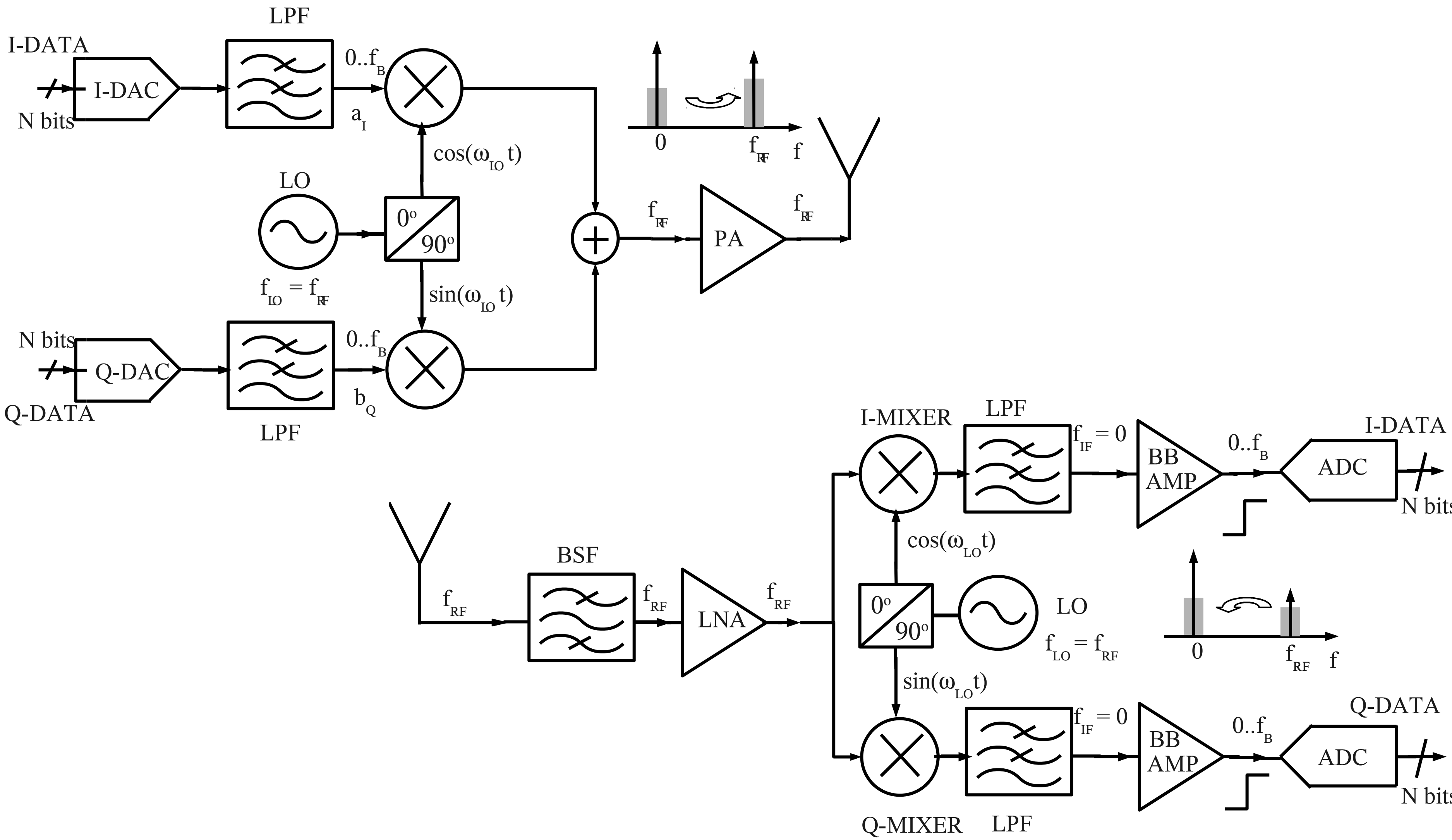
Chapter 2: HF and High Data-Rate Systems

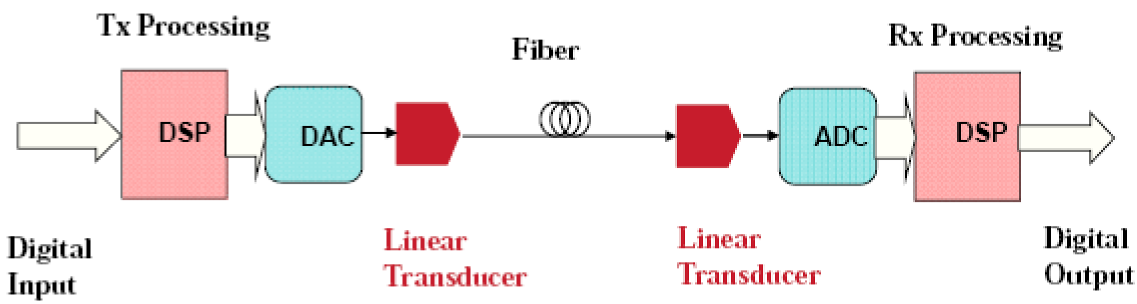
Outline

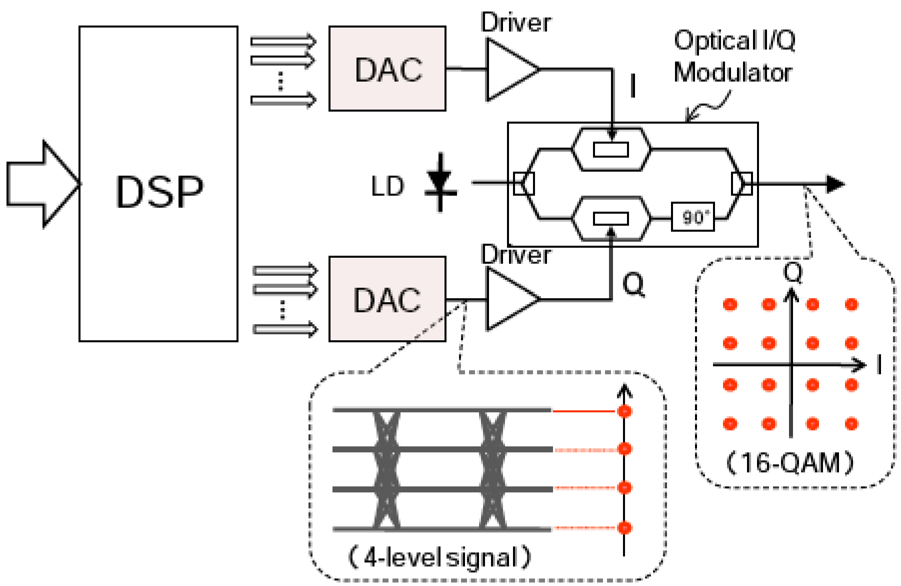
- Wireless and wireline systems
- Modulation schemes
- Transceiver architectures
- Transceiver specification
- Link budget
- Phased Arrays
- Radar and radiometers

Radio Transceiver Example



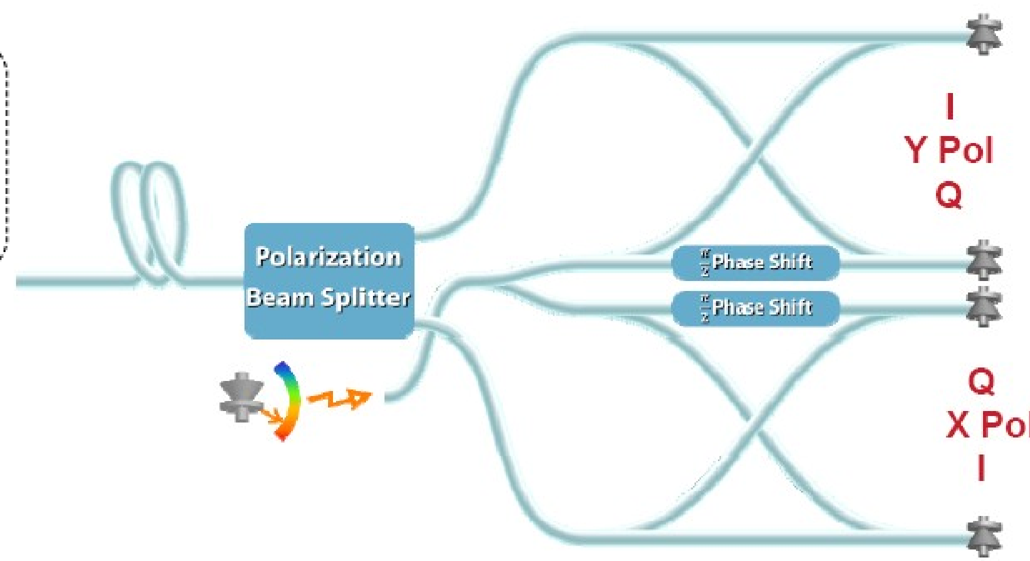
16-QAM Fiberoptic Transceiver Example



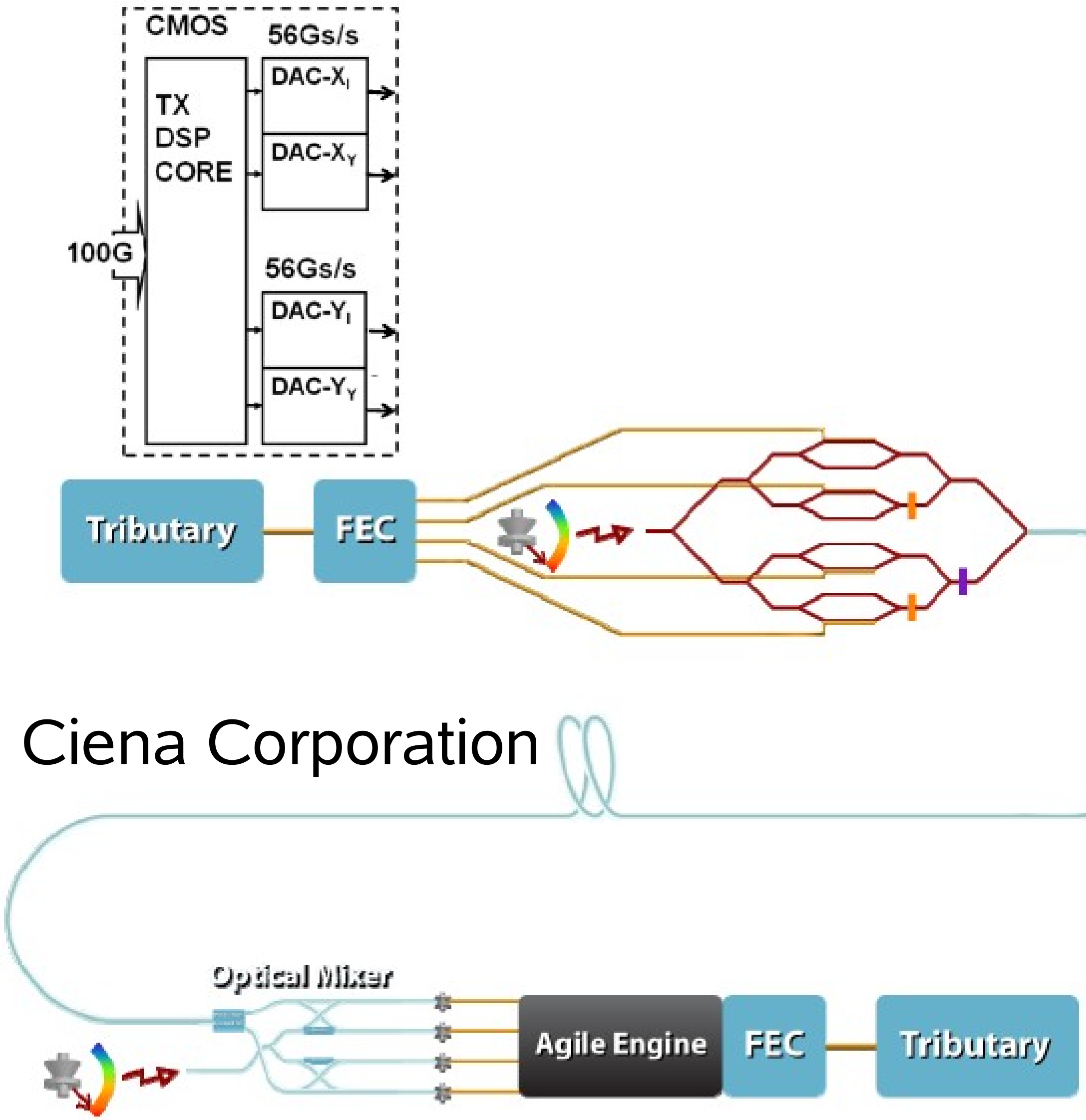


Coherent transmitter with linear upconversion

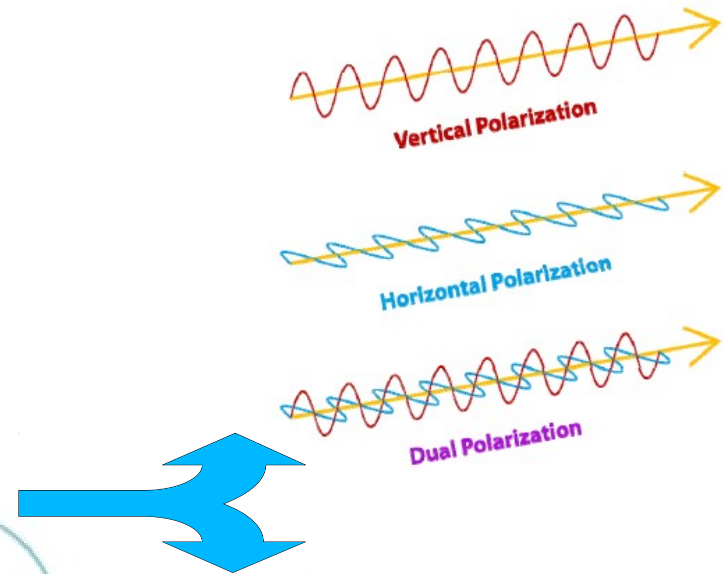
Coherent receiver

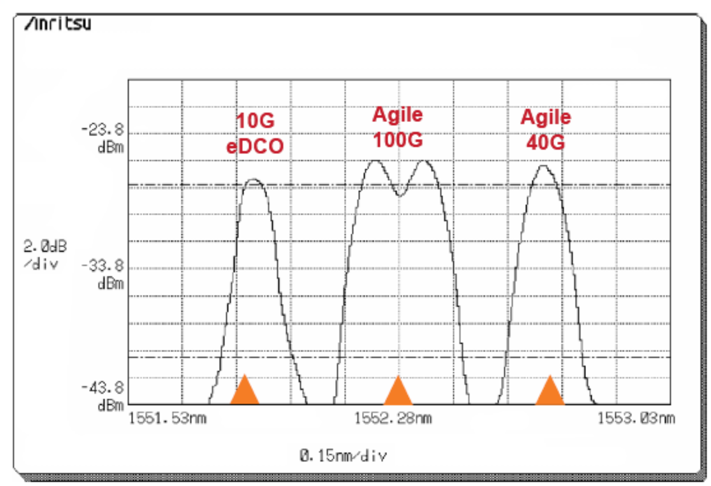


110-Gb/s DP-QPSK Fiberoptic Transceiver

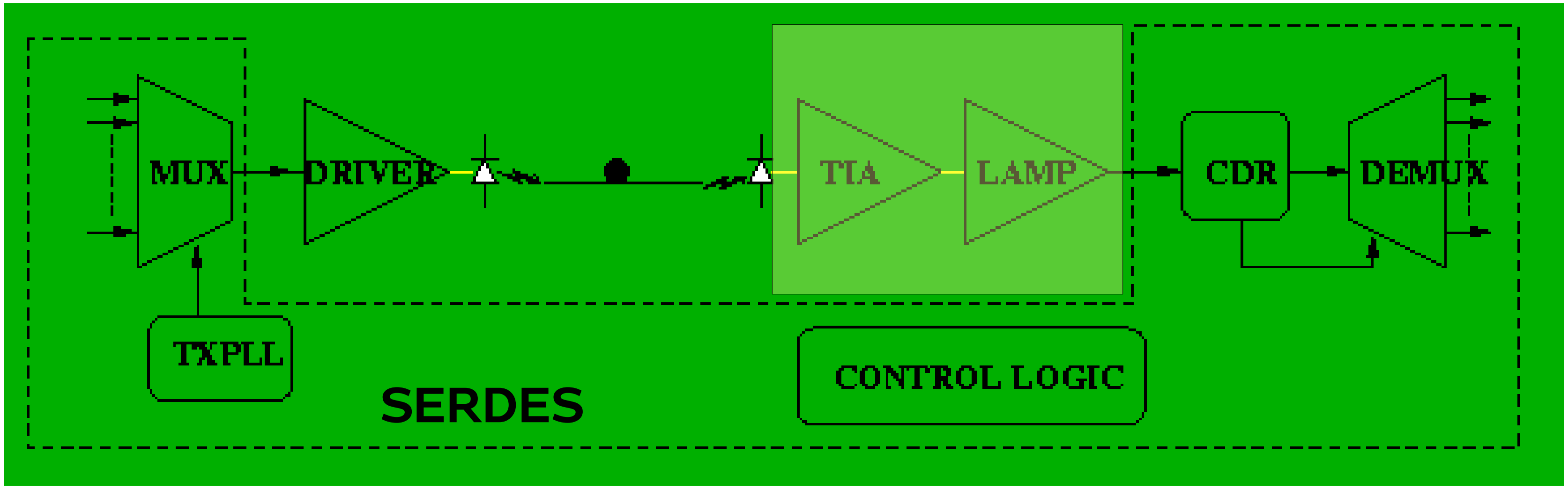


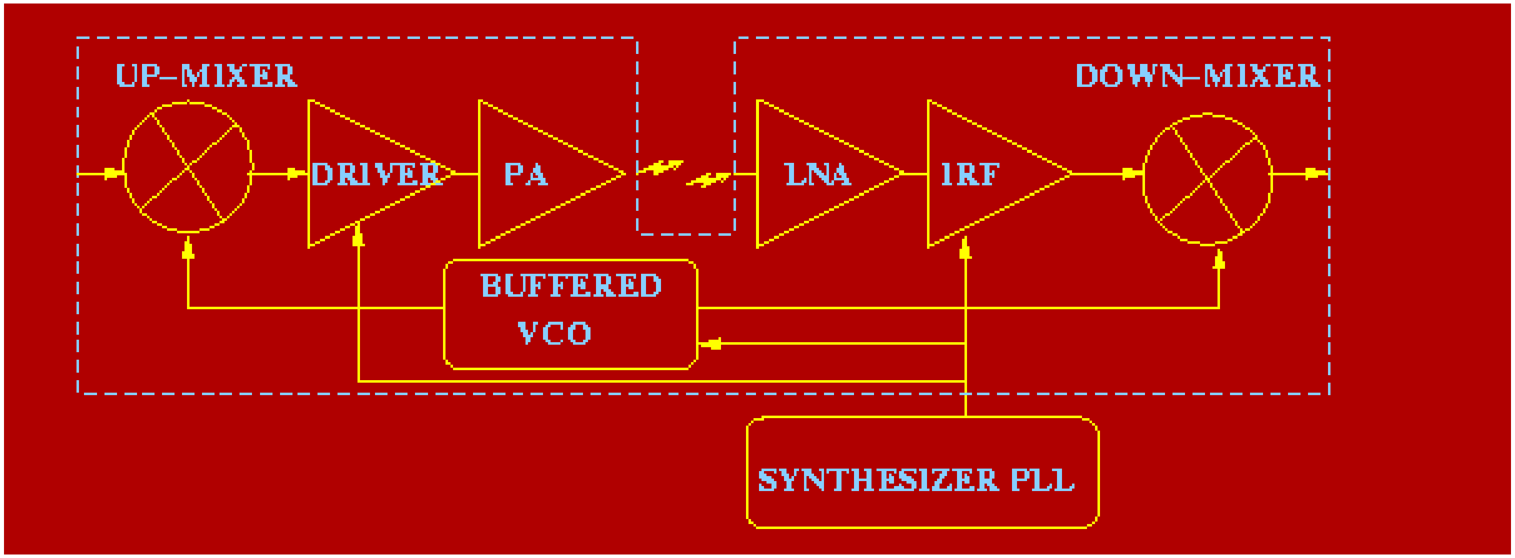
Ciena Corporation





Electronic circuits:
wireless vs. fiberoptic communications systems





Radio vs. Broadband Fiber Circuits

- Input/Output vs. Output return loss (S params)
- Receiver Sensitivity @ BER and SNR (Q)
- Linear LNA vs. linear TIA
- NF vs. equivalent input noise current
- (Non)Linear PA vs. Non-Linear TX drivers
- P_O (average) vs. V_O (p-p)
- Frequency conversion vs. MUX/DeMUX

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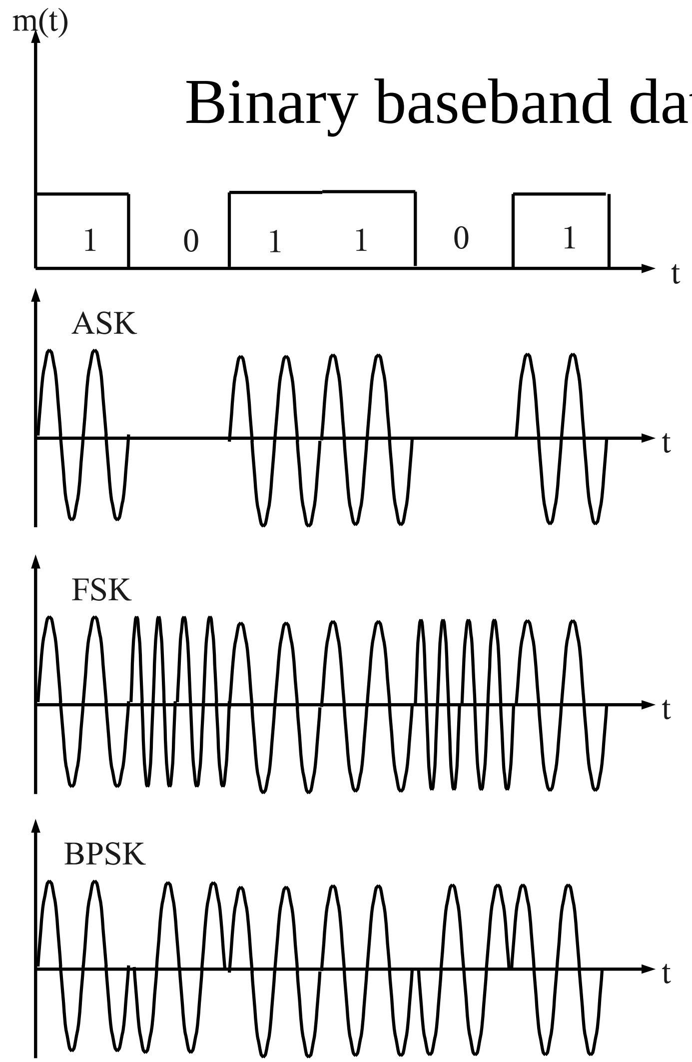
Modulation Schemes

- Information modulates carrier
  - Amplitude
  - Frequency
  - Phase

$$s(t)=A\cos(\omega t+\phi)$$

- Modulation schemes
  - Analog
  - Digital
- Digital preferred because of
  - Robustness to noise, interference, fading
  - Ease of error correction and encryption

Digital Modulation Schemes

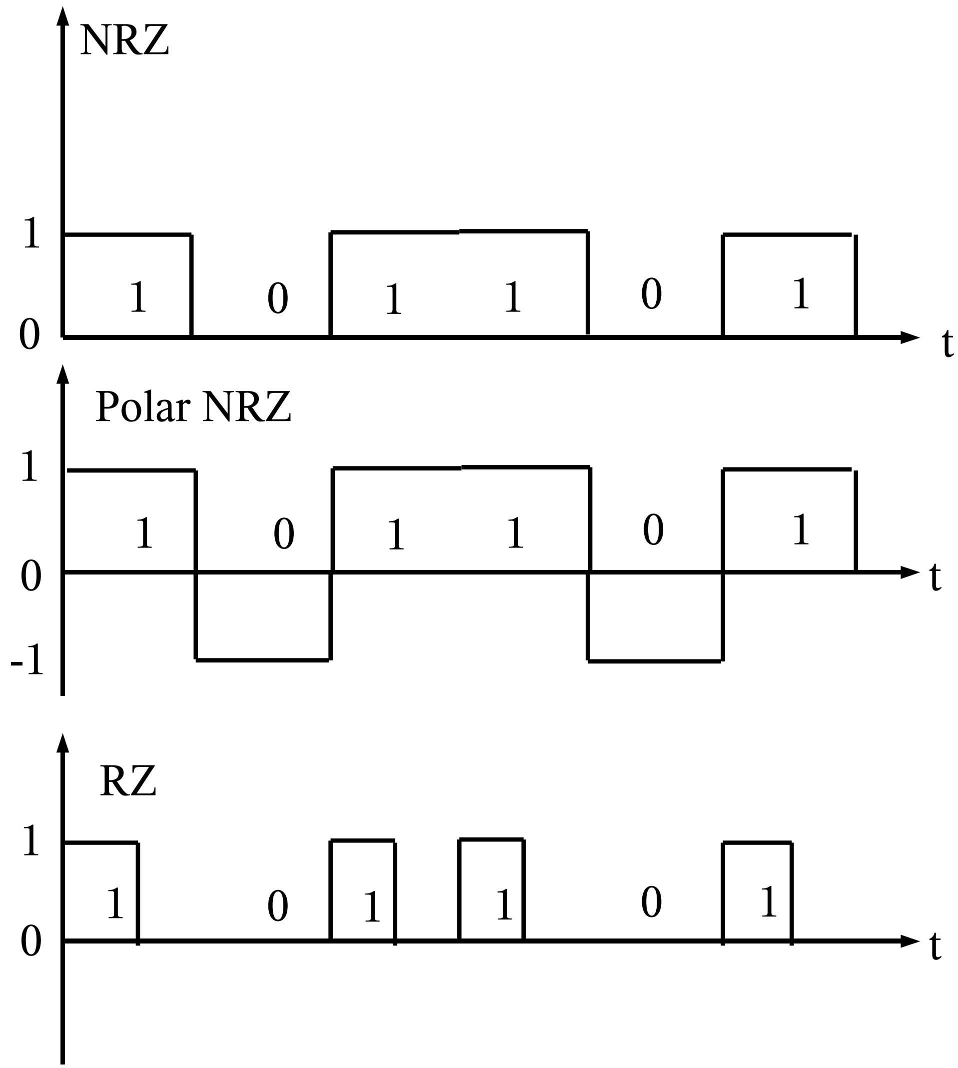


Amplitude Shift Keying or
On-Off Keying =OOK

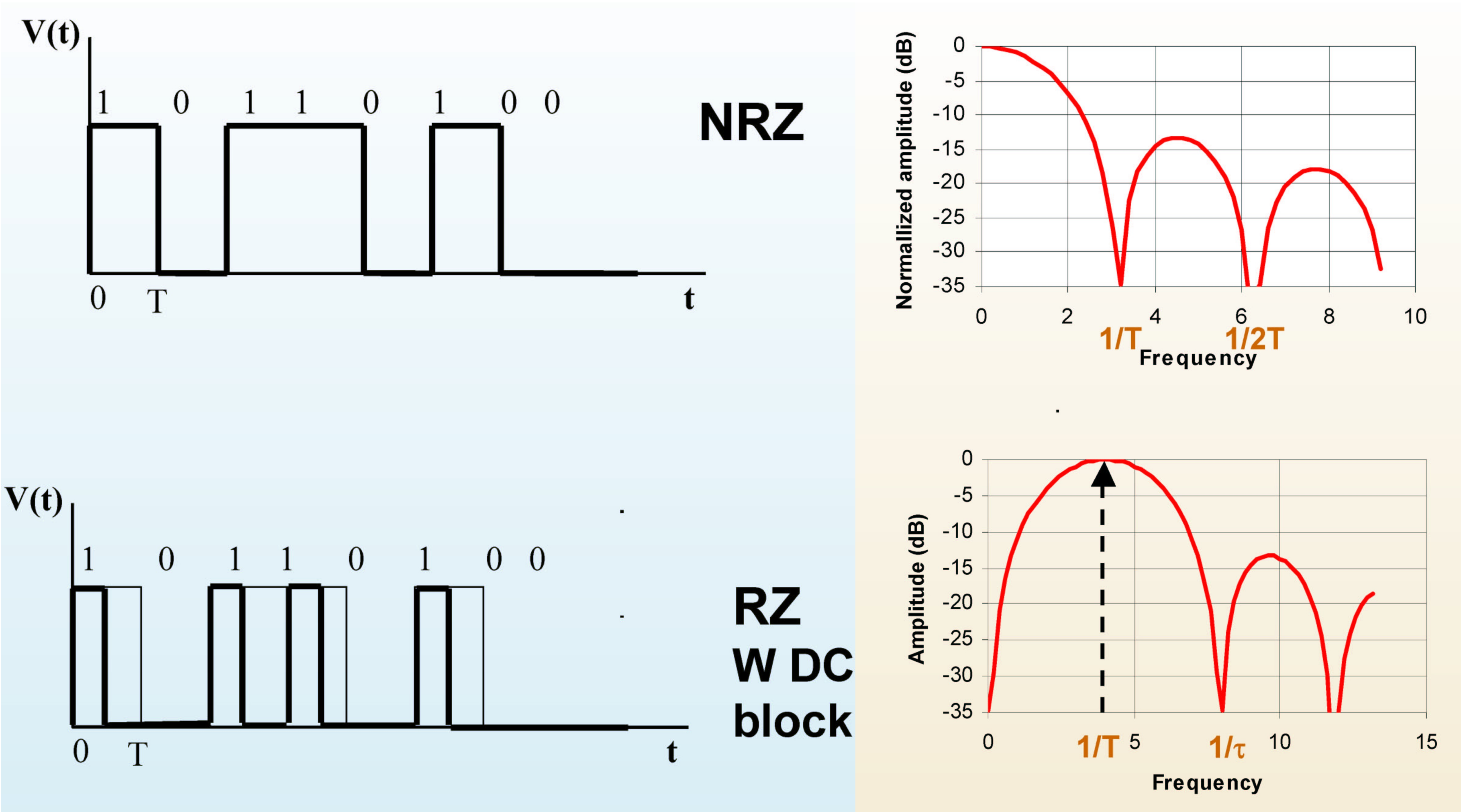
Frequency Shift Keying

Phase Shift Keying

Binary data encoding formats



Spectral content of RZ and NRZ encoded binary data



M-ary PSK and QAM

- M-ary PSK has constant amplitude

$$s_i(t) = A\cos(\omega_0 t + \phi_i)$$

where

$$\phi_i = 2\pi \frac{i}{M}$$

for i=0,1,2 ..M-1, M = 2^n and *n* is the number of bits per symbol.

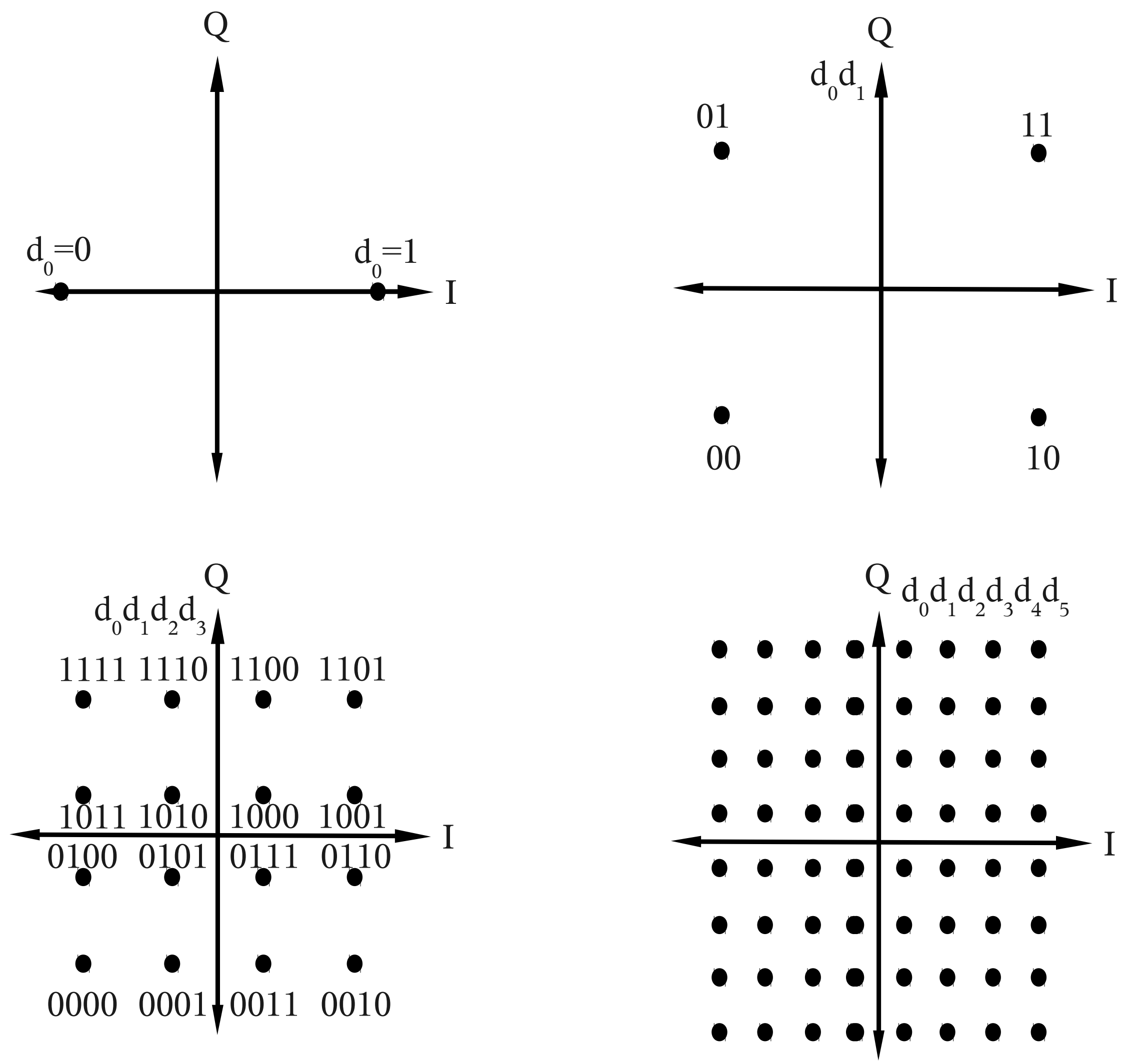
- M-ary QAM

$$s_k(t) = a_k \cos(\omega_0 t) + b_k \sin(\omega_0 t)$$

Ex. 64QAM

$$(a_k, b_k) = \left((-1)^{d_0} \frac{1+2d_2+4d_4}{7}, (-1)^{d_1} \frac{1+2d_3+4d_5}{7} \right)$$

M-ary PSK, QAM Constellations



M-ary PSK, QAM

- Generation
  - By up-conversion
  - By direct modulation
- Detection: coherent only
- QAM modulation systems need very linear power amplifiers

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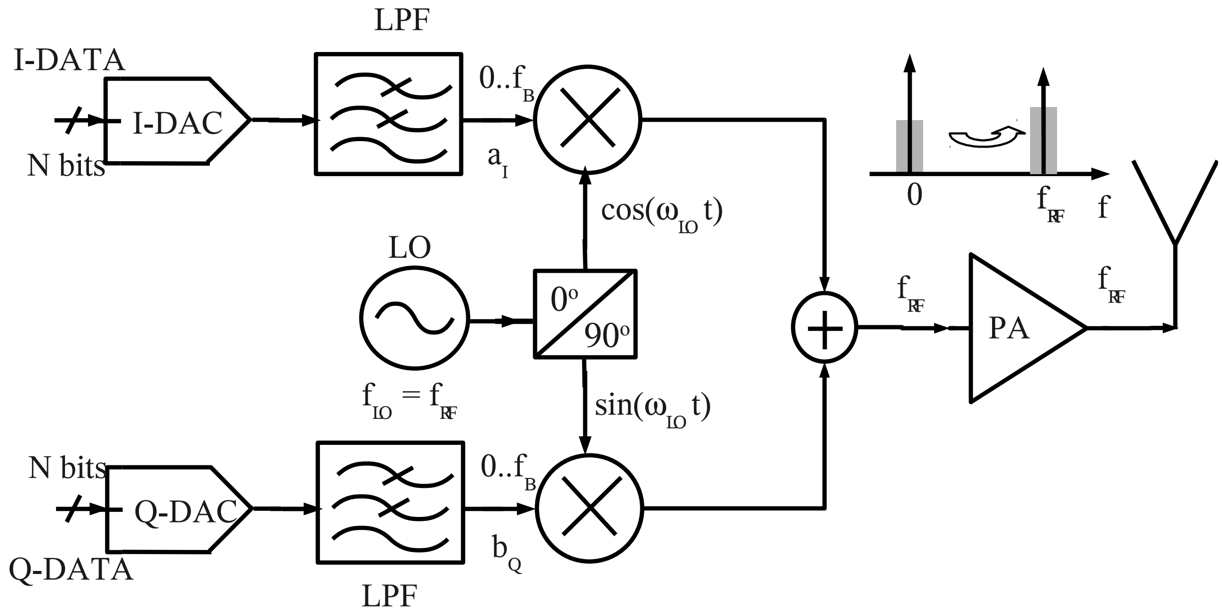
Transmitter Architectures

- The transmitter generates the modulated carrier signal
- Key components
  - carrier synthesizer
  - modulator,
  - power amplifier (PA)
- Modulation of carrier
  - by upconversion (linear)
  - directly, in amplitude, frequency, phase or pulse-width
- Direct modulation can be linear or non-linear (RF-DAC)

Direct Upconversion Transmitter

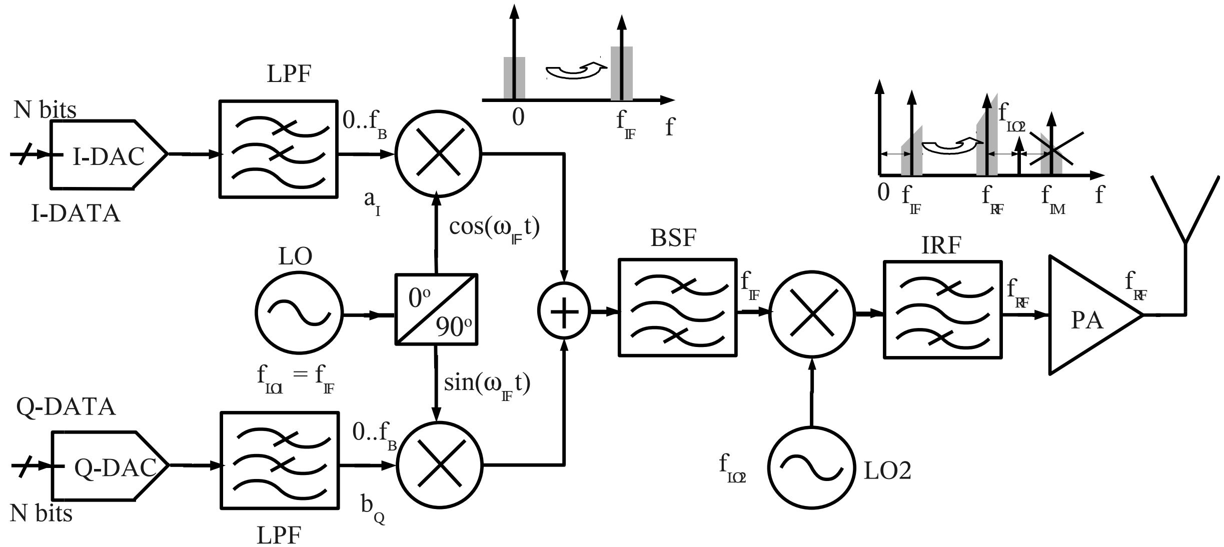


$$s(t)=(a_I+j\,b_Q)e^{-j\omega_{LO}t}=a_I\cos(\omega_{LOF}t)+b_Q\sin(\omega_{LO}t)$$

- An RF filter (not shown) must be placed before PA
- a_I and b_Q are analog baseband signals
- ✗ LO frequency pulling by PA => need lots of buffering between LO and PA

Single-Sideband, 2-step Upconversion

- Most common
- ✔ LO pulling is relieved
- ✗ Additional IF filtering
- ✗ Stringent lInearity specs

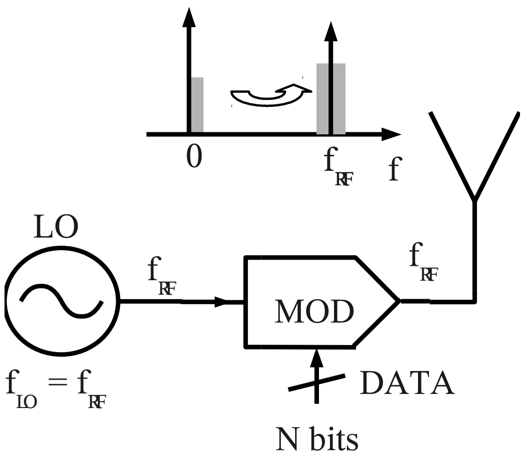


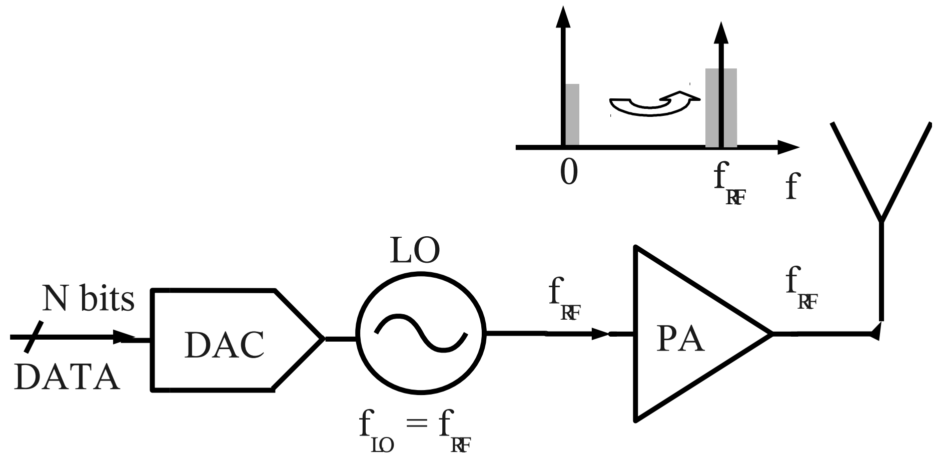
$$s_{IF}(t)=(a_I+j\,b_Q)e^{-j\omega_{IF}t}=a_I\cos(\omega_{IF}t)+b_Q\sin(\omega_{IF}t)$$

$$s(t)=[a_I cos(\omega_{IF}t)+b_Q sin(\omega_{IF}t)]cos(\omega_{LO2}t)=\frac{a_I}{2}cos(\omega_{RF}t)+\frac{b_Q}{2}sin(\omega_{RF}t)$$

$$+\frac{a_I}{2}cos[(\omega_{LO2}+\omega_{IF})t]+\frac{b_Q}{2}sin[(\omega_{LO2}+\omega_{IF})t]$$

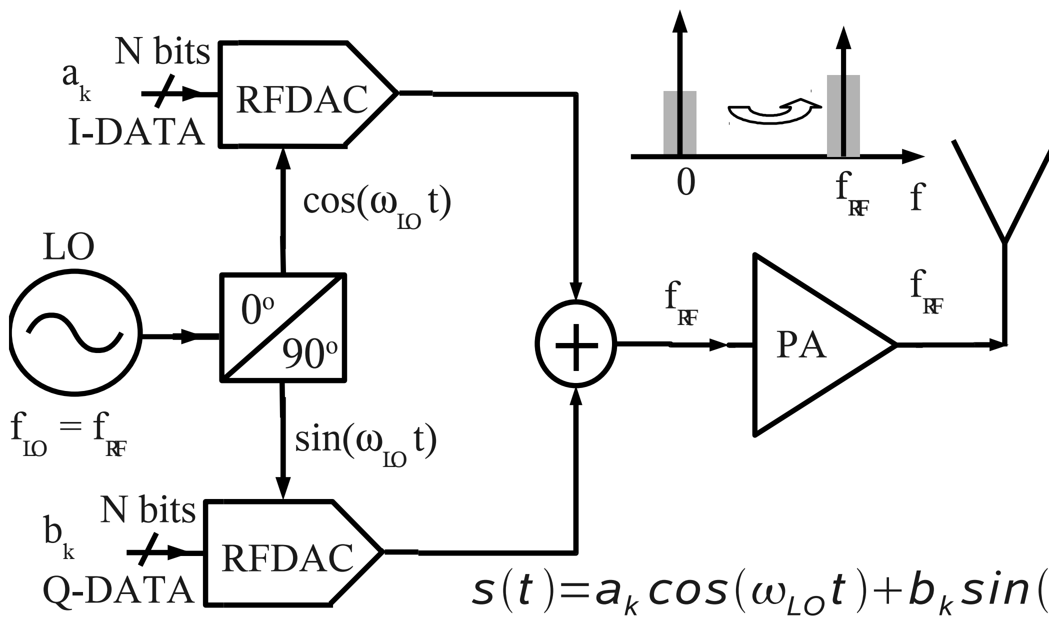
Direct Modulation Transmitter





- ASK
- M-ary QAM
- FSK
- RF-DAC

$$y[k] = a_k + jb_k.$$



$$s(t) = a_k \cos(\omega_{LO} t) + b_k \sin(\omega_{LO} t)$$

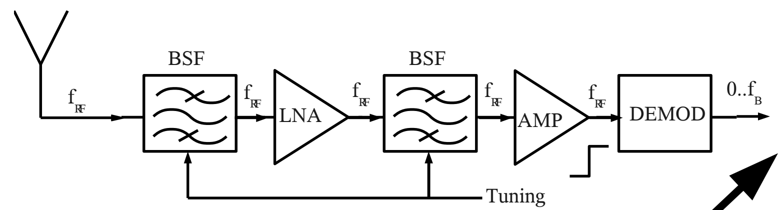
Receiver Architectures

- **Tuned homodyne** receiver (also known as a **direct detection** receiver) (Earliest)
- (Super) **Heterodyne** receiver (Reginald Aubrey Fessenden, 1906, Brant Rock Massachusetts)
- **Direct Conversion (Zero-IF)** Receiver

Direct Detection Receivers

- ✔ Simplest, no VCO/PLL
- ✗ Large gain (>50dB) RF tuned amplifier => unstable
- ✗ Tunable RF filters difficult to integrate

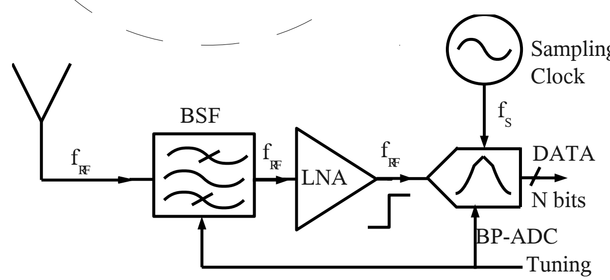
$$s(t) = A(t)\cos(\omega_{RF} t)$$



$$[A_V s(t)]^2 = A_V^2 A(t)^2 \cos^2(\omega_{RF} t) = \frac{A_V^2 A(t)^2}{2}\left[1 + \cos(2\omega_{RF} t)\right]$$

A_V = receiver voltage gain

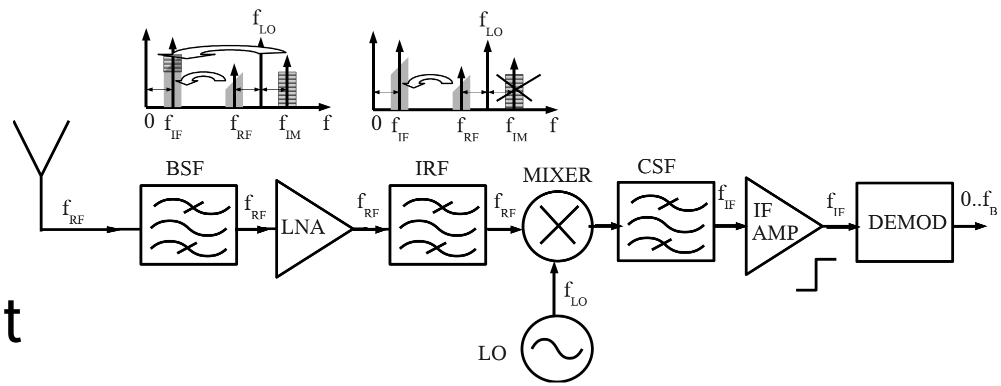
- Digital: Bandpass $\Delta\Sigma$
  - ✗ power hungry
  - ✗ needs VCO



Superheterodyne Receivers

$$s(t)=A(t)cos[\omega_{RF}t+\alpha(t)]$$

- Single step
- Two step
  - ✔ robust,
  - ✗ many filters, difficult to integrate in IC



$$cos(\omega_{LO}t)$$

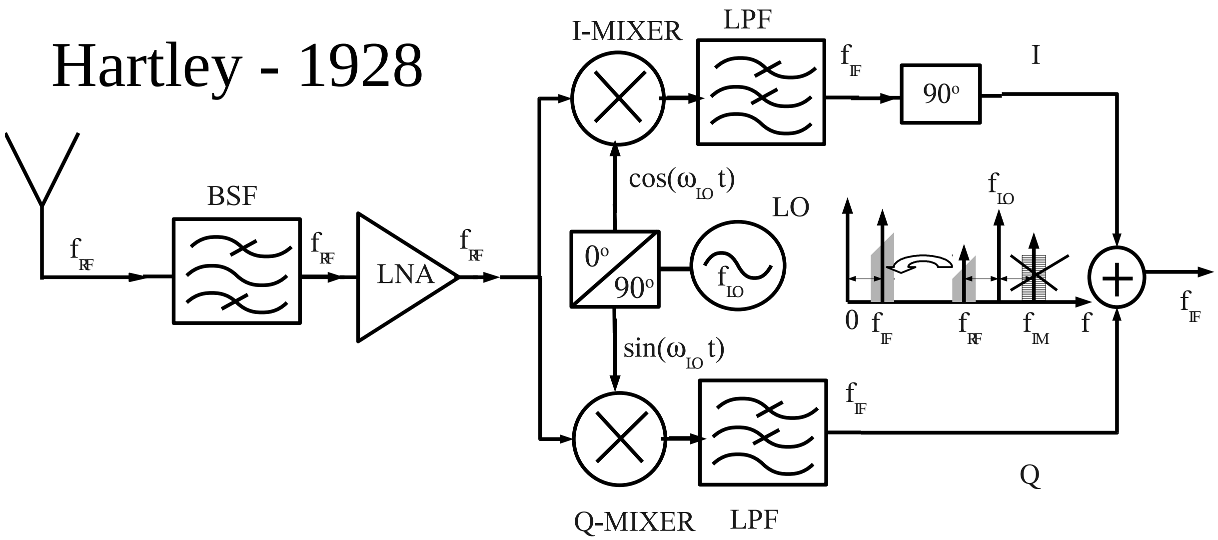
$$f_{RF}=f_{LO}-f_{IF} \;\; or \;\; f_{RF}=f_{LO}+f_{IF}$$

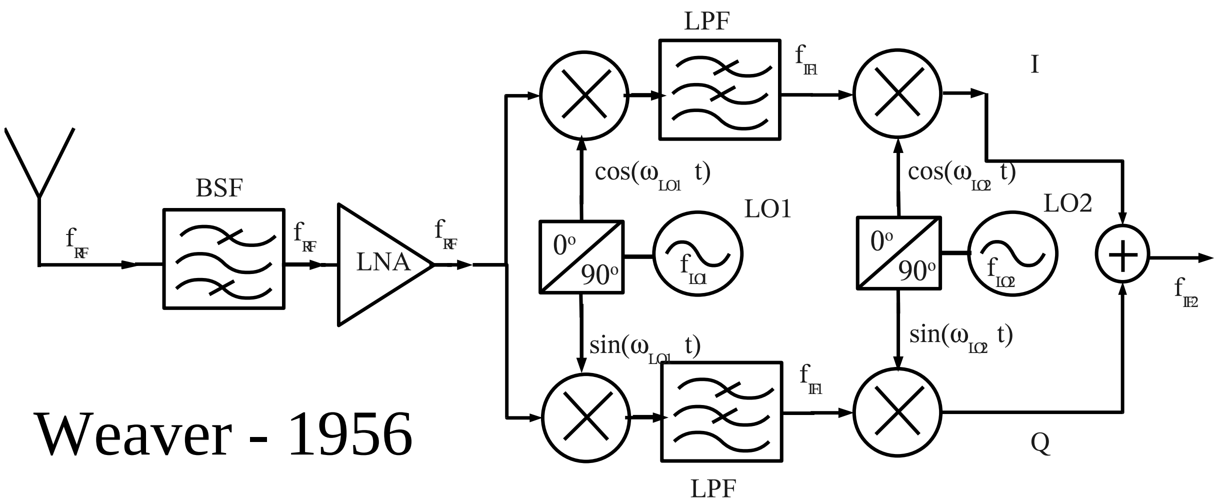
$$s_{IF}(t)=A(t)\cos[\omega_{RF}t+\alpha(t)]\cos(\omega_{LO}t)=>$$

$$\frac{A(t)}{2}\cos[\omega_{IF}t-\alpha(t)]$$

Image Rejection Superheterodyne Receivers

Hartley - 1928

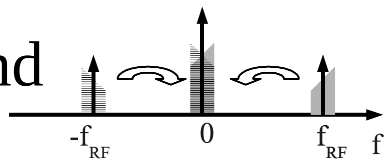


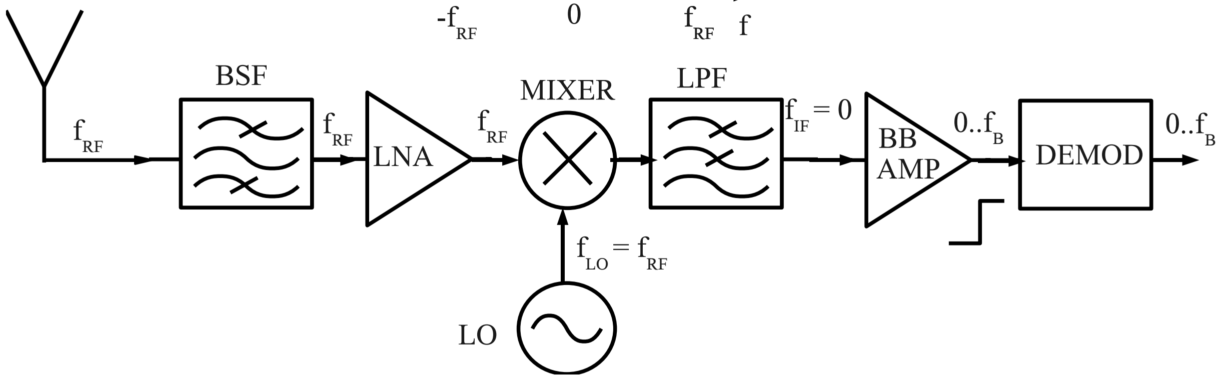


Weaver - 1956

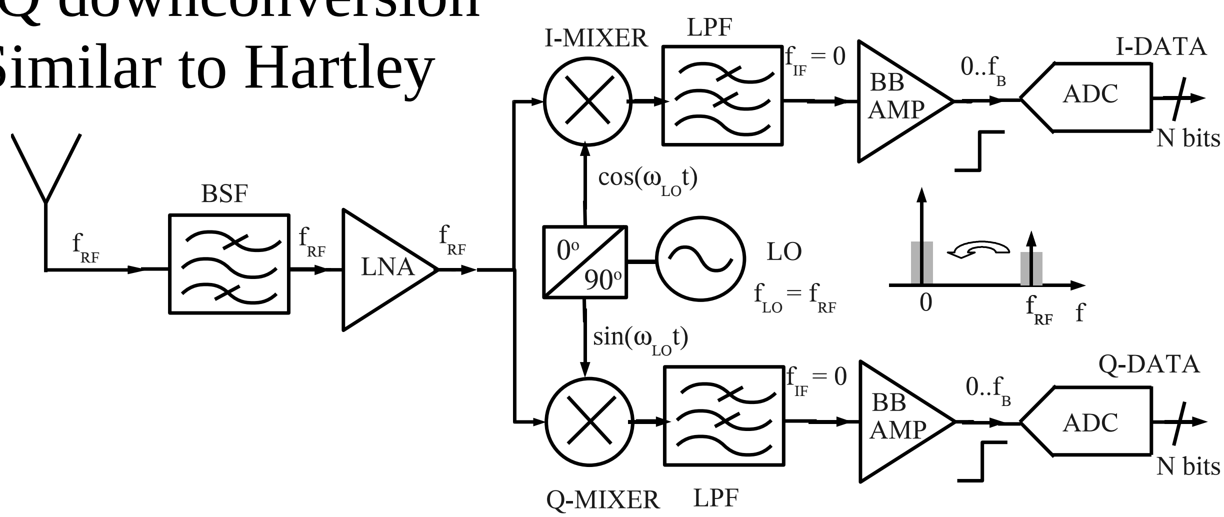
Direct Conversion (0-IF) Receivers

Double-sideband





IQ downconversion
Similar to Hartley



$s(t) = a_I \cos(\omega_{RF}t) + b_Q \sin(\omega_{RF}t)$ $\quad$ $y(t) = a_I + jb_Q.$

- ✔ Simple & easy to integrate
- ✗ LO leakage, pulling, self-mixing
- ✗ Sensitivity to 1/f noise, even order non-linearity
- ✗ Noise figure degradation in monostatic radar

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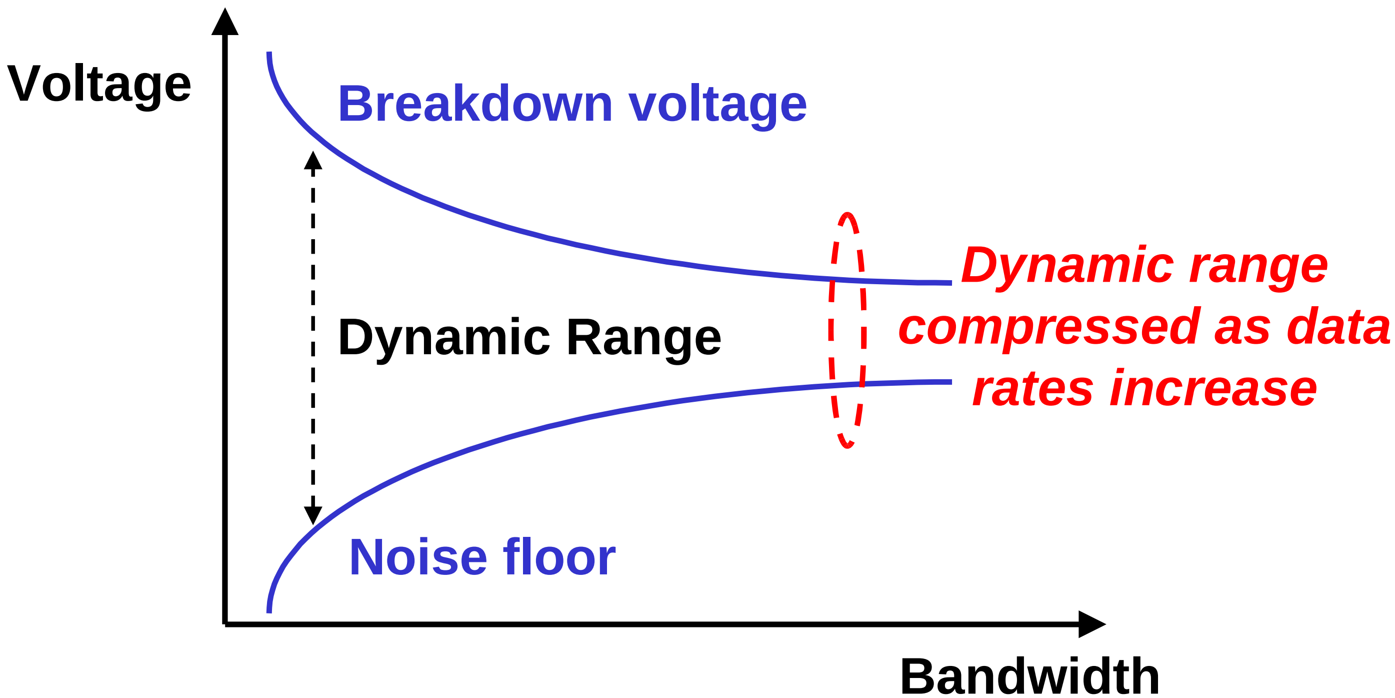
Transceiver Specification

- Receiver specification
  - Frequency, gain
  - Dynamic range
    - Noise figure, noise temperature, sensitivity
    - Linearity
    - PLL phase noise
- Transmitter specification
  - Output power
  - Error Vector Magnitude (EVM)
  - Power Spectral Density (PSD) mask, or ACPR
  - Noise

Fundamental Limitations of Dynamic Range



Emphasizes need for low-noise

Noise

Determines the threshold for minimum signal detection.

Noise can be introduced in receiver in two ways

- From the *external* environment through antenna or signal source impedance
- By *internal* generation in receiver's own circuitry

Common Types of Noise Generated in Circuits

- *Thermal noise* – $kT\Delta f$.
  - caused by the random vibration and motion of carriers due to T.
  - its spectral density kT is constant up to at least 100 GHz.
- *Shot noise* - occurs in active devices
  - due to quantum, random nature of carriers crossing potential barriers.
  - Its spectral density $2qI$ is constant up to high frequencies
- *Flicker noise* – occurs in active devices and sometimes in resistors, and its spectral density follows the $1/f$ law.

Available Noise Power

Is the power that can be transferred from a noise source to a *conjugately-matched load*, whose temperature is *0 °K*, and thus *unable to reflect* back noise power.

Although counter-intuitive, the available noise power from a device/circuit/body/antenna does not depend on its size.

$$P_{available} = \frac{\overline{v_n^2}}{4R} = \frac{4kTR\Delta f}{4R} = kT\Delta f$$

T may be

- 290 °K (terrestrial antenna)
- 30 .. 50 °K (satellite antenna)
- > 290 °K for noise diode or PIN diode (signal and noise source impedance are not equal)

2-Port Noise Factor, Noise Figure

$$F = \frac{SNR_i}{SNR_o} = \frac{SNR_i}{\frac{GP_i}{N_a + GN_i}} = 1 + \frac{N_a}{GN_i} \qquad NF = 10\log_{10} F$$

F is the degradation of the signal-to-noise ratio as the signal passes through the two-port.

G is the power gain of the two-port

N_i is the input noise power (usually the noise floor) available from the antenna

P_i is the input signal power

N_a is the noise power added by the two-port

Two-Port Noise Temperature

$$N_i = kT\Delta f \qquad T_a = \frac{N_a}{kG\Delta f} \qquad F = 1 + \frac{T_a}{T} \qquad (F-1)\times T = T_a$$

- T_a is the equivalent noise temperature of the 2-port.
- T is the ambient temperature of the signal source
- The noise figure is a function of the ambient temperature. Noise figure measurements must specify the ambient temperature at which the measurement occurred.

Friis' Cascaded Noise Factor/Temperature

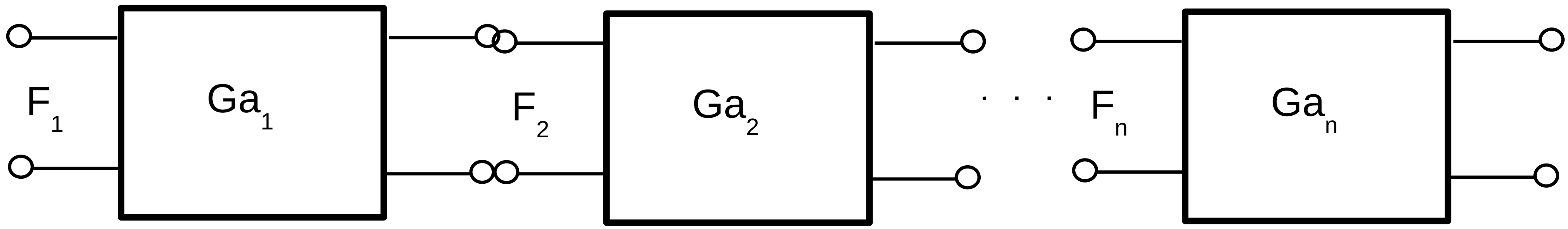
$$F = F_1 + \frac{F_2 - 1}{Ga_1} + \ldots + \frac{F_n - 1}{Ga_1 \times Ga_2 \times \ldots Ga_{n-1}}$$

$$T_a = T_{a1} + \frac{T_{a2}}{Ga_1} + \ldots + \frac{T_{an}}{Ga_1 \times Ga_2 \times \ldots Ga_{n-1}}$$

How do we derive it?

$$M = \frac{F - 1}{1 - \frac{1}{G}}$$

M = noise measure



Receiver Noise Floor and Sensitivity

- Noise floor defined at the output of the receiver, before the decision circuit or demodulator

$$\text{Noise floor} = kT\Delta f G F$$

G= overall gain of the receiver gain

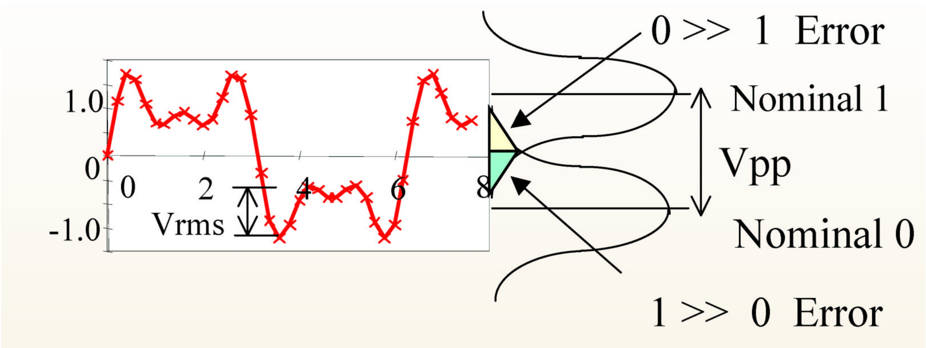
F = receiver noise factor

- Receiver *sensitivity*, S_i,
  - specified for a certain bit error rate
  - depends on the receiver $F(T_a)$ and required SNR_{RX} of the detector

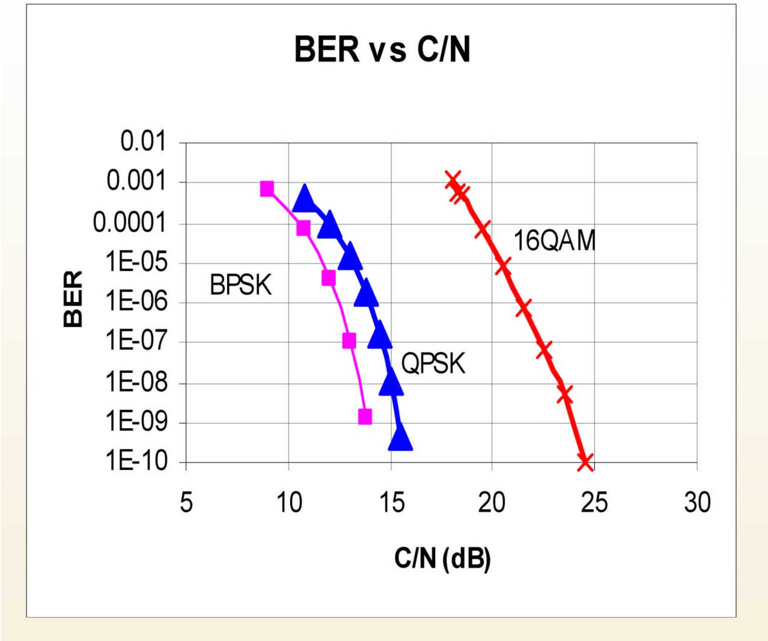
$$S_i = F \cdot SNR_{RX} \cdot k \cdot T \cdot \Delta f = \left(1 + \frac{T_a}{T}\right) SNR_{RX} \cdot k \cdot T \cdot \Delta f$$

Link Between BER and SNR

$$BER = \frac{1}{\sqrt{2\pi}} \frac{\exp\left[-\left(\frac{E_b}{N_o}\right)^2 / 2\right]}{\frac{E_b}{N_o}} \qquad \frac{E_b}{N_o} = SNR \frac{B_n}{R_b}$$



- E_b is the bit energy
- N_o is the one-sided noise power density
- B_n is the noise bandwidth: $\pi\Delta f/2$
- R_b is the data rate



SNR vs. Modulation Scheme

Modulation	Efficiency*	SNR@BER=10^{-6}
OOK-NRZ	1.0 (1) bits	13 dB
4L FSK	1.5 (2) bits	17 dB
QPSK	1.6 (2) bits	14 dB
8PSK	2.5 (3) bits	19 dB
16QAM	3.2 (4) bits	21 dB
64QAM	5.0 (6) bits	27 dB

*)Bandwidth efficiency = data rate/bandwidth = (ideal)

$$\frac{R_b}{\Delta f}$$

Receiver Sensitivity Examples

5-GHz Wireless LAN System

- NF = 6 dB, B=20 MHz, QPSK => SNR =14 dB

$$S_i(BER=1E\text{-}6)=-174\ \text{dBm}\ +6+10\log\left(\frac{20\,\text{MHz}}{1\,\text{Hz}}\right)+14=-174+6+73+14=-81\,\text{dBm}$$

60-GHz, 1.5-Gb/s Wireless LAN System

- NF = 9 dB, B = 1 GHz, QPSK => SNR =14 dB

$$S_i(BER=1E\text{-}6)=-174\ \text{dBm}\ +9+10\log\left(\frac{1\,\text{GHz}}{1\,\text{Hz}}\right)+14=-174+9+90+14=-61\,\text{dBm}$$

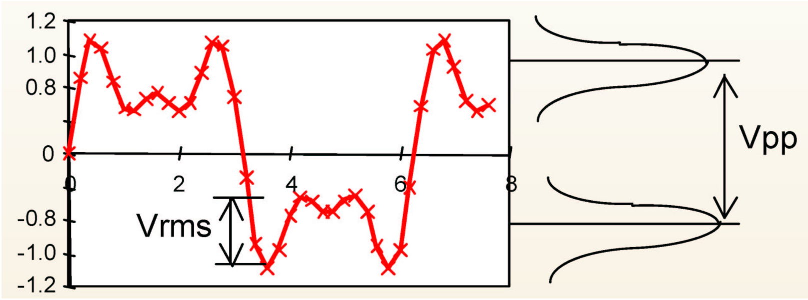
12-GHz, HDTV satellite receiver

- T=30°K, NF=1dB, B=6MHz, 64QAM=>SNR=27dB

$$S_i(BER=1E\text{-}6)=-184\ \text{dBm}\ +3.5+10\log\left(\frac{6\,\text{MHz}}{1\,\text{Hz}}\right)+27=-184+3.5+68+27=-85.5\,\text{dBm}$$

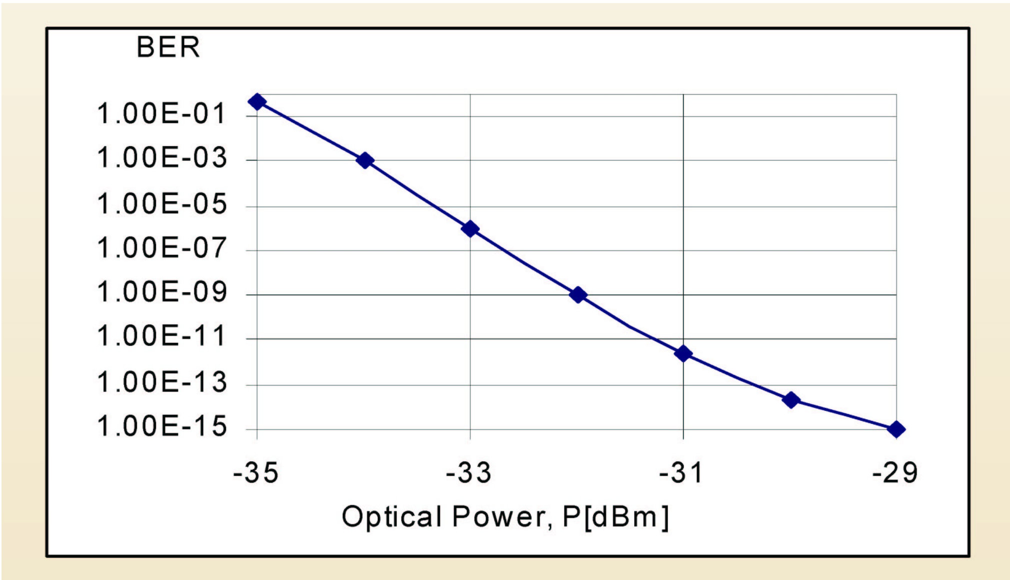
Equivalent Noise Current and Receiver Sensitivity in Fiber Systems

$$S_i = \frac{Q\, i_n^{rms}}{R} \qquad BER \approx \frac{1}{\sqrt{2\pi}} \frac{\exp\left[-Q^2/2\right]}{Q}$$

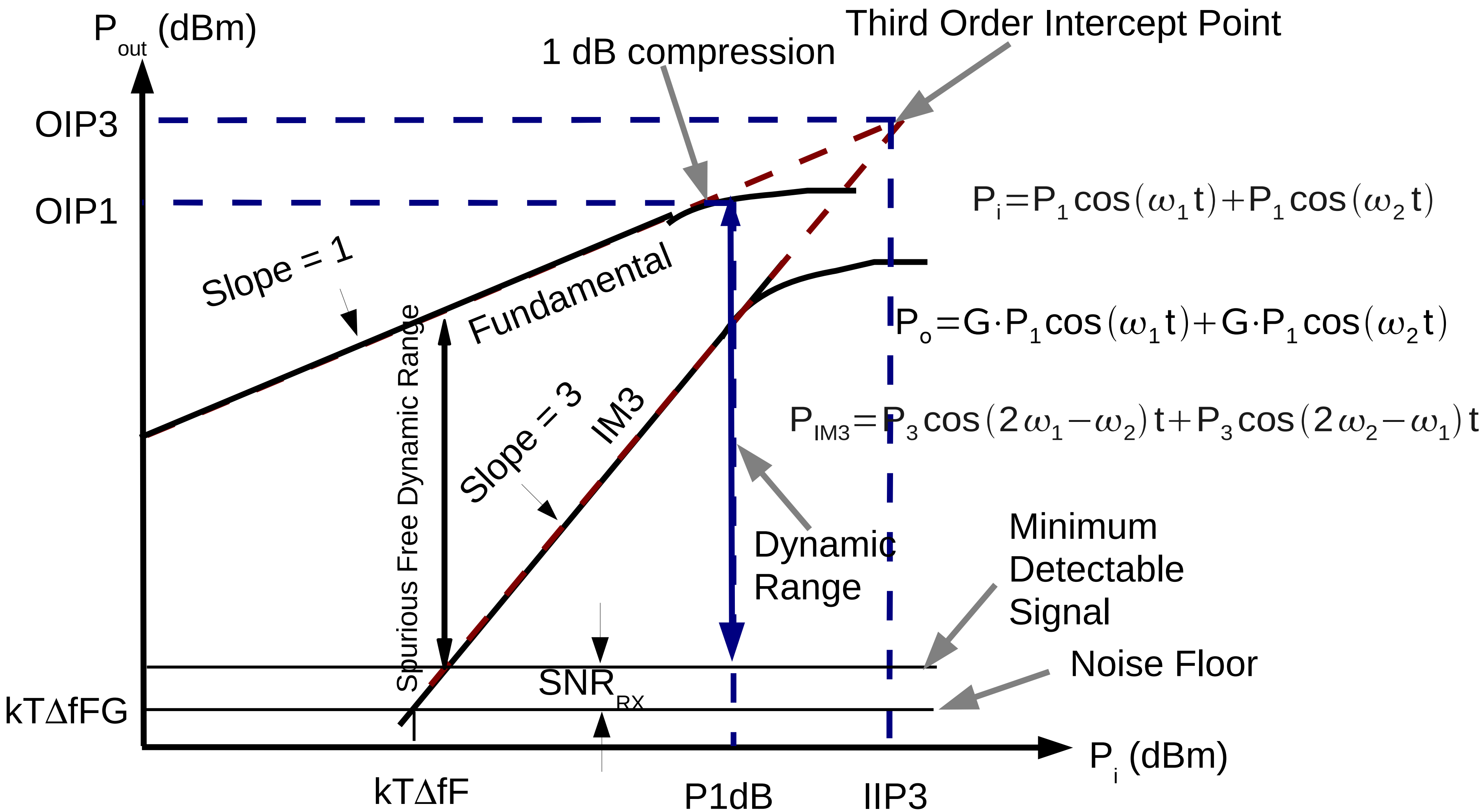


- *Q* is the eye quality factor
- I_n^{rms} is the equivalent input noise current of the fiberoptic receiver
- *R* is the photodiode responsivity (A/W)

There is a direct link between *Q* and the bit error rate BER



Linearity Figures of Merit



Linearity and Distortion

- The 1-dB compression point: P_{1dB}
- Third-order intercept point: *IIP3, OIP3*
- Second-order intercept point *IIP2* (critical in 0-IF receivers)
- Broadband measures of linearity
  - Total Harmonic Distortion (THD) = the sum of the powers of all the harmonics, except the fundamental

Equations for IIP_n and DR

- IIP_n can be obtained graphically using dB scales

$$IIP_n = \frac{nP_i - IM_n}{n-1}$$

- Output dynamic range with respect to the *n*-th spur

$$DR_n = \left(1 - \frac{1}{n}\right)(IIP_n - IM_n)$$

- Spurious signals due to an interferer *I*, must be *C* dB below the sensitivity level (*C* = margin)

$$IM_n = S_i - C \Rightarrow IIP_n = \frac{nI - (S_i - C)}{n-1}$$

Example of IIP3 spec derivation

- $I_1 = I_2 = -38$ dBm, $S_i = -60$ dBm, C = 14 dB, $f_1 = 64$ GHz, $f_2 = 62$GHz => IM3 at 60GHz

$$IIP_3 \geq \frac{3\times(-38)-(-60-14)}{2} = \frac{-114+74}{2} = -20\,\text{dBm}$$

- The most relaxed IIP3 requirement is in the absence of an interferer. Eg. in OFDM systems, the sub-carriers can interfere with each other:

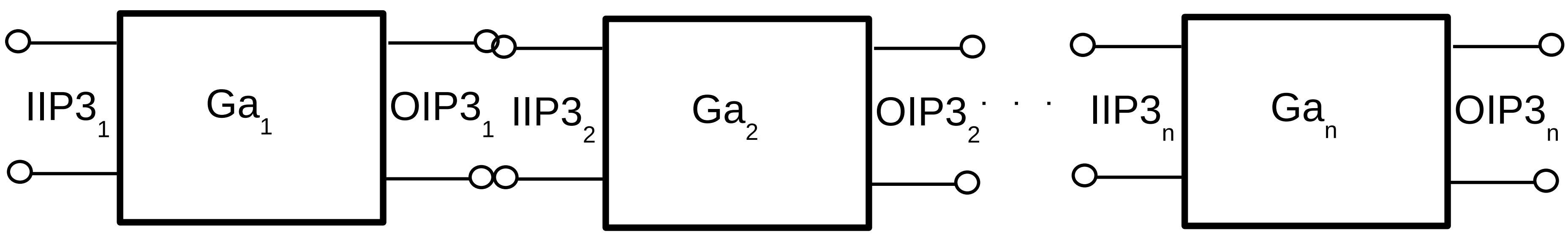
$I_1 = I_2 = -60$ dBm, $S_i = -60$dBm, C = 40dB, $f_1 = 60.005$GHz, $f_2 = 60.01$GHz => IM3 at 60GHz

$$IIP_3 \geq \frac{3\times(-60)-(-60-40)}{2} = -60+\frac{40}{2} = -40\,\text{dBm}$$

Linearity of a Chain Two-Ports

$$\frac{1}{IIP3} = \frac{1}{IIP3_1} + \frac{Ga_1}{IIP3_2} + \frac{Ga_1\,Ga_2}{IIP3_3} + \ldots + \frac{Ga_1\,Ga_2 \ldots Ga_{n-1}}{IIP3_n}$$

$$\frac{1}{OIP3} = \frac{1}{OIP3_n} + \frac{1}{Ga_n\,OIP3_{n-1}} + \frac{1}{Ga_n\,Ga_{n-1}\,OIP3_{n-2}} + \ldots + \frac{1}{Ga_2 \ldots Ga_n\,OIP3_1}$$

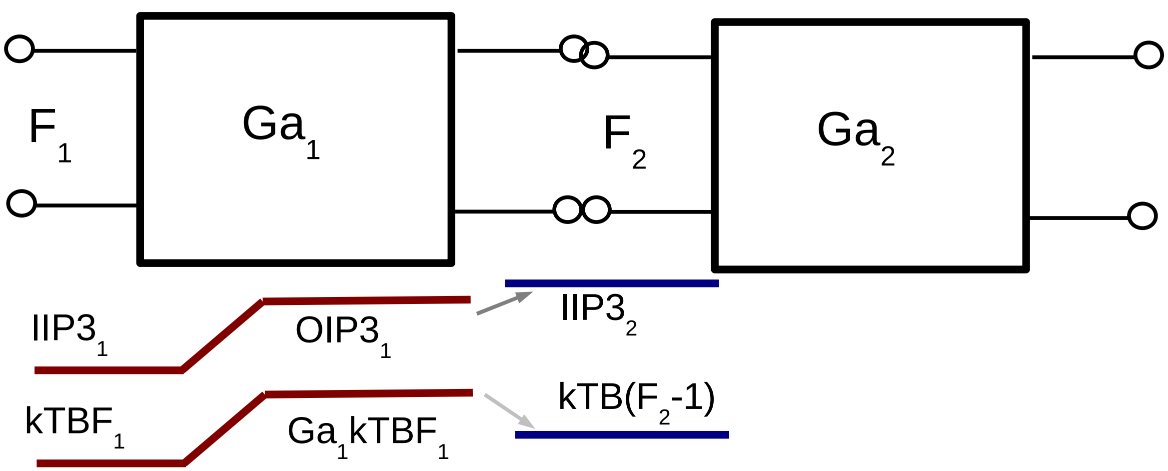


Ga_i is the available power gain of stage i (i.e. power gain when its input and output are conjugately matched to the impedance of the preceding and of the following stages.

Dynamic Range in Cascaded 2-Ports

Designing for maximum dynamic range:

- Make the noise level at the output of the first stage match that at the input of the second, third ... stages.
- Each stage contributes noise and distortion equally.

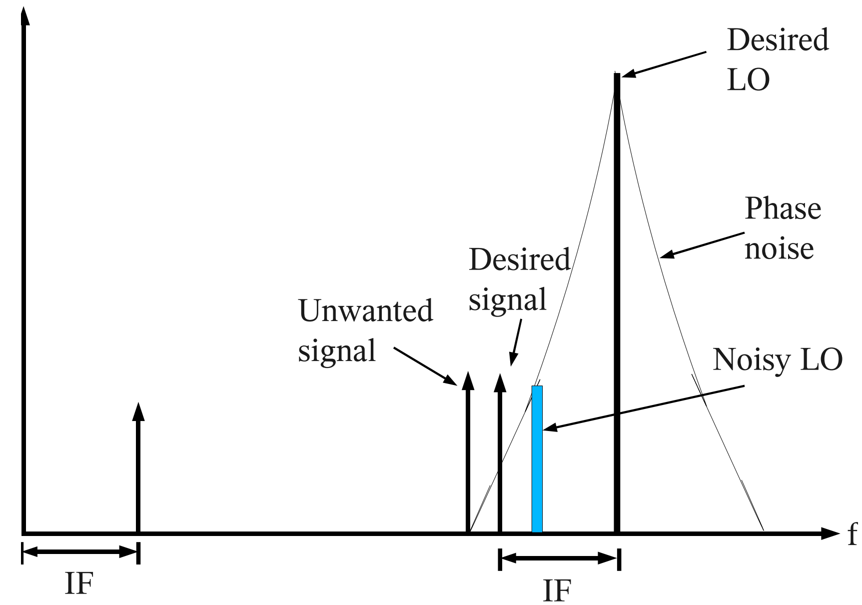


$$Ga_1 = \sqrt{\frac{F_2 - 1}{F_1}\frac{IIP3_2}{IIP3_1}}$$

$$Ga_2 = \sqrt{\frac{F_3 - 1}{F_2}\frac{IIP3_3}{IIP3_2}}$$

PLL Phase Noise: *L*

- *L* = *a* broadening of the oscillator spectrum
- Measured as noise power in dBc/Hz in a 1-Hz band at an offset frequency f_m relative to the carrier power and frequency f_{OSC}



- Degrades the receiver sensitivity by raising the noise floor and adding *rms* phase error
- Degrades transmitter *EVM*
- Affects velocity resolution in automotive cruise control radar

Ex. of PLL Phase Noise Impact

- A phase noise of -100dBc/Hz over a bandwidth of 5 MHz results in a rms phase error

$$\theta_{rms} \approx \sqrt{2 L \Delta f} = 0.0316\, \text{rads} = 1.81^{o}$$

- Oscillator phase noise mixes with an undesired signal and is downconverted to the IF/baseband raising the noise floor and dictating how closely adjacent channels may be spaced
- Formula that gives maximum allowed phase noise to achieve adjacent channel rejection of *C* dB for an interference power *I*

$$L(f_m) = P(\text{dBm}) - C(\text{dB}) - I(\text{dBm}) - 10\log(\Delta f), (\text{dBc/Hz})$$

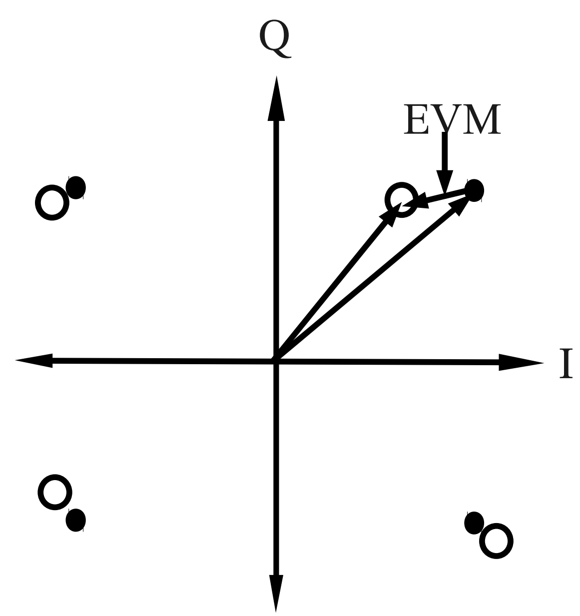
In a 60-GHz radio with P=-60 dBm, C=14 dB, I=-38dBm, Δf=2.1GHz,

f_m = 2.1 GHz, L(2.1GHz) =<-128.5 dBc/Hz

Transmitter Specification

- *Output power* in dBm ($V_{pp}/50\Omega$) in fiber systems
- *EVM =* distortion in the transmitted signal constellation
  - measured with an ideal IQ receiver

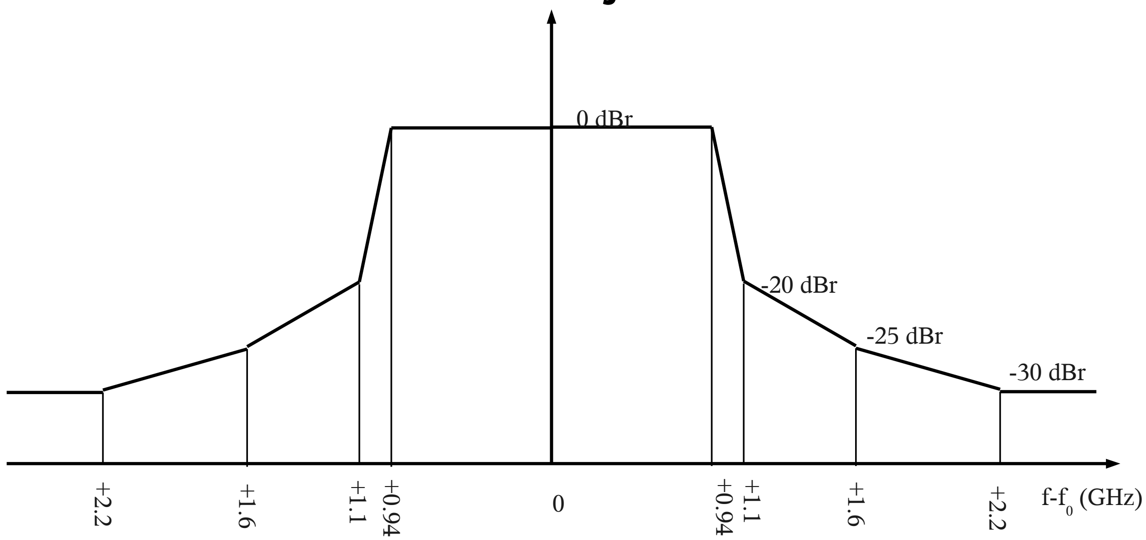
$$EVM = \sqrt{\frac{1}{1000 \cdot P_{avg}} \sum_{i=1}^{1000} \left[(I_i - I^*_i)^2 + (Q_i - Q^*_i)^2 \right]}$$



P_{avg} = average power of the constellation, (I_i^*,Q_i^*) are the complex coordinates of the i^th measured symbol, and (I_i,Q_i) are the complex coordinates of the nearest constellation point.

Power Spectral Density Mask, ACPR

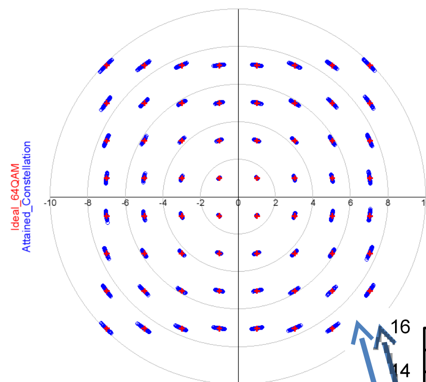
- *PSD (eye mask in fiber)* is specified in dB relative to the signal power in the center of the channel to prevent unwanted emissions into adjacent channels



- Adjacent Channel Power Ratio (ACPR) is another parameter specified in dB: 10log10.
    - Ex: <-20 dBc for 16 QAM, <-43 dB for 64 QAM

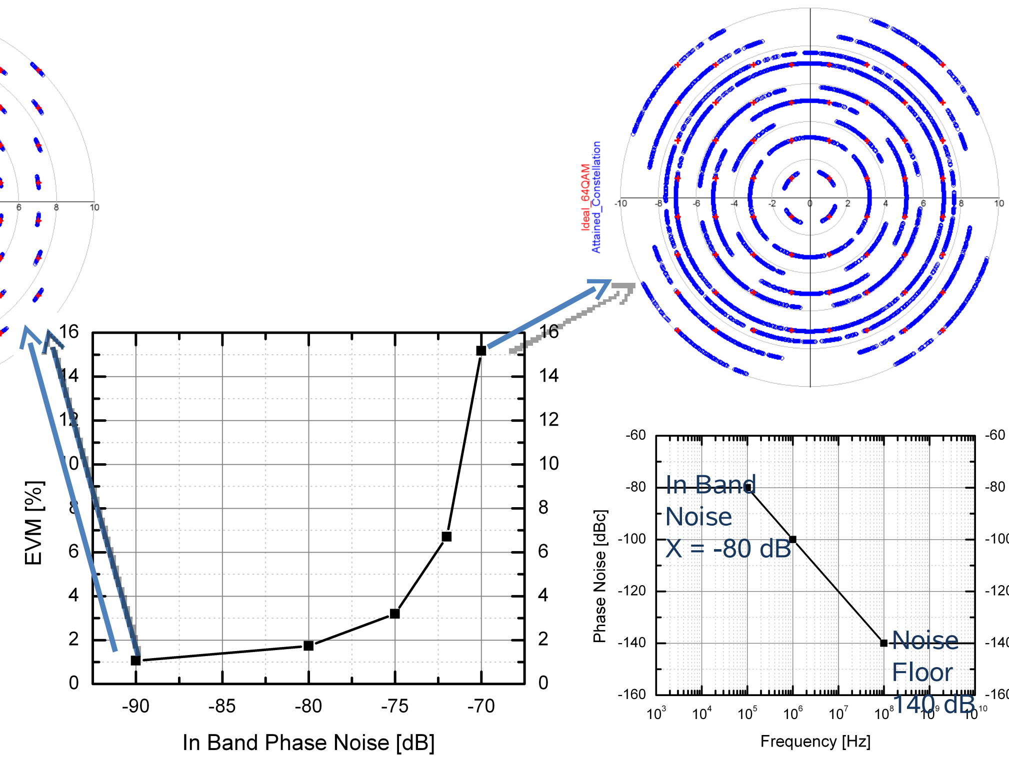
Transmitter noise

- *Transmitter noise (jitter),* mostly from PLL, degrades EVM.
- In fiberoptic transmitters the maximum allowed jitter of the output eye is specified



[Sims by A. Balteanu]

Phase Noise Profile = list(100kHz, X dB, 1MHz, (X-20) dB, 100MHz, -140 dB)



Outline

- Wireless and wireline systems
- Modulation schemes
- Transceiver architectures
- Transceiver specification
- Link Budget
- Phased Arrays
- Other wireless systems
- Baseband data formats and analysis

Link Budget. *LB*

- Describes the combined transmitter, antenna, and receiver performance

$$LB = P_{RX} - S_i \qquad\qquad P_{RX}(d) = P_{TX} G_{TX} G_{RX} \left(\frac{\lambda}{4\pi d} \right)^2$$

- λ is the wavelength
- *d* is the distance between the TX and RX >>λ
- P_{RX} is the power at the receiver input
- P_{TX} is the power at the transmitter output
- G_{TX} is the gain of the TX antenna (dBi)
- G_{RX} is the gain of the RX antenna (dBi)
- *EIRP* = $P_{TX}G_{TX}$ effective isotropic radiated power (*dBmi*)

Example: 60-GHz Link Budget

- 2-m 60-GHz link at 4 Gb/s with Δf of 2 GHz
- QPSK modulation, SNR = 14 dB, NF = 7 dB

$$S_i = -174\,\mathrm{dBm} + 10\log_{10}(2\times 10^9) + 7\,\mathrm{dB} + 14\,\mathrm{dB} = -174 + 93 + 7 + 14 = -60\,\mathrm{dBm}$$

- P_{TX} = 10 dBm, G_{TX}= G_{RX}= 8 dBi, λ = 5 mm

$$LFS = 20\cdot\log_{10}\left(\frac{\lambda}{4\pi d}\right) = 74\,dB$$

$P_{RX} = P_{TX} + G_{TX} - LFS + G_{RX} = 10+8-74+8=-48\mathrm{dBm}$

$LB = P_{RX} - S_i$ = -48 dBm + 60 dBm = 12 dB

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Channel Capacity

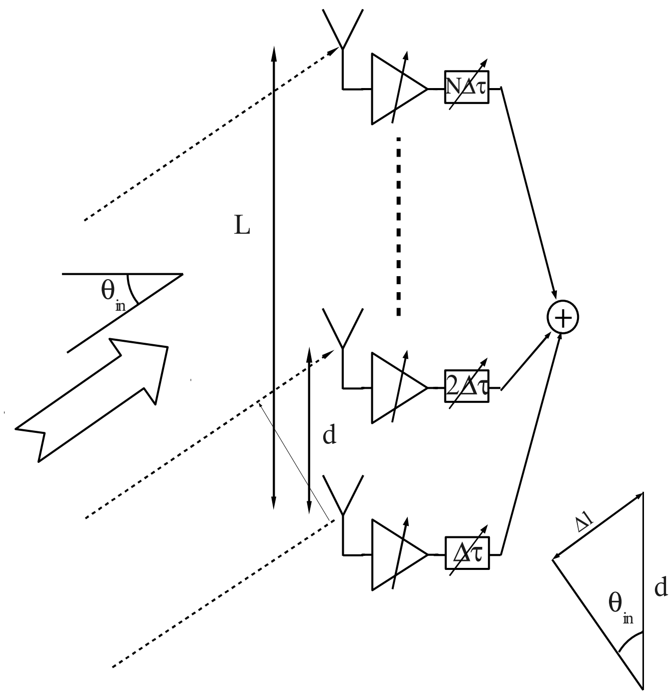
- Shannon's channel capacity, *C* [b/s]

$$C = \Delta f \cdot \log_2(1 + SNR)$$

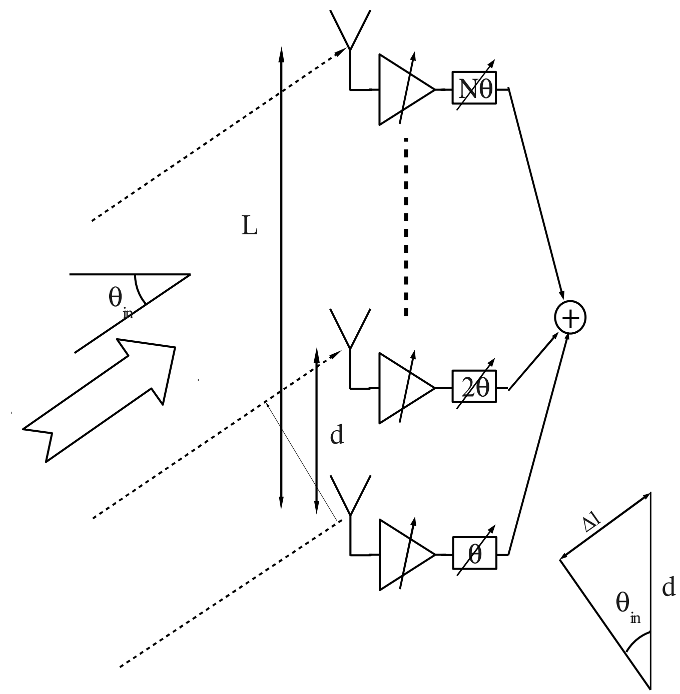
- Fundamental limit for single TX-RX systems
- Improved with spatial diversity:multiple TX, RX)
  - Multiple-Input-Multiple-Output (MIMO)
    - all channels are statistically independent
    - require nonlinear data processing to improve SNR => expensive
  - Phased Arrays
    - at least part of TX, RX is shared
    - Linear: delay + sum processing

Phase Array

- Used in Marconi's 1901 transatlantic radio transmission
- First electronically steered phased arrays in WWII

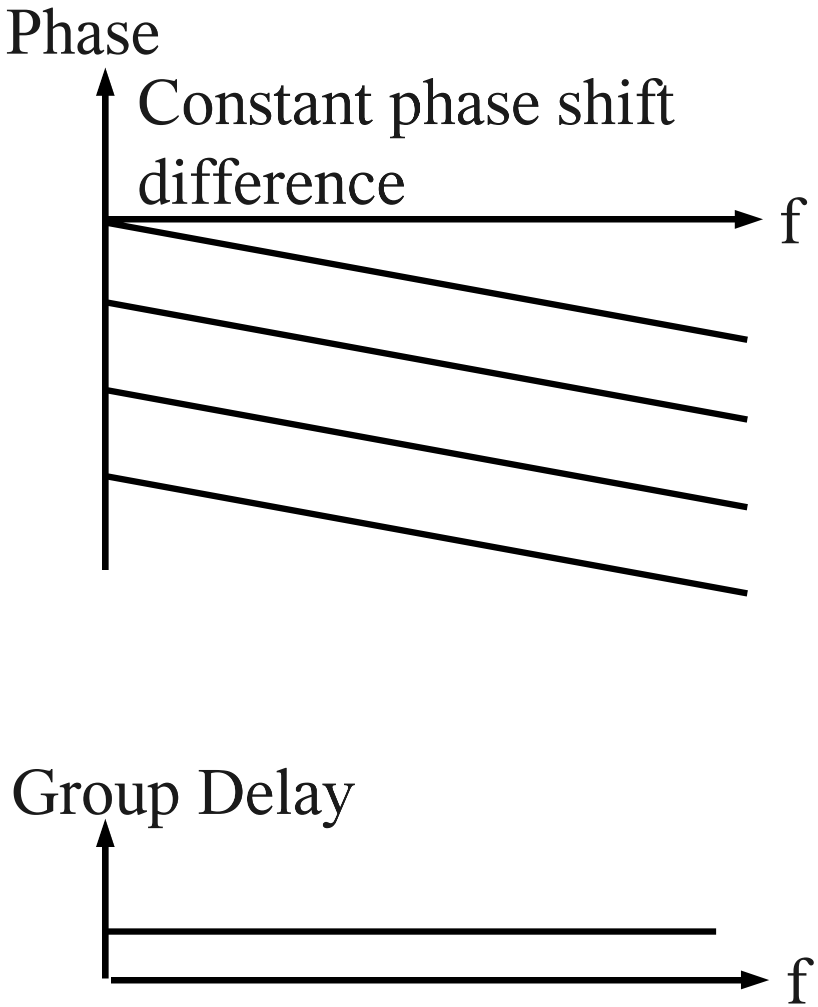


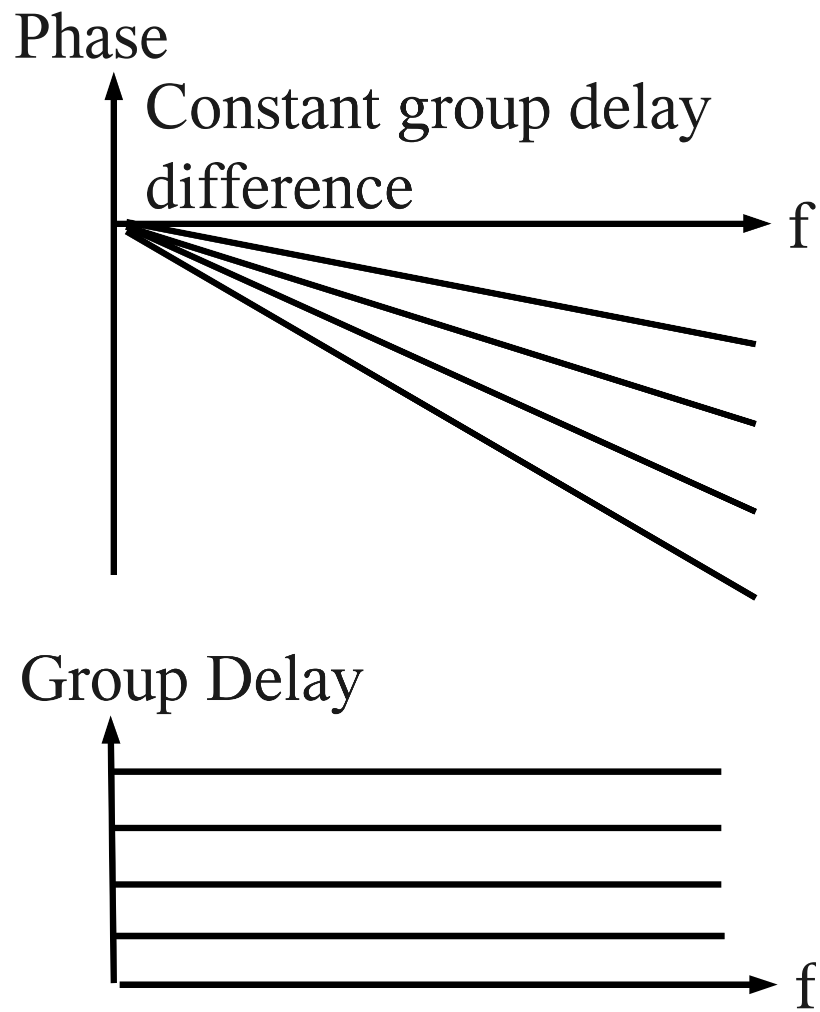
True time delay



Phase delay

Variable phase vs. variable group delay





Properties of Phased Arrays (i)

- The signal before reaching the delay cell *i*

$$s_i(t) = G_{vi} A \cos\left(\omega\left[t-(i-1)\frac{d}{c}\sin(\theta_{in})\right]\right)$$

- To compensate for the propagation delays

$$\Delta\tau = \frac{d}{c}\sin(\theta_{in})$$

- After summing node, signal scales linearly in *N*

$$s_{out}(t) = \sum_{i=1}^{N} G_v A \cos[\omega[t-(i-1)\frac{d}{c}\sin(\theta_{in}) + i\Delta\tau]] = N G_v A \cos[\omega(t+\Delta\tau)]$$

Phased Array Parameters

- **Array gain** (factor) =power gain of array divided by power gain of single element

P_{in} = (N/2)A^2, P_{out} = $N^2G^2A^2/2$, G = P_{out}/P_{in}= NG^2_V

- **Beam-pointing angle** = angle of incidence for which the array gain is maximized

$$\theta_m = \arcsin\left(\frac{c}{d}\Delta\tau\right)$$

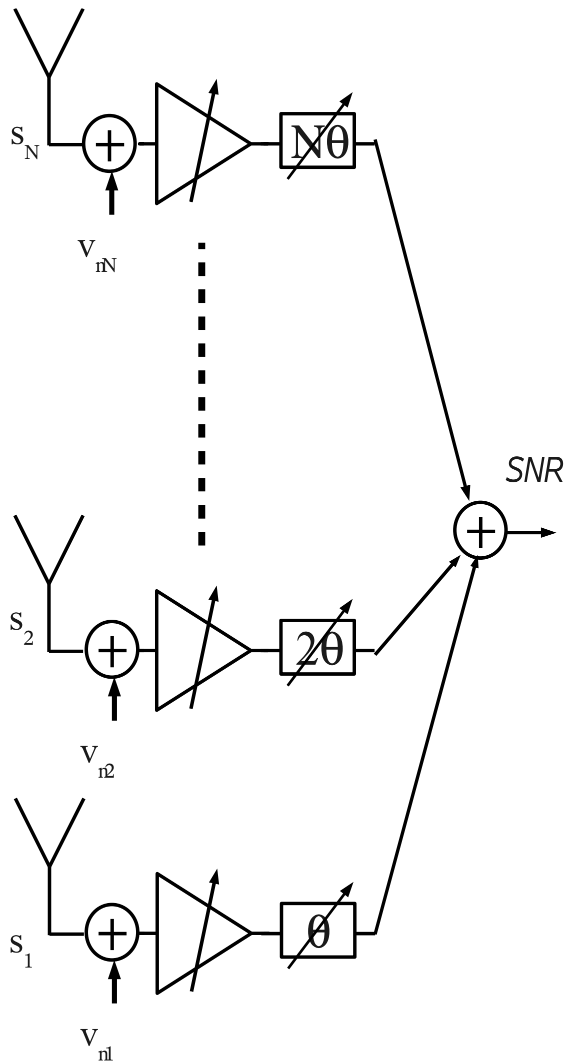
- **Beam width** = λ/L where L = (N-1)d

Typically d =λ/2

Benefits of Phased Arrays

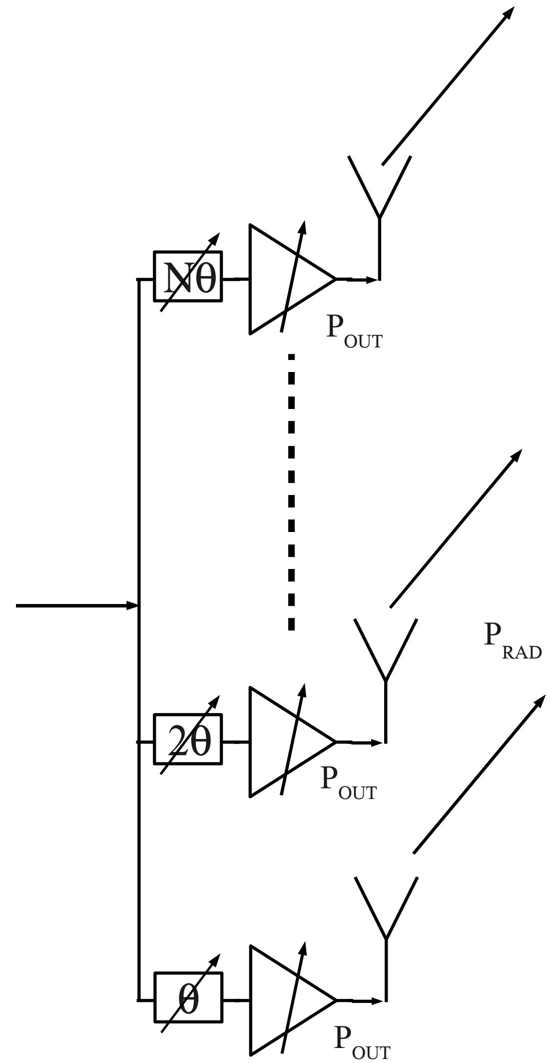
- increased *signal level* at receiver output (the aggregate antenna gain increases N times)
- increased overall output power, scales with N
- Increased *EIRP* which scales as N^2
- immunity to interferers, but only within the bandwidth of each receiver and only after the signal-summation block,
- immunity to multi-path fading through antenna diversity in both RX and TX.

SNR and P_{out} Improvement



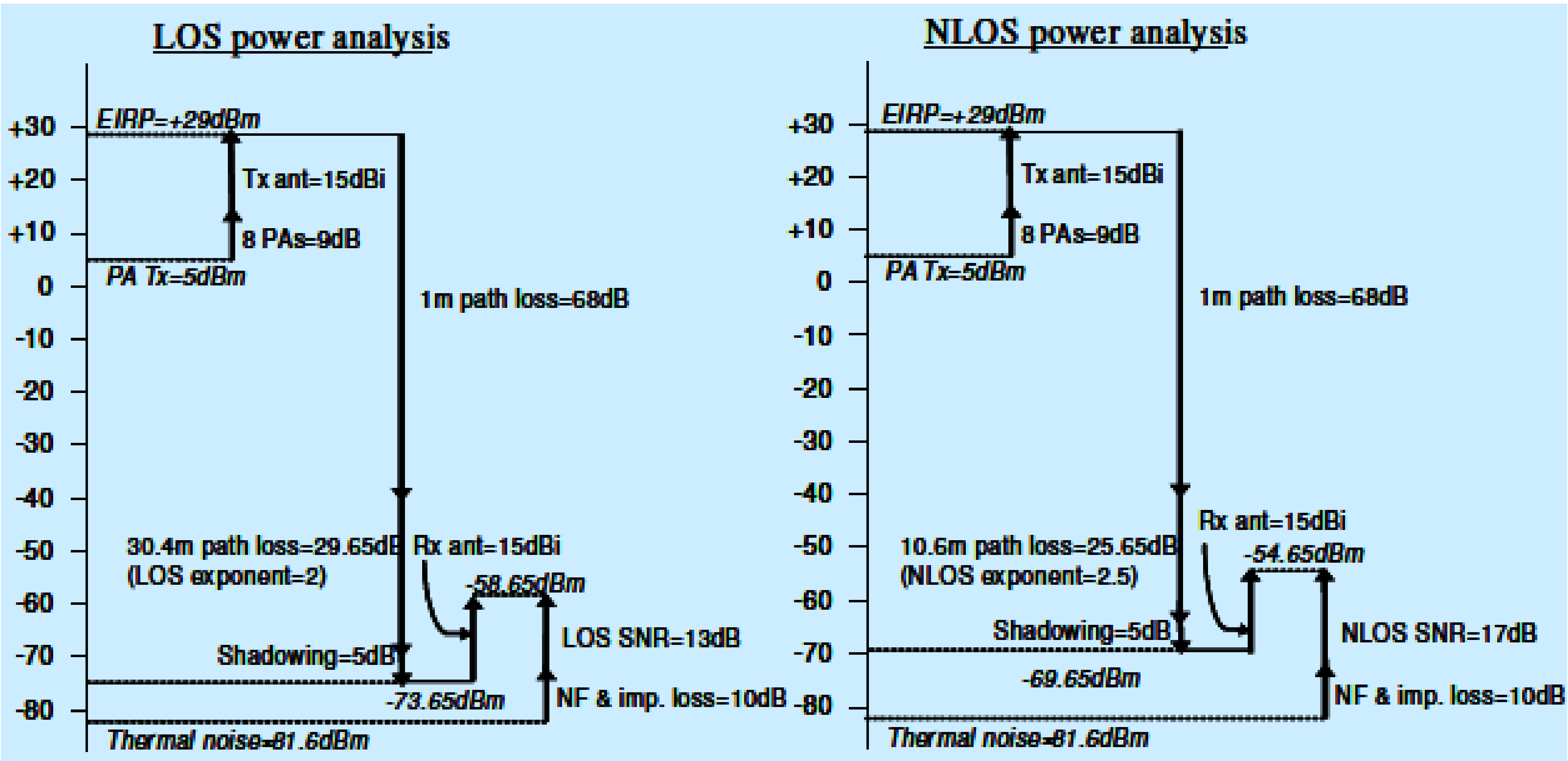
$$SNR_{OUT} = \frac{N^2 A^2 G_V^2}{2\,N\,G_V^2 v_n^2} = N\,\frac{A^2}{2\,v_n^2}$$

$$where\ v_{ni} = v_n$$



$P_{RAD} = N\,P_{OUT}$

Ex. 60-GHz LOS, NLOS link budgets



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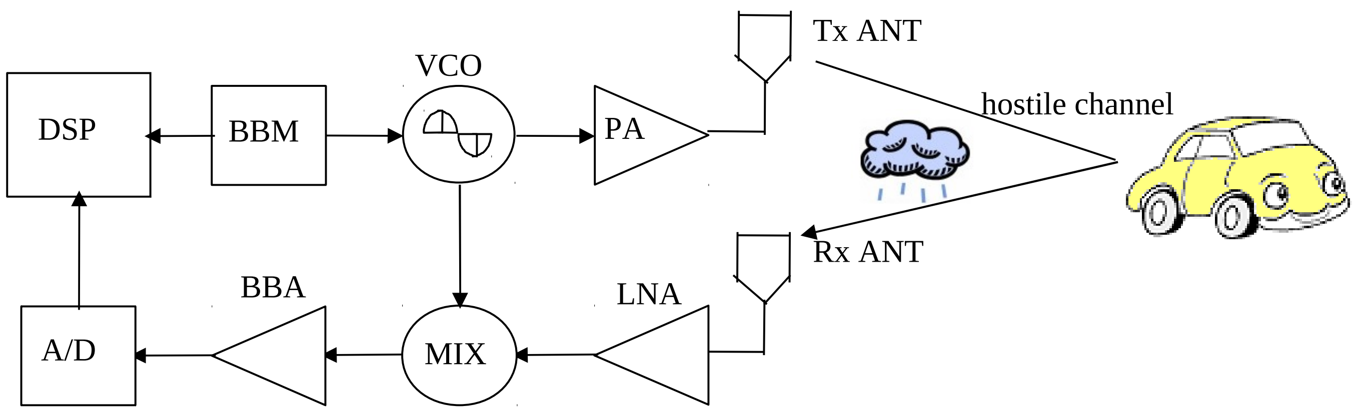
Automotive Radar

- An FMCW Doppler radar = a direct conversion radio

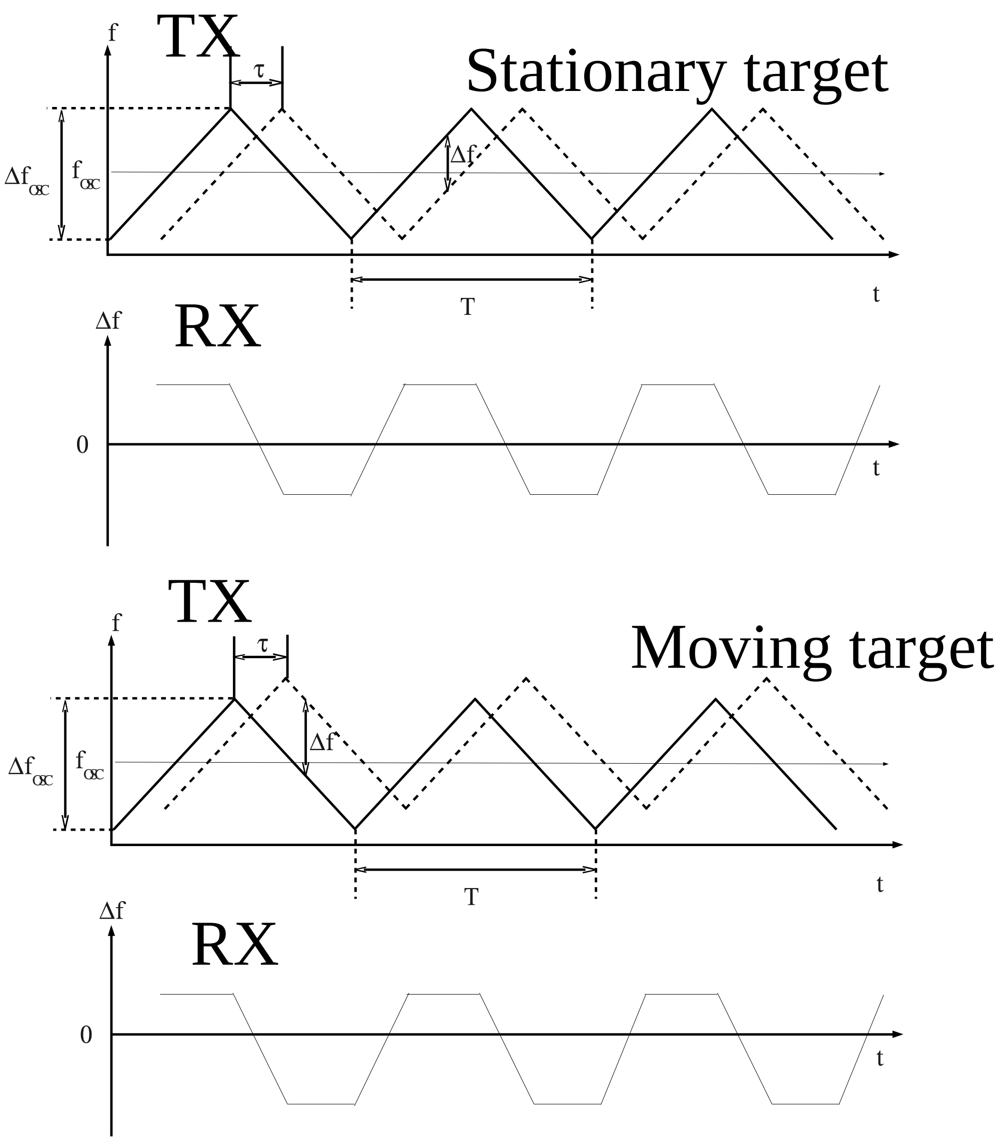


- Link Budget Equation

$$P_{RX} = \frac{\lambda^2 \sigma P_{TX} G_{TX} G_{RX}}{(4\pi)^3 r^4}$$

σ is the radar cross section of the target in m^2.

FMCW Radar Waveforms



- Range

$$r = \frac{1}{2} c \tau$$

$$r = \frac{1}{2} c \tau = \frac{c \Delta f T}{2 \Delta f_{OSC}}$$

- Velocity

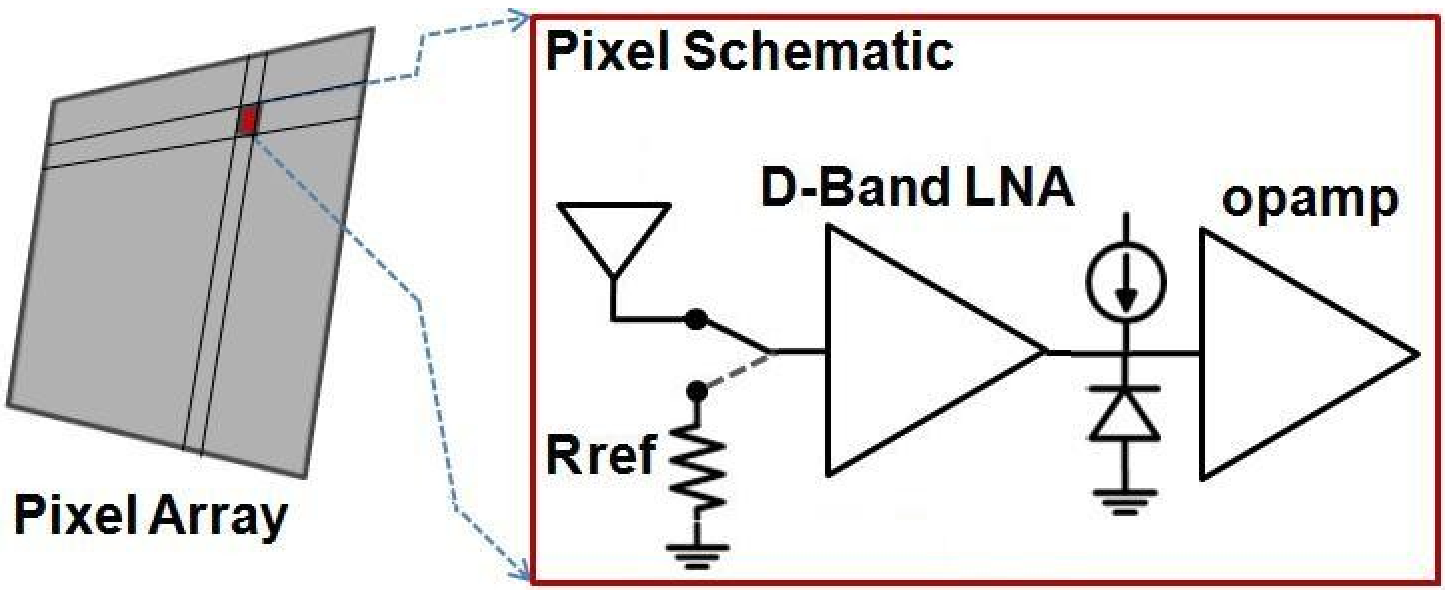
$$v = \frac{c \Delta f_d}{2 f_{OSC}}$$

Δf_d = Doppler shift

c = speed of light

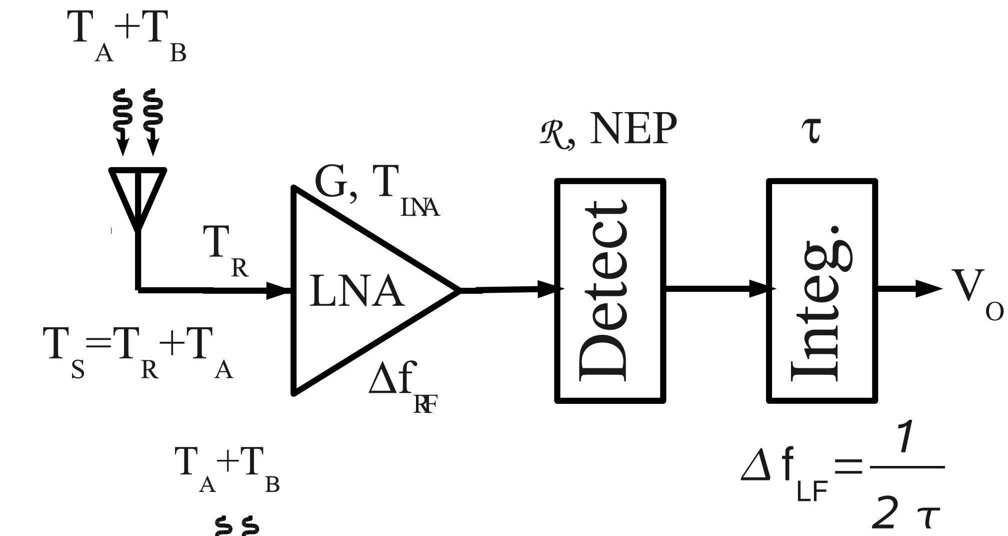
Radioastronomy and remote sensing

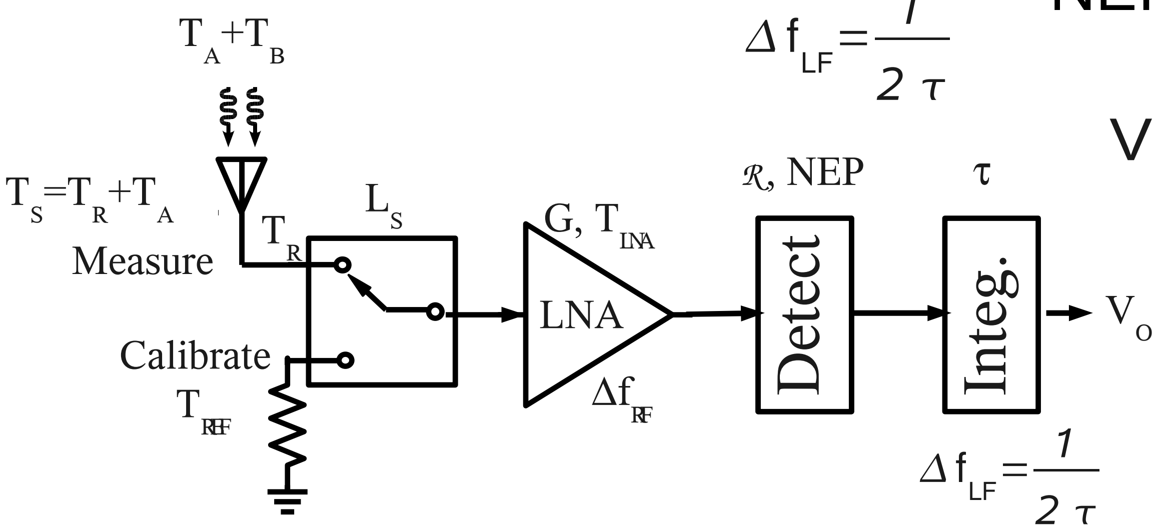
- W-band and D-Band “cameras” (“full body scanners”)
- low-noise, broadband receiver measuring “black-body” radiation emitted by all bodies: $P = k\Delta f T$



Reference: Pozar

Total Power Radiometers



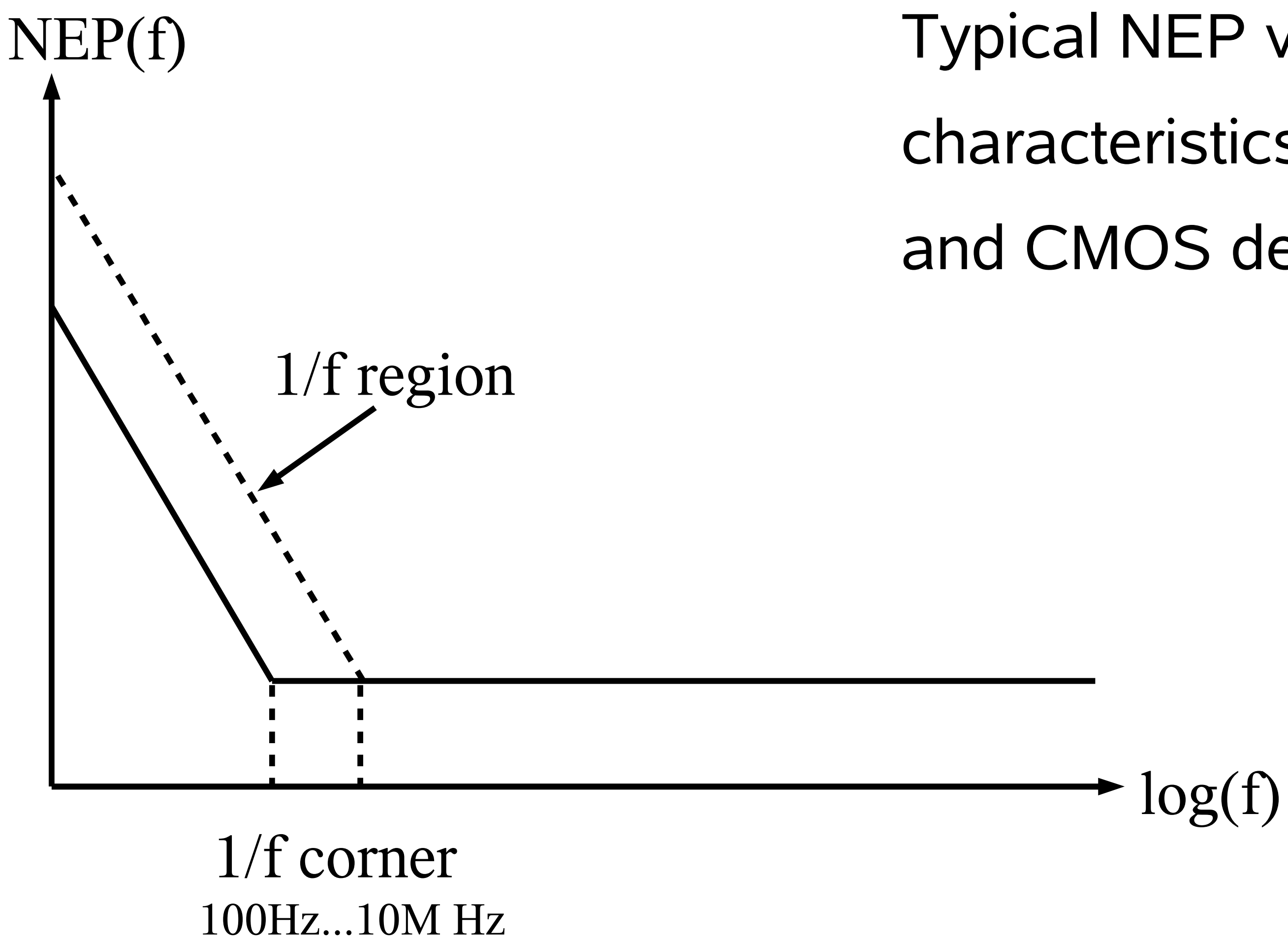


T_A = antenna noise temperature
T_R = receiver noise temperature
T_S = system noise temperature
R = detector responsivity [V/W]
NEP = noise equivalent power

$$V_o=(kT_B\Delta f_{RF}+kT_S\Delta f_{RF})GR$$

$$\Delta T_{MIN}=(2)(T_A+T_R)\sqrt{\frac{1}{\Delta f_{RF}\tau}+\left(\frac{\Delta G}{G}\right)^2}$$

Detector NEP



Typical NEP vs. frequency characteristics for III-V, SiGe and CMOS detectors

Example of Total Radiometer

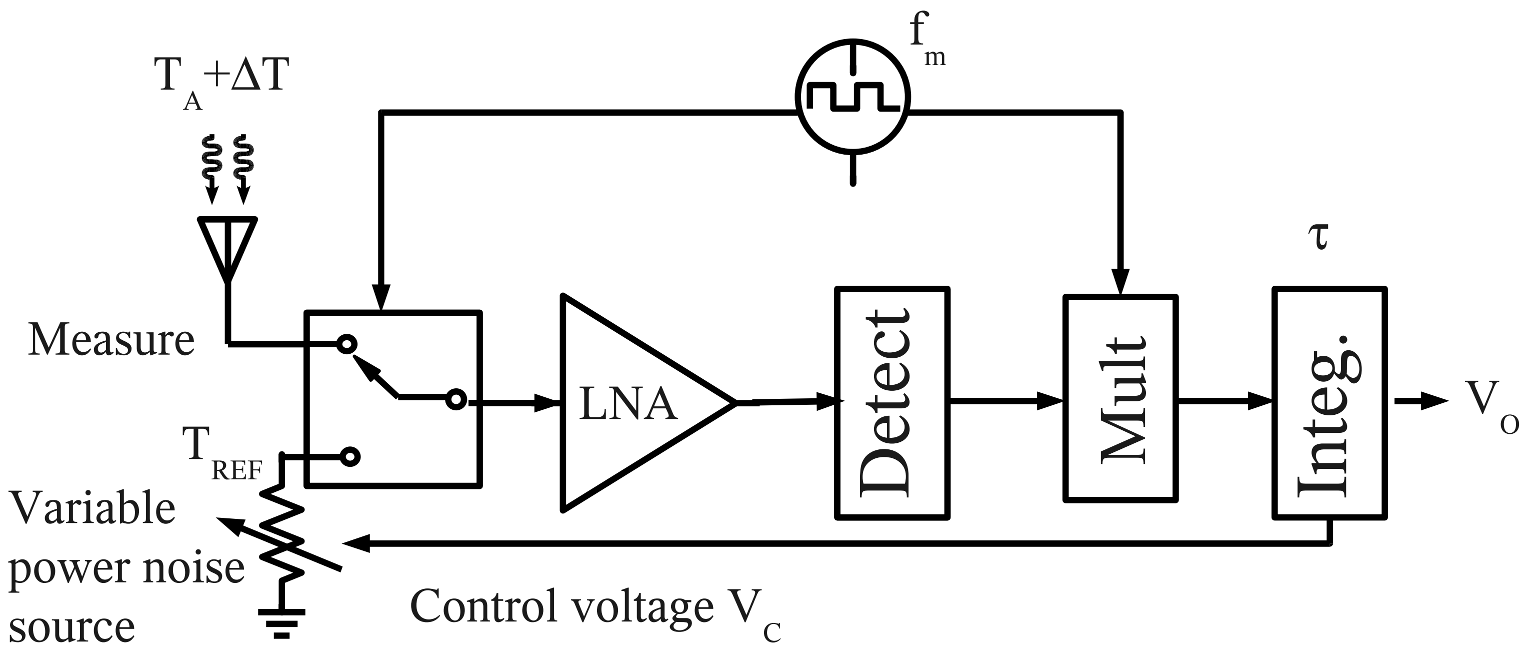
- 94-GHz total power radiometer
- Δf_{RF}=10 GHz, G = 30dB (1000), T_S = 400K and τ = 20 mS,

$$\Delta T_{MIN} = 0.028 \text{ K.}$$

This is an outstanding resolution!

- If the LNA gain fluctuates by 0.05dB, ΔT_{MIN} increases dramatically to 4.62 K,
- For most practical applications, a resolution of at least 0.5 K is considered necessary.

Dicke Radiometer



Reduces impact of 1/f noise and gain fluctuations

Fiber and Backplane Systems

- Fiber
  - Direct modulation ASK (OOK) transmitter
  - Direct detection receiver
  - New systems at 110+ Gb/s are based on QPSK and 16QAM
- Backplane and cable transceivers
  - Unique example of a baseband system
  - No carrier modulation
  - NRZ or 4/8/128-PAM data transmission directly over backplane or cable

Back-up

Amplitude Shift Keying

$$s(t)=m(t)\cos(\omega_0 t) \text{ where } m(t)=0,1$$

- m(t) = data signal
- cos(ωt) = carrier

- Generation:
  - By upconversion: mixing data with carrier
  - By direct modulation: oscillator+ on-off switch
- Demodulation
  - Synchronously (coherent) by downconversion + low-pass filter (LPF)
  - Asynchronously (direct detection): square law or envelope detector +LPF

Amplitude Shift Keying (ii)

Coherent demodulation (better performance, costlier)

$$v(t) = s(t) \times \cos(\omega_0 t) = m(t)\cos^2(\omega_0 t)$$

$$\frac{1}{2} m(t)[1 + \cos(2\omega_0 t)] \rightarrow \frac{m(t)}{2}$$

$$P_e = \frac{1}{2} erfc\left(\sqrt{\frac{E_b}{4 n_0}}\right)$$

$$E_b = \int_{t=0}^{T} s^2(t)\, dt$$

$$P_e = \frac{1}{2} erfc\left(\sqrt{\frac{SNR}{4} \frac{\Delta f}{R_b}}\right)$$

P_e = probability of error in detector

E_b = energy of signal over 1 bit period in Ws

n_0 = noise power spectral density in W/Hz

R_b = data rate

Δf = bandwidth

Amplitude Shift Keying (iii)

Direct detection (1dB worse SNR)

$$v(t) = s^2(t) = m^2(t)\cos^2(\omega_0 t)$$

$$\frac{m^2(t)}{2}\left[1 + \cos(2\omega_0 t)\right] \rightarrow \frac{m^2(t)}{2} = \frac{m(t)}{2}$$

Frequency Shift Keying

- Carrier frequency switches between two values f_1, and f_2, generated with VCO, constant amplitude, no need for linear power amplifier

$$s(t) = cos[(\omega_2 + m(t)\Delta\omega)t]$$

Where

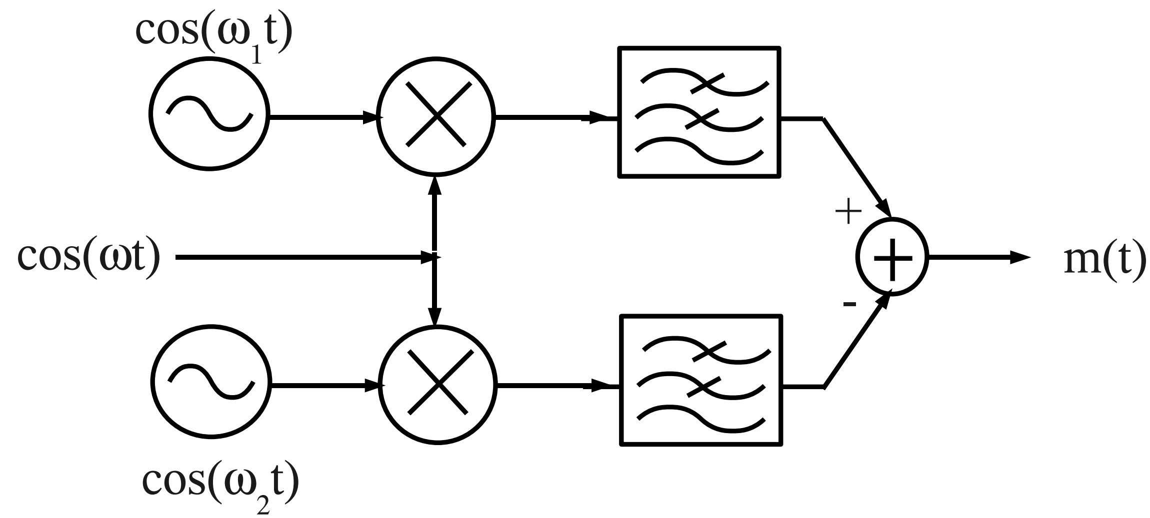
  - m(t) is a binary data signal,
  - $\Delta\omega = 2\pi$ Δf and
  - $\Delta f = f_1 - f_2$.

- Spectrum effective bandwidth

$$B = 2\left(\Delta f + \frac{2}{T}\right)$$

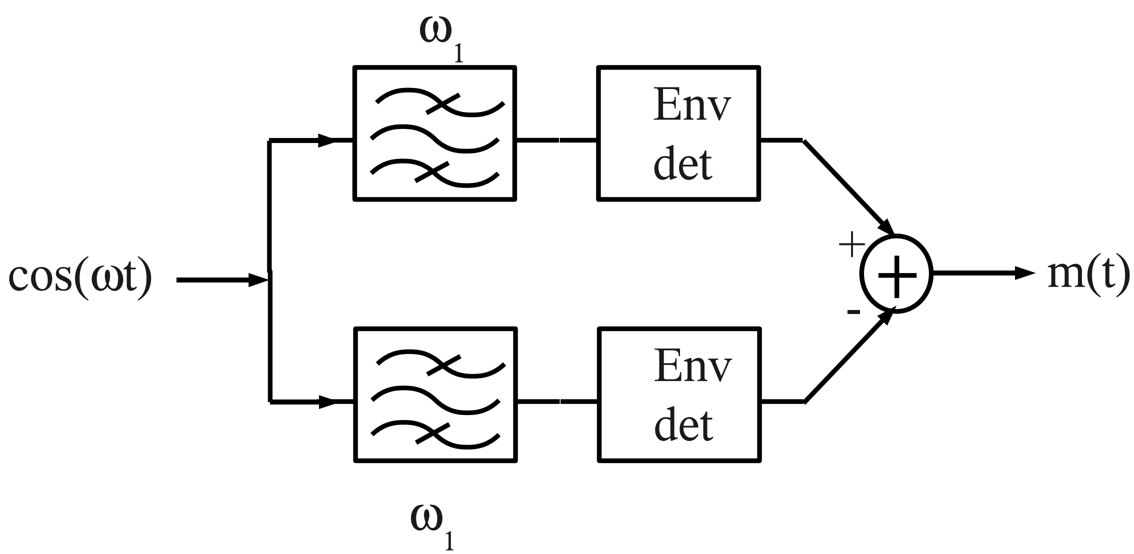
T= period of binary data signal

Frequency Shift Keying: Detector



- Coherent (synchronous)

$$P_e = \frac{1}{2} erfc\left(\sqrt{\frac{E_b}{2n_0}}\right)$$



- Direct detector

Phase Shift Keying

$$s(t) = m(t)\cos(\omega_0 t)$$

Where m(t) is in polar NRZ format, i.e. m(t) =+/-1

- Generation:
  - mixing(upconversion) m(t) with carrier $\cos(\omega_0 t)$
  - Direct phase (sign) modulation of differential VCO signal
- As FSK, has constant amplitude, good with non-linear power amplifiers

Phase Shift Keying: Detection

- Only coherently (as coherent ASK)

$$v(t)=s(t)\times cos(\omega_0 t)=m(t)cos^2(\omega_0 t)$$
$$\frac{1}{2}m(t)[1+cos(2\omega_0 t)]\rightarrow\frac{m(t)}{2}$$

- Difference from ASK: m(t) is polar, hence decision threshold =0, does not depend on amplitude

$$P_e=\frac{1}{2}erfc\left(\sqrt{\frac{E_b}{n_0}}\right)$$

- P_e is 6 dB better than ASK and 3 dB better than FSK, but ASK has 3 dB lower average transmit power