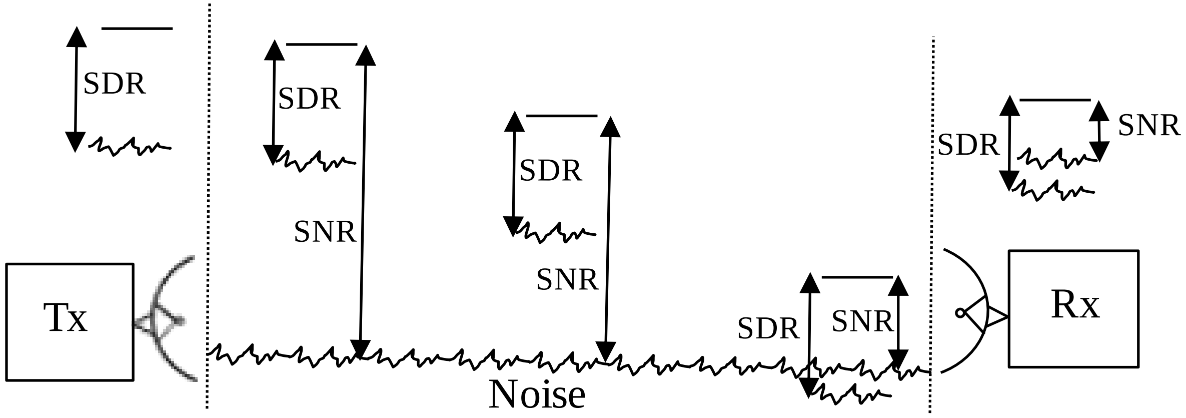
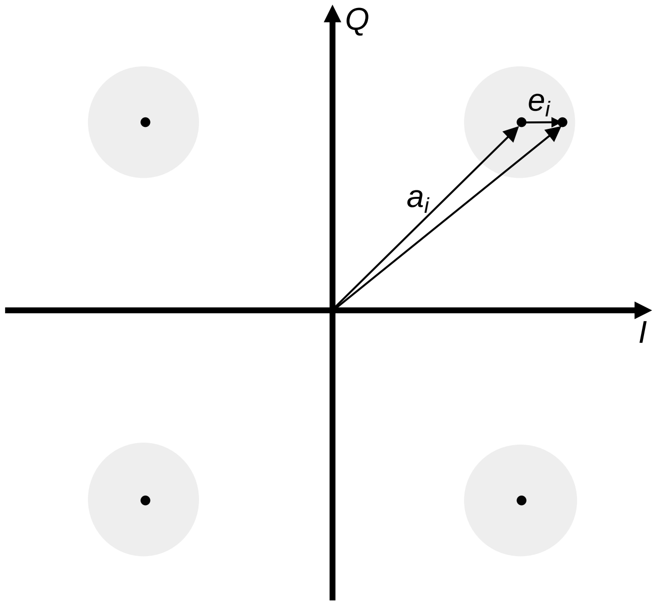
RF Systems Course: RF Systems Issues II

EVM in Transmitters

- EVM is another way to measure signal to noise and distortion ratio.
- EVM used more often in transmitters compared to receivers as a measure of modulation accuracy instead of SNR.
- When signal is generated in BBSP it is nearly perfect.
- Once signal passes through Tx, due to imperfections waveform will contain original information plus some distortion -> has finite signal-to-distortion level (SDR).
- If SDR is not larger than required SNR at Rx input then information transmission will not be successful even over zero distance.
- good rule of thumb SDR should be at least 6 dB higher than required SNR.
- As signal is transmitted over distance it is exposed to a noisy channel.
- signal power drops and while SDR stays constant SNR gets reduced.
- At max transmit distance it is expected that SNR may be less than SDR, but it should still be greater than min SNR required for detection.
- signal then passes through Rx where signal power is again amplified, but SNR and SDR are both reduced due to noise and non-idealities in Rx.



EVM in Transmitters



- In any real system some errors will always be introduced and vector that is transmitted will be different (an error) from what was intended.

$$EVM_i = \frac{|e_i|}{|a_i|}$$

- Normally EVM averaged over large number of data samples *N*:

$$EVM = \sqrt{\frac{1}{N}\sum_{i=1}^{i=N}\left(\frac{|e_i|}{|a_i|}\right)^2}$$

- are many sources of EVM in a transmitter and overall effect of various sources can be added together

$$EVM_{tot} = \sqrt{(EVM_1)^2 + (EVM_2)^2 ... + (EVM_M)^2}$$

Example:
A transmitter is using an OFDM, 16QAM modulation with 64 subcarriers.
The required BER for transmission is 10^{-3}.
What is the required EVM of the transmitter?
Assume a 6 dB safety margin.

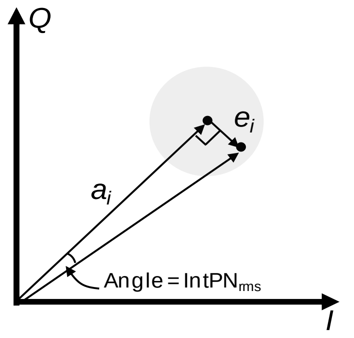
Solution:
For BER of 10^{-3} with 16QAM, SNR = 17 dB is required.
With 6 dB safety margin EVM of 23 dB or about 0.5% is required.
Remember EVM is just a different name for SNR in transmitter, because signal corruption in a transmitter comes more often from linearity and mixing than noise.

EVM Due to Phase Noise

- PN can be integrated to give rms phase variation in radians $IntPN_{rms}$.
- This phase variation will affect angle of reference vector.
- PN will generate an error vector orthogonal to reference vector.
- error vector will have a magnitude of

$$|e_{iPN}| = a_i \sin(IntPN_{rms}) \approx a_i IntPN_{rms}$$

$$EVM_{PN} = \frac{|a_i| \cdot IntPN_{rms}}{|a_i|} = IntPN_{rms}$$



Example:
What PN is required to achieve an EVM of 1%?
Assume BB signal BW is 500kHz and PN of synthesizer is flat across entire BW.

Solution:
An EVM of 1% or 0.01 means that integrated rms PN = 0.01 rads.
If PN is flat across entire band of 500 kHz that means that PN must be 200 prads/Hz or -97 dBc/Hz

EVM Due to IQ Mismatch

- Note that a phase mismatch between I/Q paths either created by LO or mismatch in baseband paths will behave exactly same way in regards to EVM as PN does.
- if there is a phase mismatch in LO of θ_{LO} then EVM due to IQ phase mismatch will be

$$EVM_{LO} = \theta_{LO}$$

- Another source of EVM is gain mismatch in baseband I/Q paths.
- If there is some random shift in amp from ideal amp in I / Q paths δ such that mag of I path = $I(1+\delta)$ and mag. of Q path = $Q(1-\delta)$ then error vector will be

$$e_i = \sqrt{(\delta I)^2 + (\delta Q)^2}$$

- reference vector will have a magnitude of

$$a_i = \sqrt{I^2 + Q^2}$$

- EVM will be simply

$$EVM_{IQ} = \delta$$

EVM Due to Carrier Feedthrough

- Carrier feedthrough can be due to finite isolation from LO to RF and from LO to IF and DC offset as well as other factors.
- Whatever the source, any energy from the carrier will distort the waveform and cause EVM.
- If we define carrier suppression C_s as ratio of desired RF power P_t to LO leakage P_{CFT} then EVM due to carrier feed through is

$$EVM_{CFT} = \sqrt{\frac{P_{CFT}}{P_t}} = \sqrt{C_s}$$

EVM Due to Linearity

- Linearity in Tx may can also cause EVM, depending on type of modulation.
- If system uses phase-only modulation, linearity is of much less concern.
- With amp sensitive modulations like QAM best to operate output with a power level much less that P_{1dB}.
- If Tx is reasonably linear and linearity can be modeled by a 3rd-order power series then instantaneous EVM is:

$$EVM_i = \frac{\frac{3}{4} k_3 v_{RFi}^{3}}{k_1 v_{RFi}} = \frac{0.285 \frac{v_{RFi}^{2}}{v_{1dB}}}{Gain}$$

- EVM_i must be computed for each constellation point and then divided by total number of constellation points

$$EVM_{nonlin} = \frac{1}{n} \sum_{i=1}^{n} \frac{0.285 \frac{v_{RFi}^{2}}{v_{1dB}}}{Gain}$$

EVM Due to Linearity Example

Determine EVM of 16QAM Tx after passing through the PA.
The constellation in one quadrant is: (1,1), (3,1), (1,3) and (3,3).
Choose P_{1dB} for PA.

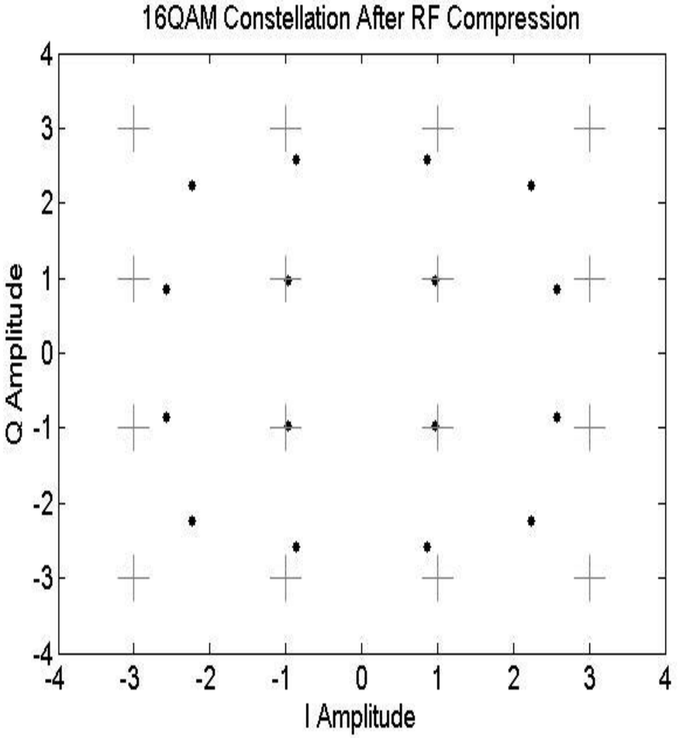
Solution:

With this modulation scheme need to first determine the signal amplitudes.

With random bit stream it can be expected that symbols at distance (1,1) or voltage amp =1.414 V transmitted 25% of the time, symbols at amp 3.162 (points (3,1) or (1,3)) transmitted 50% of the time, symbols at amp 4.243 transmitted 25% of the time.

As a starting point P_{1dB}=4.5 V is selected.
assume amp gain is one.

$$EVM_{nonlin} = \frac{1}{4}\left(0.285 \cdot \frac{1.414^2}{4.5} + 2 \cdot 0.285 \frac{3.162^2}{4.5} + 0.285 \frac{4.243^2}{4.5}\right) = 14.1\%$$



simulated with no band limiting filters and after ideal demodulation output constellation for 1000 symbols transmitted.
outer points are far more compressed than the inner ones.
actual simulated EVM was 16.4% which is close to theoretical value.

EVM Using OFDM Modulation

- Systems that use OFDM have a different set of linearity requirements than systems that use simpler modulation schemes.
- If we assume that Tx is operating well below P_{1dB}, dominant contribution to EVM degradation due to linearity of Tx will be triple-order beats of all subcarriers dumping energy on top of sub carrier of interest.
- power in triple-beat tone is

$$TB = OIP_3 - 3(OIP_3 - P_s) + 6 = 3P_s - 2OIP_3 + 6$$

- P_s is power of tones generating triple beat (in this case power in one sub carrier)
- EVM can be computed as ratio of power of the subcarrier to the power of all triple beats dumping power into that channel.

$$TB - P_s = 2P_s - 2OIP_3 + 6 + 10\log\left(\frac{3}{8}N^2\right)$$

- EVM due to transmitter nonlinearity is

$$EVM_{lin} = 10^{\frac{\left[2P_s - 2OIP_3 + 6 + 10\log\left(\frac{3}{8}N^2\right)\right]}{20}}$$

- where N is the number of sub carriers.
- total transmitted power is related to P_S

$$P_t = P_s + 10\log N$$

$$20\log(EVM_{lin}) = 2P_t - 10\log N^2 - 2OIP_3 + 6 + 10\log\left(\frac{3}{8}N^2\right)$$

- if output power is close to P_{1dB} this formula will become less accurate.

EVM Using OFDM Modulation

Example: Determining Transmitter Linearity
A transmitter is required to transmit at a power level of 25 dBm and is using a 64QAM OFDM modulation with 64 subcarriers. What is required linearity of Tx to achieve EVM = 2.5%?
Solution:

$$-32 = \left[2(25) - 2OIP_3 + 6 + 10\log\left(\frac{3}{8}\right) \right]$$

$$OIP_3 = 41.9dBm$$

output P_{1dB} would be about 32.3 dBm which gives 7.3 dB of backoff.

EVM and Noise and Filters

- If Tx is backed off and transmitting at low power level then noise in Tx circuits can become an important contribution to EVM.
- If total noise power density coming from Tx is N_0 then EVM contribution from noise is

$$EVM_{noise} = \sqrt{\frac{N_o \cdot BW}{P_t}}$$

- Filters in signal path can also be a source of EVM if their BW is close to signal BW
- usually only necessary to consider BB filters or perhaps an IF filter if it is only a couple of channels wide.
- Since it is impossible to build an ideal filter, any real filter will have a non-zero impulse response during adjacent bit periods.
- power will be spread from one symbol period to another.
- ISI can also lead to EVM.
- Knowing a filter's impulse response $h_f(t)$, its contribution to EVM can be found as

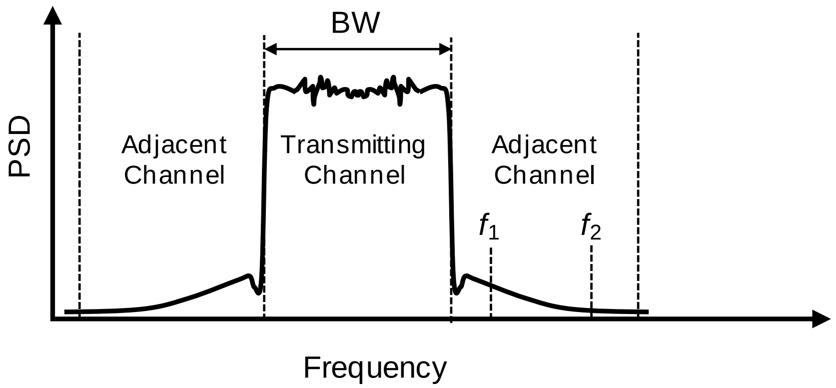
$$EVM_{filter} = \sqrt{\frac{\sum_{k=-\infty, k\neq 0}^{k=\infty} \left| h_f\left(t_o + kT_s\right)\right|^2}{\left| h_f\left(t_o\right)\right|^2}}$$

Adjacent Channel Power

- In any transmitter there can be power transmitted beyond the BW of the channel.
- not desired as it has potential to interfere with other communications.
- there are stringent requirements on how much power leakage can take place in any transmitter.
- adjacent channel power ratio (ACPR):

$$ACPR = \frac{\int_{f_1}^{f_2} PSD(f)df}{\int_{f_{ch}-\frac{BW}{2}}^{f_{ch}+\frac{BW}{2}} PSD(f)df}$$

- BW over which adjacent power is measured is not always equal to channel BW and often some faction of complete channel



Adjacent Channel Power

- possible to have a low ACPR with a single carrier modulation, probably be due to transmitter being pushed beyond P_{1dB}.
- Simple theory would not be accurate -> simulation would be needed to determine the actual value of ACPR.
- With multi-carrier signals ACPR becomes much more of an issue even if Tx is linear compared to signal power levels.
- To start consider case where 2 carriers are present in channel.
- They are spaced so that IM3 products will fall in channel above and below transmitting channel.
- IM3 tones will result in finite ACPR value.
- Assuming that power level of each tone is P_t and that OIP_3 of Tx is known, power of IM3 tones is:

$$IM3 = OIP_3 - 3(OIP_3 - P_t)$$

- Since each tone represents half total transmit power P_{TR} :

$$ACPR = OIP_3 - 3(OIP_3 - P_t) - 2P_t$$
$$= P_t - 2OIP_3$$

- since total transmit power will be $2P_t$, ACPR is

$$ACPR = \frac{P_{TR}}{2} - 2OIP_3$$

- Even though this result is derived for 2 tones in general with N tones this result will still hold.
- Each tone gets $1/N$ of total power and each will generate a smaller IM3 tone, but total will add up to be almost the same in the end.
- result will be more like noise and will be distributed across adjacent channel.
- If only a fraction of the adjacent channel is used then ACPR is

$$ACPR \approx \frac{P_{TR}}{2} - 2OIP_3 + 10\log\left(\frac{f_2 - f_1}{BW}\right)$$

Adjacent Channel Power Examples

Example: ACPR Requirements

A radio is required to receive a signal with SNR = 10dB.

If min detectable signal is -80dBm and max adjacent power is -40dBm what is max allowable ACPR for radio to operate properly?

Solution:

If min detectible signal is -80dBm and requires a SNR of 10dB, means that max interference power is -90dBm.

If adjacent signal power is -40dBm therefore ACPR of that signal must be -40dBm – (-90dBm) = 50dBc.

Adjacent Channel Power Examples

Example: Determining ACPR

Consider a radio using a 32 carrier OFDM signal where at base band each subcarrier uses a center freq of 1, 2, 3, …32 Hz.

The subcarriers are modulated with simple BPSK and are then up converted to a 200 Hz transmit freq.

If OIP3 of Tx is 27.8 dBV and power density in channel is -5 dBV/Hz what is ACPR if adjacent power is measured in a 10 Hz BW?

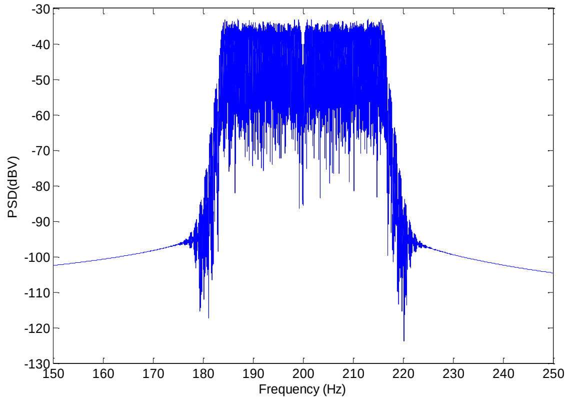
Solution:

With 32 subcarriers and a max carrier freq of 32 Hz can estimate pass band BW = 34 Hz assuming each subcarrier has a BW of 1 Hz and allowing for some guard bands. Total transmit power will be

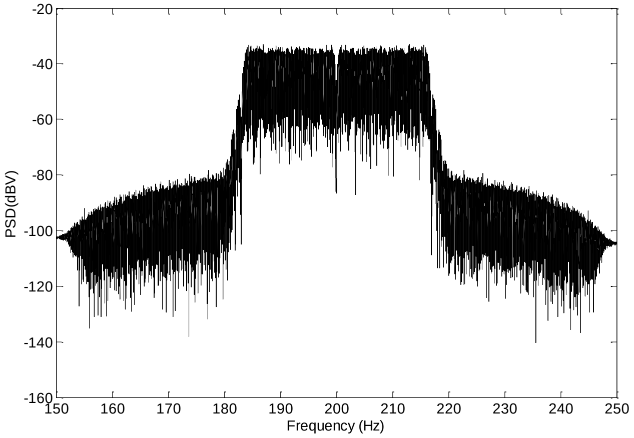
$$P_{TR} = -5dBV/Hz + 10\log(34) = 10.3dBV$$

$$ACPR \approx \frac{10.3}{2} - 2(27.8) + 10\log\left(\frac{10}{34}\right) = -55.8dBC$$

Now this waveform must be created and simulated. The baseband waveform is ideally band limited and modulated to desired carrier freq.



- waveform is passed through a nonlinear circuit with an OIP3 of 27.8 dBV.
- resolution on these two plots is 0.001 Hz.
- adjacent channel between 165 and 175 Hz power level is about -85 dBV/0.001Hz = -55 dBV/Hz or -45dBV over 10 Hz of BW.
- channel power is 10 dBV so ACPR is -55 dBc which is very close to the predicted value

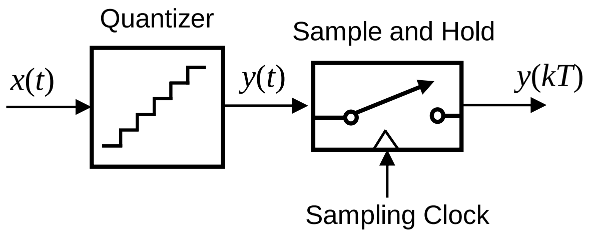


ADCs and DACs

- When specing ADC, need to specify number of bits, max input voltage swing, required jitter on clock driving the ADC, sampling freq.
- AGC step size will affect ADC performance requirements.
- larger the step size the larger the input signal amp. variation.
- If AGC does not cover whole dynamic range then ADC will have to handle signal level variation.
- SNR of ADC must be at least as high as min required SNR of signal.
- If signal has a non-unity peak-to-average ratio then ADC must be designed to handle max signal level without introducing distortion.
- There may also be a DC offset in baseband of the radio.
- this will be an added signal that must be accommodated without distorting or saturating ADC.
- quality of baseband filters will also play a role in determining requirements of the ADC.
- If out of band interferers are not completely filtered by the BB filters then ADC will have to handle these signals as well without saturating or aliasing them into receive band.
- DACs are similar to ADCs, however they deal with known signals while ADCs have to be able to deal with less ideal inputs.
- harmonics of sampling clock must be considered as these may cause adjacent channel power.
- quality of transmit baseband filters can be traded off against the amount of aliasing to higher freqs that can be allowed in DAC.
- dynamic range of DAC must be large enough to handle peak-to-average ratio of signal as well as still have a high enough SNR so that quantization noise of DAC does not become limiting factor in determining EVM of the transmitter.

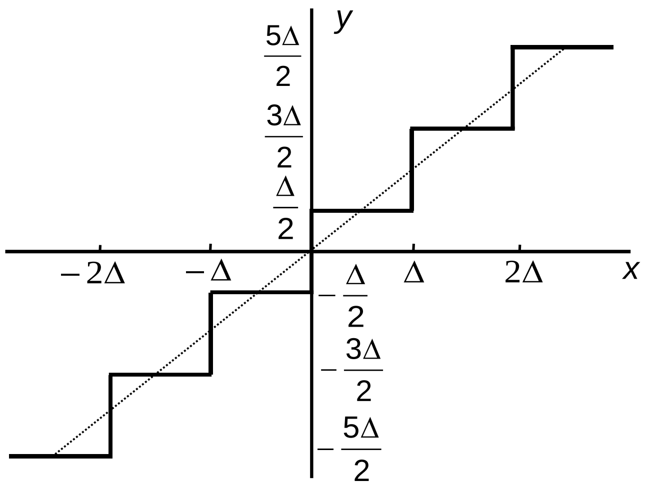
ADC and DAC Basics

- A basic ADC consists of a quantizer and a sample and hold circuit
- While many ADCs may be much more complex, this will serve to illustrate many key basic points



- quantizer converts continuous analog signal x to discrete signal according to quantization rule, where x is analog input and y is quantized output.
- y not yet sampled therefore still a continuous signal
- y is function of input, but has discrete levels at equally spaced intervals.
- unless input happens to be an integer multiple of quantizer resolution (step size), there will be error in representing input.
- This error e will be bounded over one quantizer level by a value of

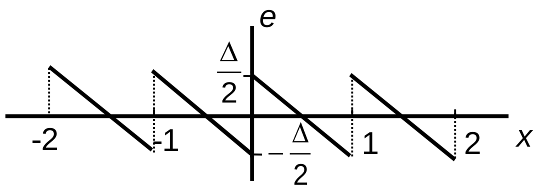
$$-\frac{\Delta}{2} \leq e \leq \frac{\Delta}{2}$$

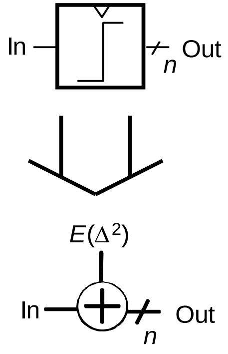


- quantized signal y can be represented by linear function with error e

$$y = \Delta \cdot x + e$$

- error is a straight line with a slope of –Δ.
- If input is “random” then instantaneous error will also be random.
- error is uncorrelated sample to sample -> can be treated as “noise”.
- Quantization and resultant quantization noise can be modeled as a linear circuit including an additive quantization error source





ADC and DAC Basics

- quantization noise for a random signal can be treated as additive white noise having a value anywhere in the range from $-\Delta/2$ to $\Delta/2$.
- quantization noise has a uniform probability density

$$p(e) = \begin{cases} \dfrac{1}{\Delta} & \text{if} \quad -\dfrac{\Delta}{2} \le e \le \dfrac{\Delta}{2} \\ 0 & \text{otherwise} \end{cases}$$

- mean square rms error voltage e_{rms} can be found by integrating square of error voltage and dividing by quantization step size

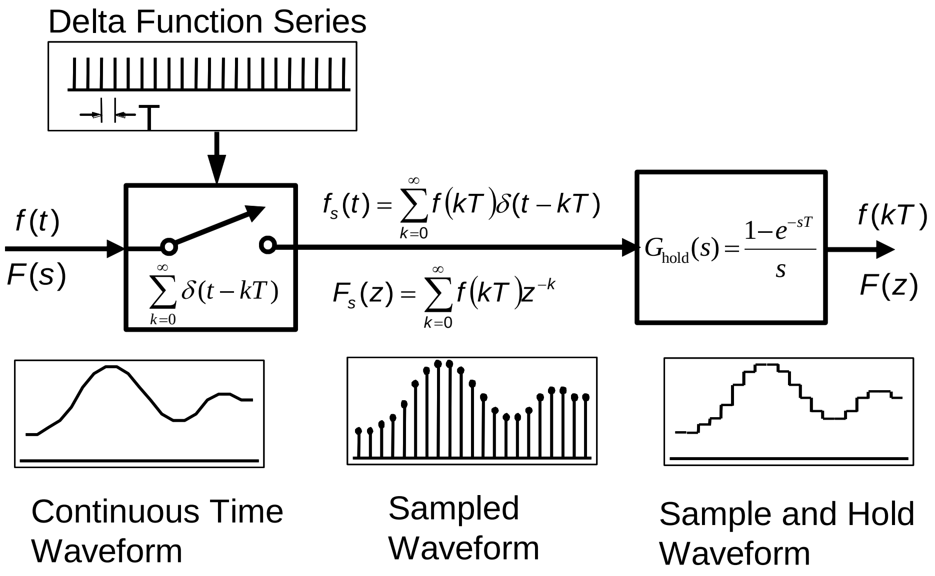
$$e_{\text{rms}}^2 = \int_{-\infty}^{+\infty} p(e) e^2 de = \frac{1}{\Delta} \int_{-\Delta/2}^{+\Delta/2} e^2 de = \frac{\Delta^2}{12}$$

- Sampling is act of measuring a system's output at periodic points in time.
- only system's outputs at instances when sampling occurs are of interest, anything system does in between sampling instances is lost.
- Sampling is done once per clock cycle
- sampled value is held until next sampling instance.
- to have good digital representation of waveform -> must be sampled at least twice as fast as highest freq of interest $f_{\max}$

$$f_s \ge 2 f_{\max}$$

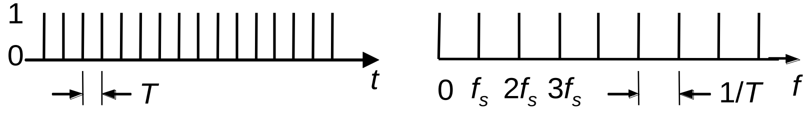
- This is known as Nyquist sampling rate and although this is theoretical min in practice often choose $f_s > 2.2\, f_{\max}$
- sampler also causes delay -> it can be thought of as a delay block.
- Sampling can be described mathematically as multiplying waveform by a series of delta functions, sometimes called an impulse train
- waveform $f(t)$ sampled every T time units can be written as

$$f_s(t) = f(t) \sum_{k=0}^{\infty} \delta(t - kT) = \sum_{k=0}^{\infty} f(kT) \delta(t - kT)$$

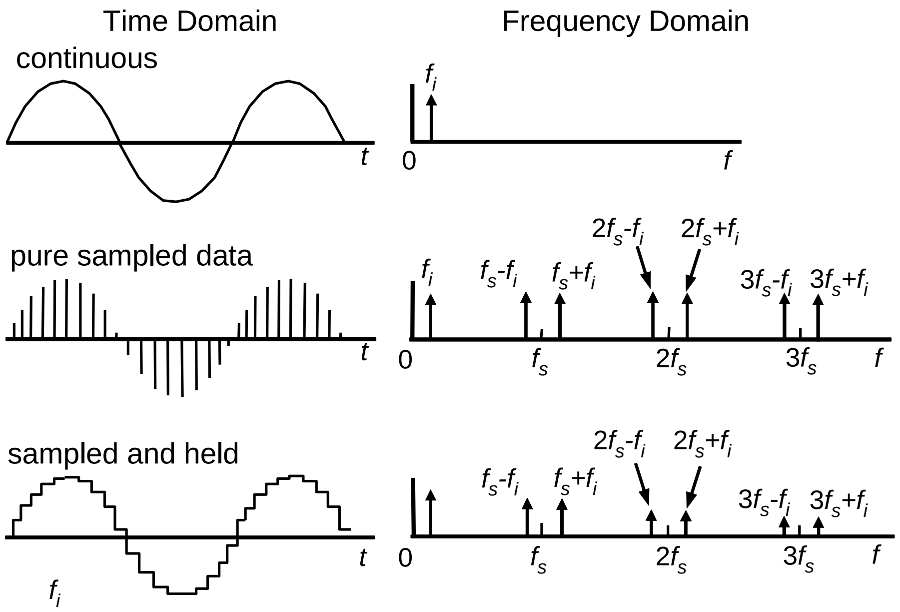


ADC and DAC Basics

- Fourier transform of a series of equally spaced delta functions in time domain (to represent sampling) is a set of delta functions in freq domain.
- These are now convolved with spectrum of the input waveform



- effect of convolving input waveform with a delta function is to re-center waveform at freq of the delta function.
- These impulses can be seen to occur at dc, at sampling freq f_s, at $2f_s$, $3f_s$, etc.
- freq spectrum of continuous waveform repeats around f_s and around all multiples of sampling freq.
- This mixing property of sampled systems is called replication.
- A consequence of replication is that signals fed into system at freqs close to the sampling freq, or a multiple of sampling freq will produce an output signal in baseband (somewhere between dc and $f_s/2$).
- Such signals are indistinguishable from signals fed in close to dc.
- This typically unwanted output signal is called an aliased component.



- When sampling is accompanied by a hold function, freq response is first convolved with delta function to represent sampling function.
- result is multiplied by the hold function.
- The hold function can be shown to have a transfer function of

$$G_{\text{hold}}(j\omega)=\frac{1-e^{-j\omega T}}{j\omega}\frac{e^{j\omega T/2}}{e^{j\omega T/2}}=Te^{-j\omega T/2}\frac{e^{j\omega T/2}-e^{-j\omega T/2}}{2j\omega T/2}=Te^{-j\omega T/2}\frac{\sin(\omega T/2)}{(\omega T/2)}$$

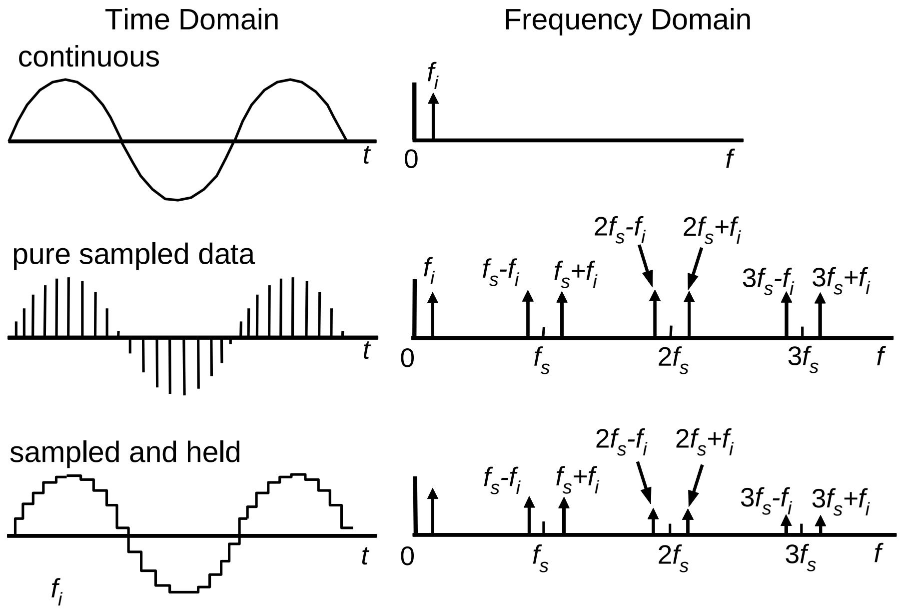
ADC and DAC Basics

- frequency spectrum of a sampled system repeats once every sampling freq, and therefore spectrum of quantization noise in a sampled system will be centered around dc and spread out to half of sampling freq $f_s/2$ and there will be a copy of noise spectrum from $f_s/2$ to $3f_s/2$ and so on.
- Considering that all the noise power lies in range of positive freq band, i.e., $0 \le f \le \infty$, quantization noise power thus folds into the band from dc to $f_s/2$.
- Assuming white noise, power spectral density $E^2(f)$ of the quantization noise is

$$E^2(f) = \frac{e^2_{\text{rms}}}{f_s/2} = 2Te^2_{\text{rms}}$$

- where the sample period $T = 1/f_s$.
- For band limited signal $0 \le f < f_0$ with BW = f_0, quantization noise power that falls into the signal band can be found as

$$n_0^2 = \int_0^{f_0} E^2(f)df = 2f_oTe^2_{\text{rms}} = \frac{\Delta^2 f_o}{6 \cdot f_s} = \frac{\Delta^2}{12 \cdot \text{OSR}}$$



- where *oversampling rate* (OSR) is defined as ratio of sampling freq f_s to Nyquist freq $2f_o$

$$\text{OSR} = \frac{f_S}{2f_0}$$

ADC and DAC Basics

- In an N-bit sampled system, if quantizer has 2^N quantization levels equally spaced by Δ, then max peak-to-peak amp is

$$v_{\max} = \left(2^N - 1\right) \cdot \Delta$$

  - This equation is an approximation, but if signal is sinusoidal and if N is large approximation will be a small one.
  - The associated signal power is

$$P = \frac{1}{8}\left(2^N - 1\right)^2 \cdot \Delta^2$$

  - dynamic range (DR) (the difference between highest input level and the noise floor)

$$\mathrm{DR} = 10\log\left(\frac{\frac{1}{8}\left(2^N - 1\right)^2 \Delta^2}{n_0^2}\right) \approx 10\log\left(\frac{3 \cdot 2^{2N}\,\mathrm{OSR}}{2}\right)$$

$$\mathrm{DR} \approx 6.02 \cdot N + 3 \cdot \log_2(\mathrm{OSR}) + 1.76$$

- DR improves by 6 dB for every bit added to the quantizer.
- For same amount of total quantization noise power, every doubling of sampling freq reduces in-band quantization noise by 3 dB.
- ADCs have been talked about here, but this result would hold for a DAC as well.

ADC and DAC Basics

- other main source of noise in an ADC or DAC is timing jitter of reference clock.
- rms timing jitter t_{jitter} is related to integrated PN of reference clock by

$$t_{jitter} = \frac{IntPN_{rms}}{2\pi} \cdot T_{clk}$$

- Now if input to ADC or output of a DAC is assumed to be a sine wave at a freq f_{in} amount of noise caused by reference clock jitter can be estimated.
- If input waveform is

$$v_{in}(t) = A\sin(2\pi \cdot f_{in}t)$$

$$\frac{dv_{in}(t)}{dt} = 2\pi \cdot f_{in}A\cos(2\pi \cdot f_{in}t)$$

- This slope will be highest at waveform zero crossings and zero at waveform peak and if there is an error in time waveform is sampled t_{jitter} then there will be an rms error in sampled voltage

$$v_{in\_error\_rms} = 2\pi \cdot \frac{f_{in}A}{\sqrt{2}} \cdot t_{jitter}$$

- Note that root 2 comes from making this an rms value.
- SNR will be given by

$$SNR_{jitter} = -20\log(2\pi \cdot f_{in} \cdot t_{jitter})$$

- now possible to estimate # of bits and clock jitter required to assure that DR of ADC or DAC does not degrade overall system SNR by more than required amount.

ADC and DAC Basics

- Example: Specifying an ADC
- An ADC is required to have a SNR of 30 dB and the signal bandwidth is 20 MHz.
- Ignoring any out of band interferers give as much detail as possible about required performance of ADC.

- Solution:

- First assume to use a Nyquist rate ADC for this design and therefore ADC will be clocked at 40 MHz and OSR=1.
- To give some margin in design we will calculate # of bits required for a SNR of 33 dB.

$$N = \frac{SNR - 1.76}{6.02} = 5.2$$

a 6-bit ADC will be adequate.
Next calculate requirement on reference clock jitter.
assuming 33 dB to give some margin

$$t_{jitter} = \frac{\log^{-1}\left(\frac{-SNR_{jitter}}{20}\right)}{2\pi \cdot f_{in}} = 0.18\text{ns}$$