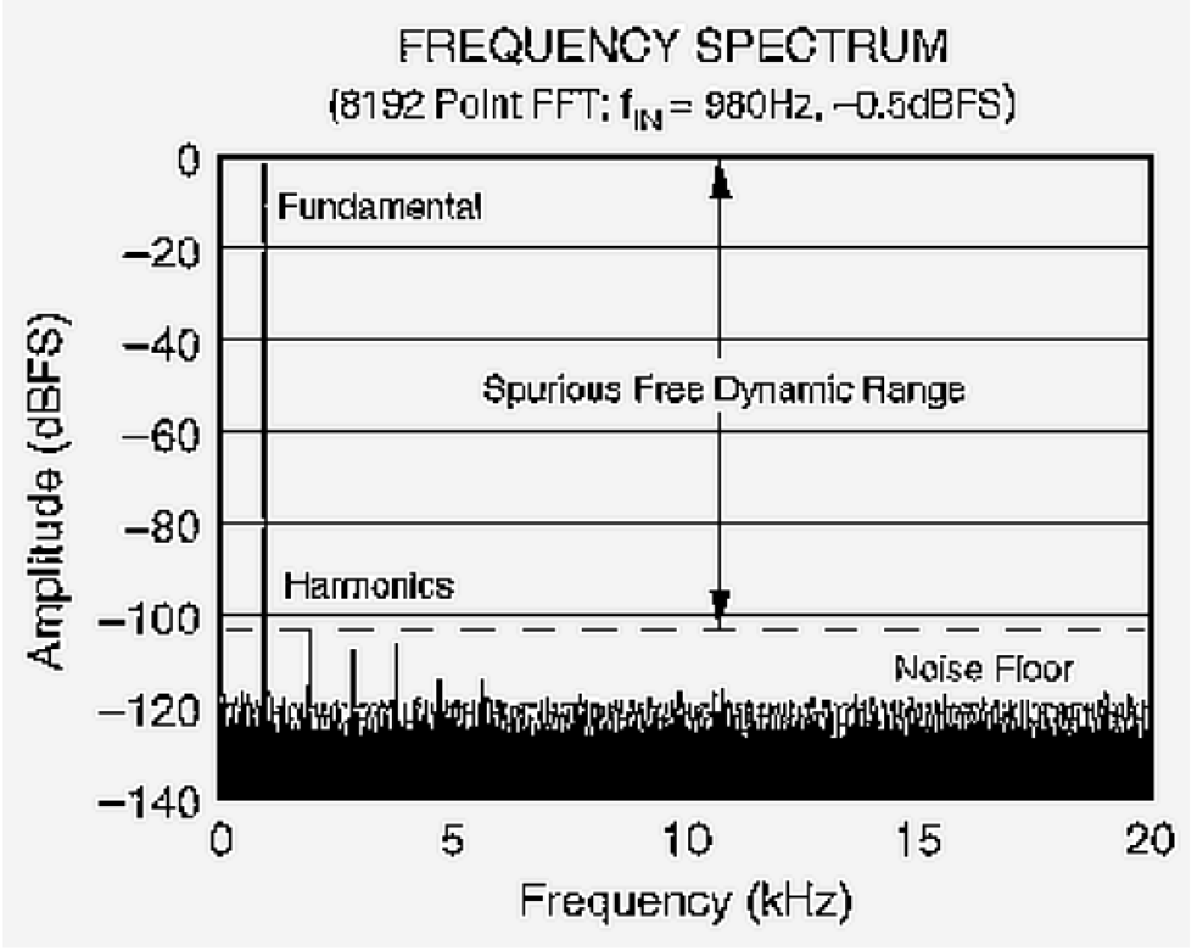
Outline

- Metrological specs (30 min)
  - Static specs
  - **Dynamic specs**

Dynamic Specs

- Signal-to-Noise Ratio (SNR) family
  - SNR, SINAD, SNHR, SFDR
  - THD,TSD
- Effective number of bits (ENOB)
- CMRR vs NMRR
- Dynamic DNL INL
- Step-response parameters

Signal-to-Noise Ratio (SNR) family

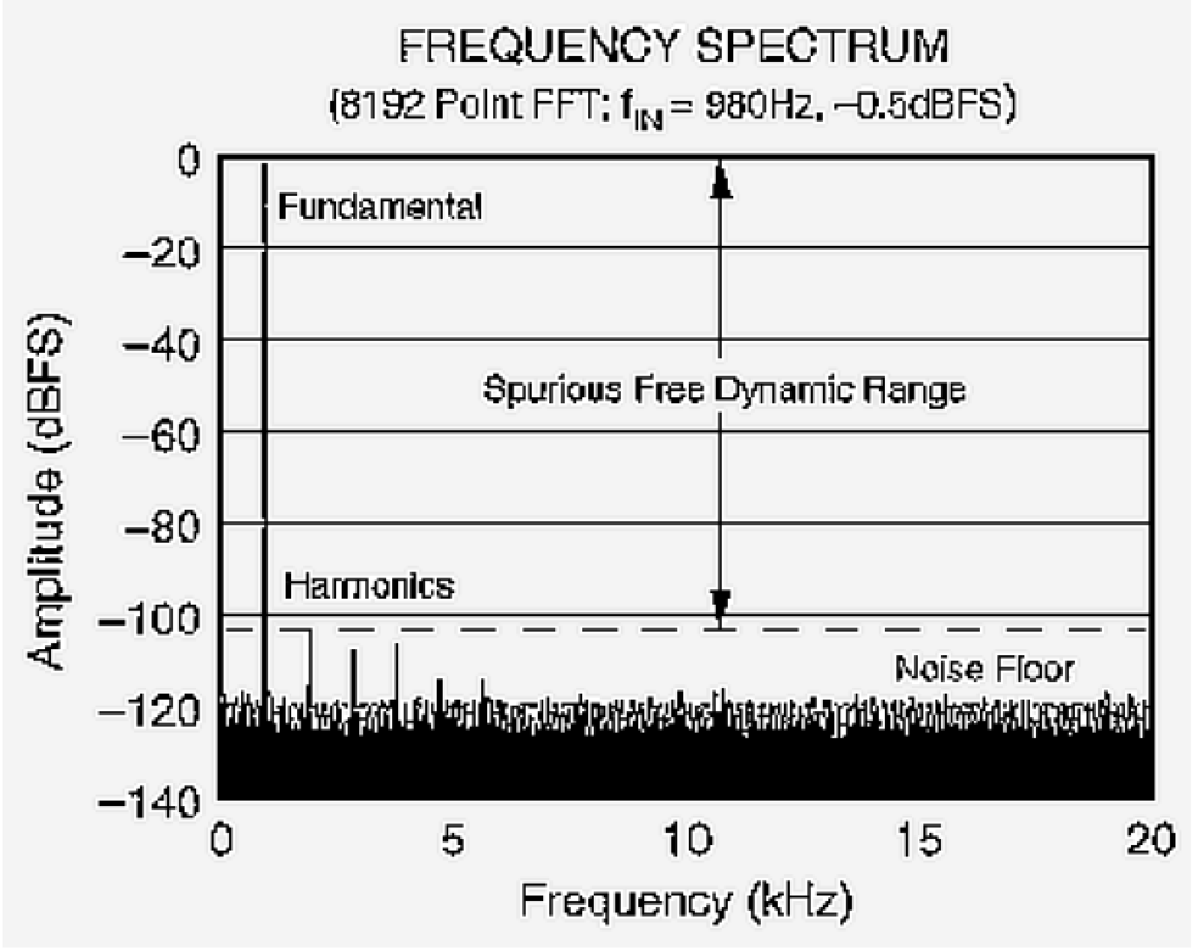


Signal-to-Noise Ratio (SNR) family

- **SNR** is a measure of the broadband noise introduced into the signal from the ADC and the sampling process. SNR compares the magnitude of the input sine wave to the sum of all other frequencies, except those representing harmonics of the fundamental

- **SNR and distortion (SINAD)** is the ratio of the power in the fundamental frequency bin to that in all other bins, including harmonics.
- **Signal to Non-Harmonic Ratio (SNHR)** is the ratio of the rms amplitude of the ADC output signal to the rms amplitude of the output noise, which is not harmonic distortion.
- **Spurious-free dynamic range (SFDR)** is the difference in magnitudes of the fundamental and the highest noise component.

Signal-to-Noise Ratio (SNR) family

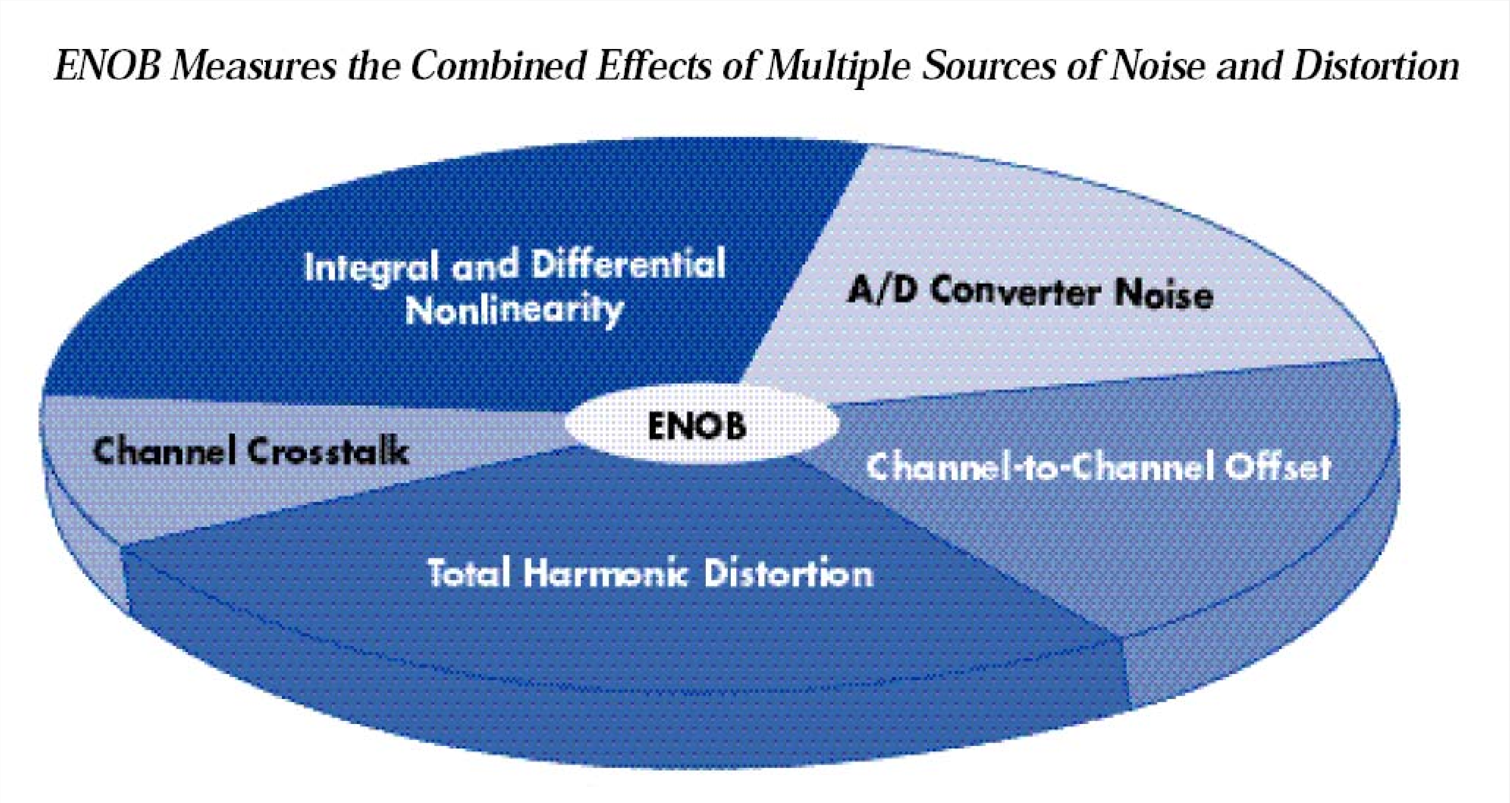


Signal-to-Noise Ratio (SNR) family

- **Total harmonic distortion (THD)** is the ratio of the fundamental to the sum of the harmonics.

- **Total Spurious Distortion (TSD)** is the rhe rss (root-sum-of-squares) of the spurious components in the spectral output of the ADC. TSD is often expressed as a dB ratio with respect to the rms amplitude of the output component at the input frequency.

Effective Number of Bits (ENOB)



ENOB Measures the Combined Effects of Multiple Sources of Noise and Distortion

Effective Number of Bits (ENOB)

ENOB is a measure of the SINAD used to compare actual ADC performance to an ideal N-bit ADC.

"the number of bits of an ideal ADC having the quantization noise equal to the overall noise of the actual ADC"

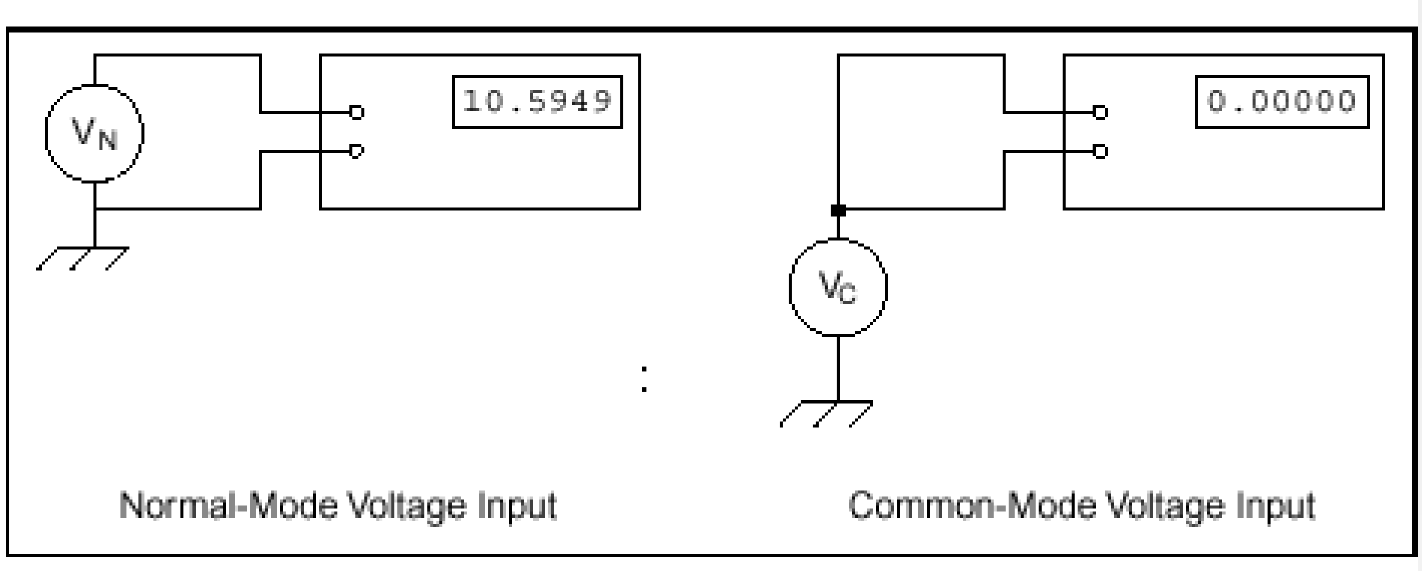
Useful because test engineers thinks quality of digital measurements in terms of achievable noise-free bits

Effective Number of Bits (ENOB)

$$ENOB = N - \log_2 \frac{rms\_noise}{ideal\_rms\_quantization\_error}$$

$$ideal\_rms\_quantization\_error = \frac{Q}{\sqrt{12}} = \frac{FullScale}{2^N \sqrt{12}}$$

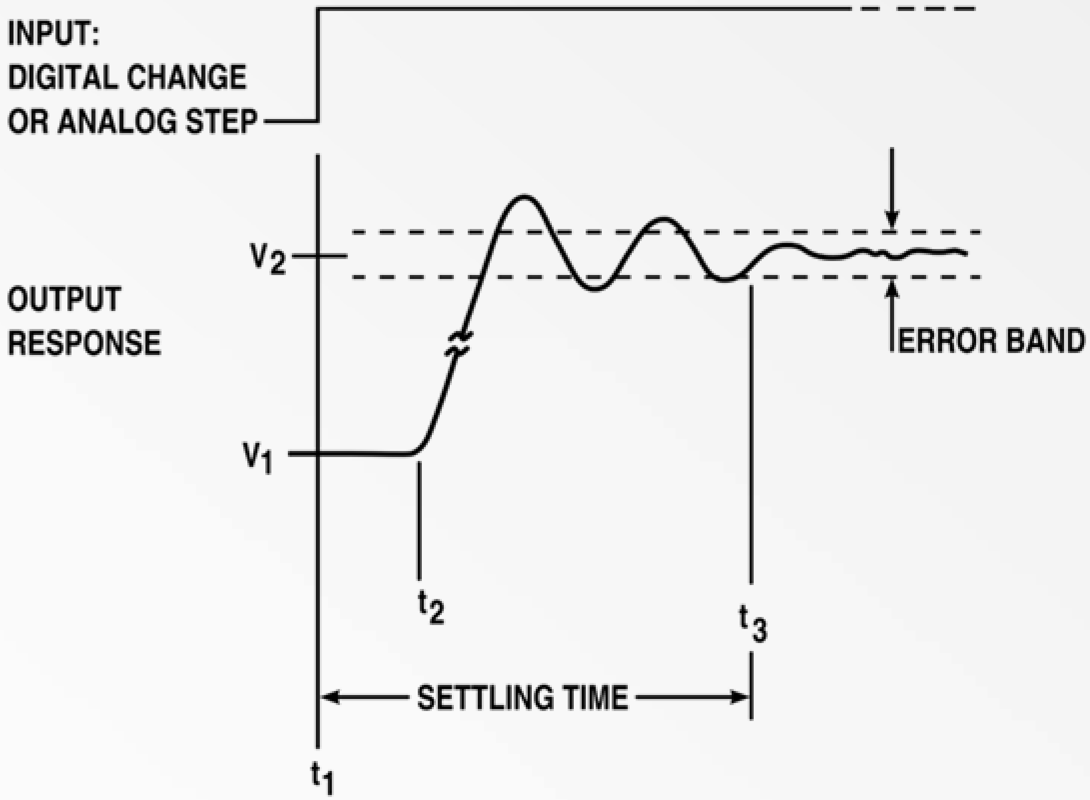
CMRR vs NMRR



$$\mathrm{NMRR} = 20\log(V_{measured}/V_{in})$$

$$\mathrm{CMMR} = 20\log(\text{Differential Gain}/\text{Common-Mode Gain})$$

Step-response Parameters



Outline

- Design hints
  - Integration effects (10 min)
  - Oversampling effects (10 min)
  - Noise effects (10 min)
  - Architecture evolution (30 min)

- Characterization hints
  - Metrological specs (30 min)
  - **Metrological tests (30 min)**

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Metrological tests

Static tests

- calibration (accuracy and precision, gain and offset)
- stability test
- transition level location (INL and DNL)
- parametric tests, and so on

Dynamic tests

Testing signals

- sine waves
- non sine wave signals - ramps, chirps, steps
- pulse and step signals

Basic dynamic test techniques use sine wave testing signal

- Sine wave curve fit test
- FFT test
- Histogram test

Sine wave curve fit test

Basic method for determination of ENOB

PROCEDURE:

- a full scale sine wave signal is digitized by the n-bit digitizer under test;
- a sine wave fit to the measured data using least-mean-square algorithm is assumed to represent the actual input signal;
- differences between the data record and best fit sine wave are assumed to be digitizer errors

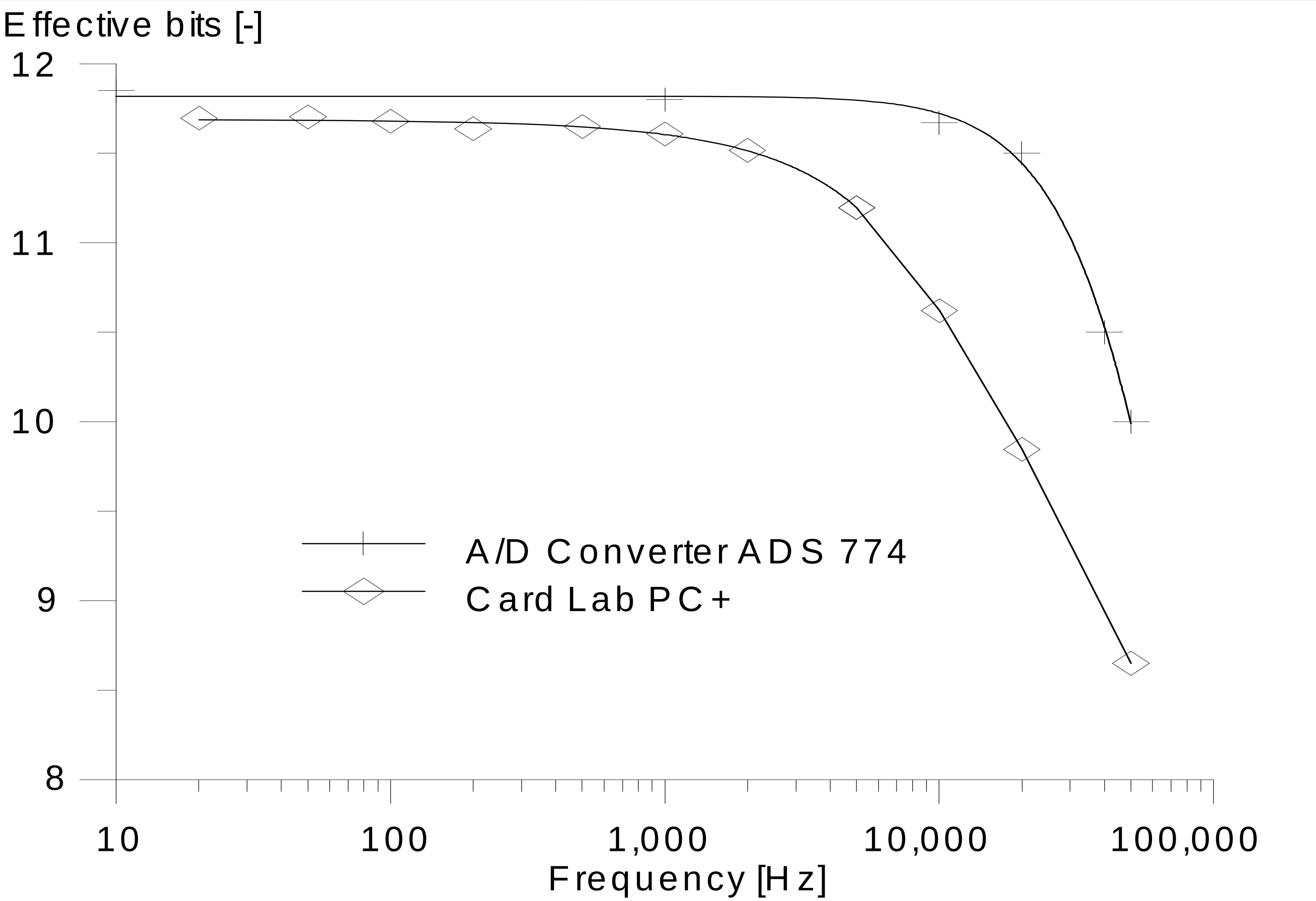
$$rms\_error = \sqrt{\frac{1}{M}\sum_{k=1}^{M}\left[X_k - A\sin\left(\omega t_k + P\right) - C\right]^2}$$

Sine wave curve fit test

Fitting algorithms

- **three parameter least squares fit** = “fixed-frequency” algorithm (frequency is known) - closed form, noniterative solution;
- **four parameter least squares fit** - iterative solution, diverge for bad initial estimates, or especially corrupted data

ENOB vs Frequency



Frequency Domain Analysis
FFT test

- The sampled data *x(n)* are transformed using Discrete Fourier Transform (DFT) to the discrete frequency spectrum *X(k)* according to the following formula:

$$X(k) = \sum_{n=0}^{M-1} x(n) e^{-j2\pi nk/M}$$

where *M* is number of samples.

- The DFT is usually calculated through the Fast Fourier Transform algorithm requiring *M* as a power of 2.

FFT test

The number of samples M determines the frequency resolution Δf and the noise floor NF

$\Delta f = f_S / M$,

where f_S is sampling rate

NF = - (6,02N + 1,76 +10 log (M/2)) (dB)

- Most dynamic performance parameters can be determined from the resulting FFT (SINAD, SNHR, SFDR, THD)
- ENOB can be calculated as:

$$ENOB = \frac{(SINAD - 1.76)}{6.02}$$

FFT test

- DFT assumes that the data record repeats with a period of M/f_S.
- If sampling is not coherent (the data record does not contain an integer number m of signal periods Tx):

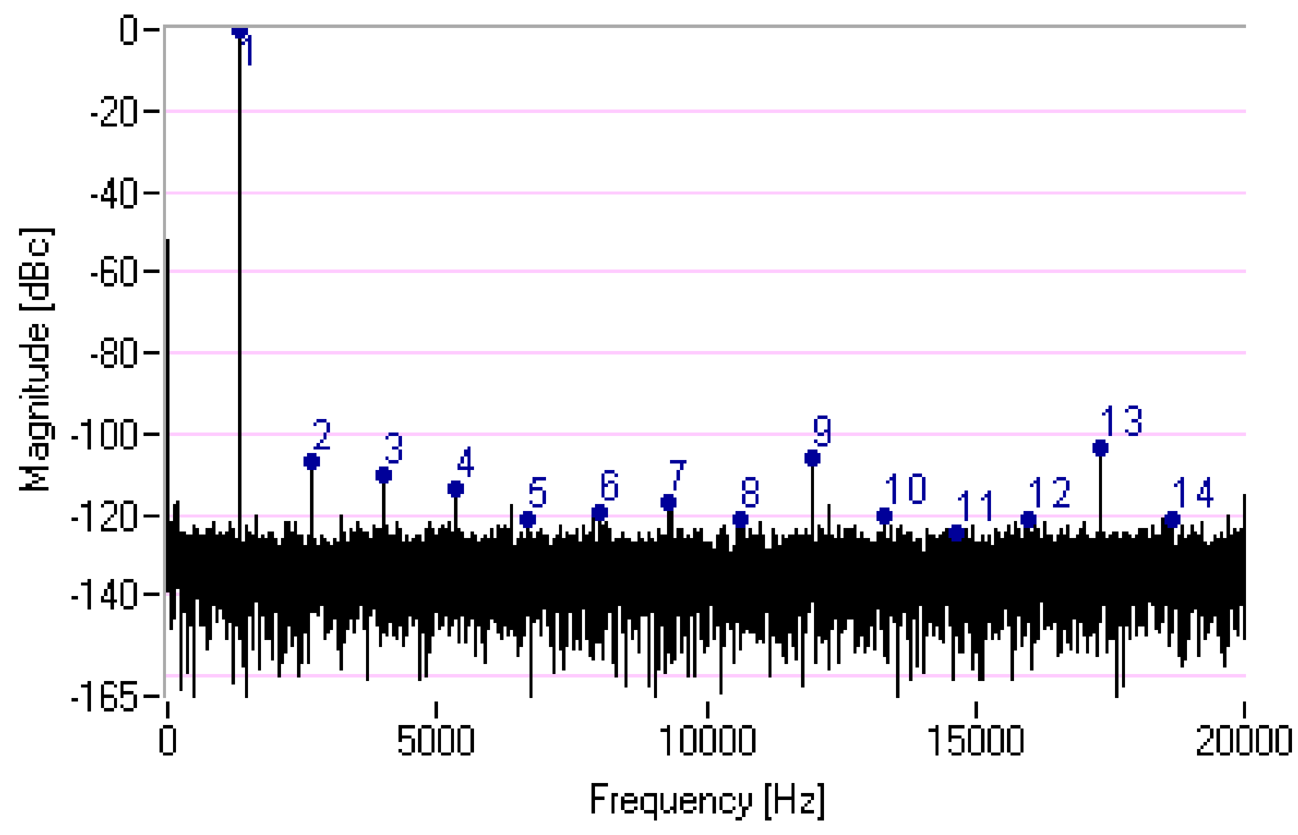
  $$m \times Tx = Mc \times \tau s$$

  where m and M are integer and co-prime

  (coherent and effective sampling)

  discontinuities at the boundaries of the observation interval cause a truncation error
- In the frequency domain, a spread of the spectral components over a broad range of frequencies arises (leakage effect). This effect is minimized through windowing.

16-bit ADC, non-coherent sampling, 4-term Blackman-Harris window, M = 128 kSa, *ENOB* = 14.01, *SINAD* = 86.08 dB, *SNR* = 86.28 dB



Histogram test

This test detects:

- differential non linearity;
- missing codes;
- gain and offset errors.

PROCEDURE:

- statistically significant number of samples is taken of the input sine wave as a record;
- the number of occurrences of each code is recorded on a histogram plot; for an ideal ADC the shape of plot would be the probability density function of a sine wave (input and sample frequencies must be relatively independent)

Histogram test

- the differential nonlinearity *DNL(i)* is a measure of how each code bin varies in size with respect to ideal;
- *DNL(i)* is computed using formula:

$$DNL(i) = \frac{actual\_P(i\ \mathrm{th} code)}{ideal\_P(i\ \mathrm{th} code)} - 1$$

where

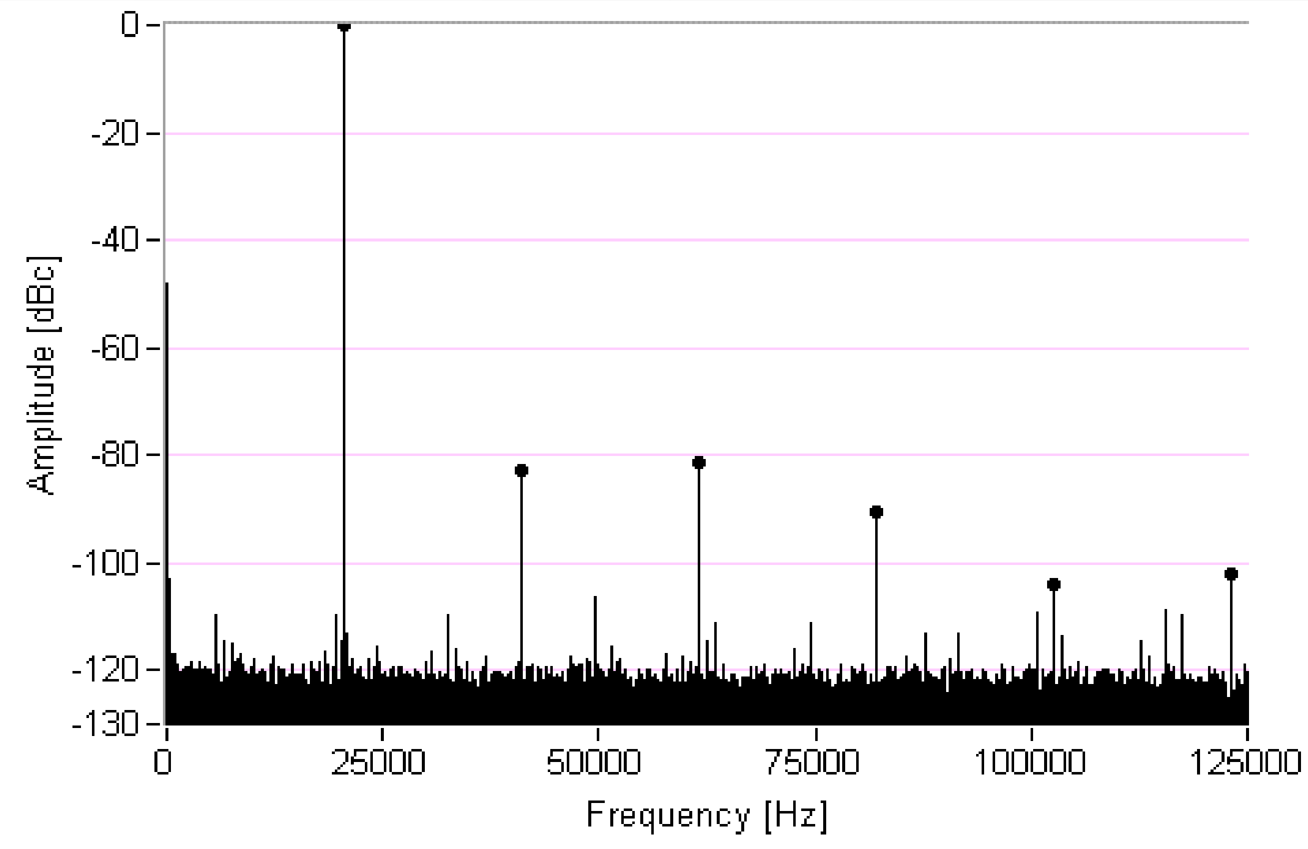
actual_P(i th code) - measured probability of occurrence for code bin i

ideal_P(i th code) - ideal probability of occurrence for code bin i

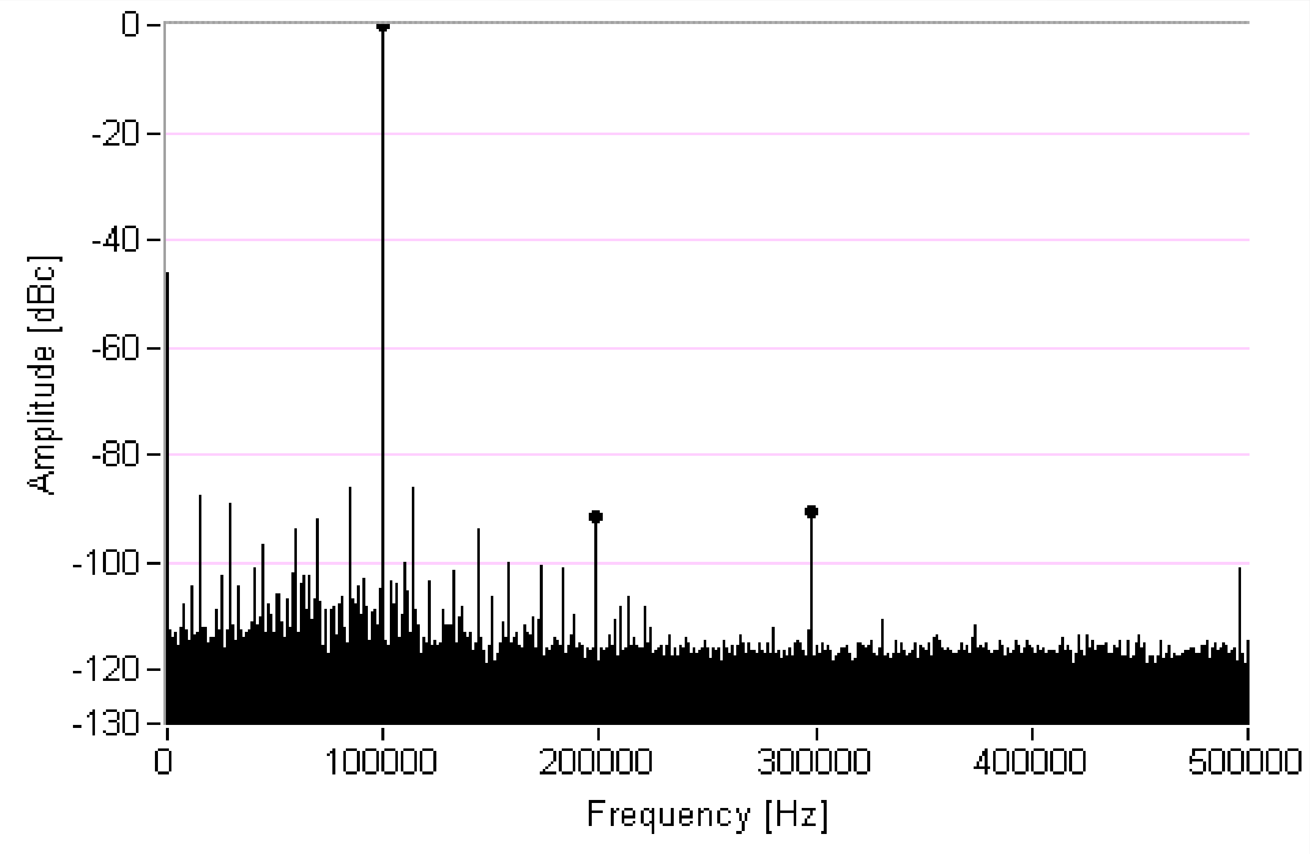
Testing low-distortion sine wave generators

- spectral quality of test signals should be evaluated using reliable measuring methods;
- ultra-low distortion sine wave generators can not be measured directly - measurement is influenced by non-linearities of measuring instrument (spectrum or distortion analyzer);
- in this case a non-distorting notch filters have to be used.

Spectrum of a sine wave signal 2 V_{PP} / 20.52 kHz from the Hewlett Packard HP33120A arbitrary waveform generator



Spectrum of a sine wave signal 2 V_{PP} / 99.2 kHz from the DS 360 generator - measured using the HP E1430A digitizer



Conclusions

Design

- Electronics for measurement systems have to take care about quality (i.e. metrology), not only effectiveness
- Influence parameters play a crucial role: be careful to noise-related issues
- A new generation of digital instruments is arising, exploiting key issues such as oversampling, digital integration, and noise modulation

Conclusions

Characterization

- Metrological specifications hide a conceptual world: take care about words
- New dynamic testing techniques for digital instrumentation are not sufficiently used, even though they are standardized