# Elements of communication theory and RF systems

## ICS905 – FARE

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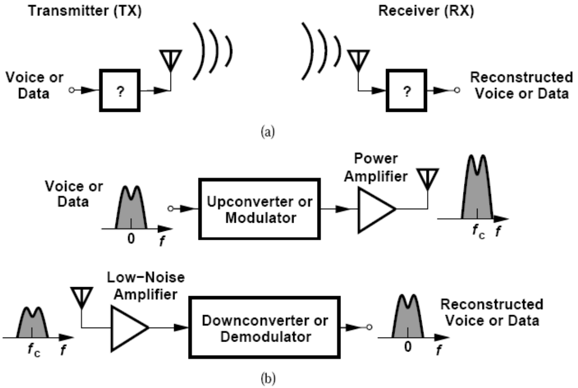
# Outline

# Credits

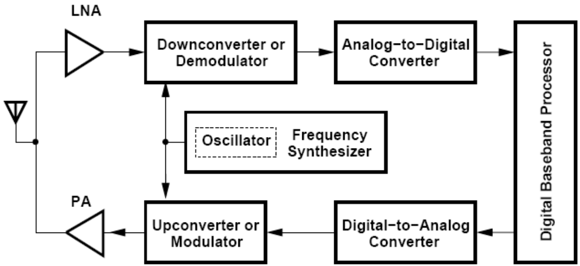
- Title slides have annotations
  - ♣ reuses materials from
    - PowerPoints for RF Microelectronics, Prepared by Bo Wen, UCLA
    - based on : Behzad Razavi, *RF Microelectronics*.
    - Available online : Razavi, PowerPoints for RF Microelectronics | Pearson
  - ♠ reuses materials from
    - RF transceivers short course, Rayan MINA, 2010
  - ♦ reuses materials from
    - Introduction to Digital Modulation, EE4367 Telecom. Switching & Transmission, Prof. Murat Torlak
    - Available online : lecturedm.pdf
  - ▲ reuses materials from
    - Lecture 19 Basics of Wireless Communication, 6.976 High Speed Communication Circuits and Systems, Michael Perrott, MIT
    - Available online : lec19.pdf
  - ▼ reuses materials from
    - BER calculation, Vahid Meghdadi
    - based on *Wireless Communications* by Andrea Goldsmith
    - Available online : ber_awgn.pdf

# The Big Picture: RF Communication (♣)

- TX: Drive antenna with high power level
- RX: Sense small signal (amplify with low noise)

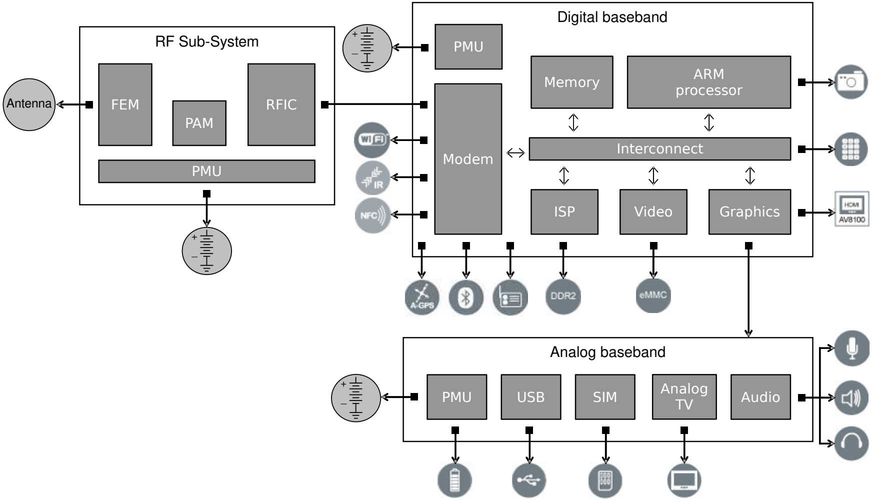


# The Big Picture: Generic RF Transceiver (♣)



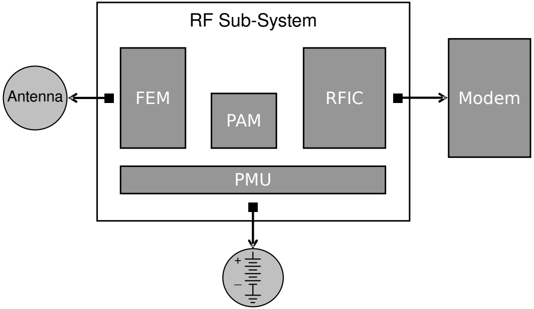

- Signals are upconverted/downconverted at TX/RX, by an oscillator controlled by a Frequency Synthesizer

# Considered system: Mobile platform overview (♠)



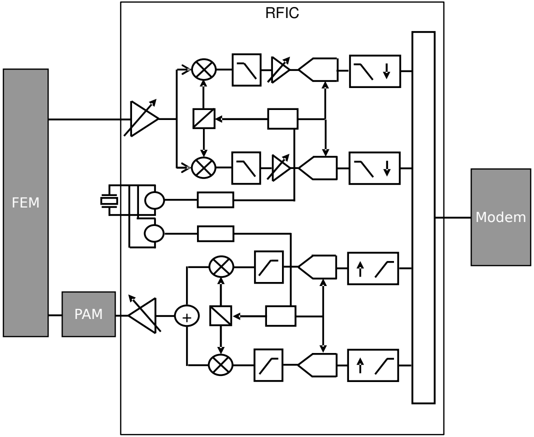
# Considered system: RF Sub-system overview (♠)

- Radio-Frequency Integrated Circuit (RFIC)
  - Receive chain (RX)
  - Transmit chain (TX)
  - Frequency synthesizer (SX)
- Power Amplifier Module (PAM)
  - Power amplifier stages
  - Biasing circuits
  - Harmonic rejection filters
- Front-End Module (FEM)
  - RF switches
  - RF filters, duplexers
  - Diplexers
  - RF coupler
- Power Management Unit (PMU)
  - Low-DropOut voltage regulators (LDO)
  - Charge pumps circuits and DC level shifters



# Considered system: RFIC overview (♠)

- RX
  - Low-noise amplifier (LNA)
  - Down-conversion Mixer
  - Analog baseband filters (ABB)
  - Analog-to-digital converter (ADC)
  - Digital front-end (RxDFE)
- TX
  - Digital front-end (TxDFE)
  - Digital-to-analog converter (DAC)
  - Analog baseband filters (ABB)
  - Up-conversion mixer and IQ modulator
  - Pre-power amplifier (PPA)
- SX
  - Voltage controlled oscillator (VCO)
  - Clock dividers
  - Low-pass filters
  - Buffers
- Digital serial interface (DigRF)



# Outline

# Considered main issues of RF design in this lecture[1]

- Nonlinearity
  - Harmonic Distortion
  - Compression
  - Intermodulation
- Noise
  - Noise Spectrum
  - Device Noise
  - Noise in Circuits

[1] Please see (♣) for exhaustive RF design issues (Ex :Impedance Transformation)

# Effects of Nonlinearity: Harmonic Distortion (quick recall) (♣)

$$y(t) = \alpha_1 A \cos\omega t + \alpha_2 A^2 \cos^2 \omega t + \alpha A^3 \cos^3 \omega t \tag{1}$$

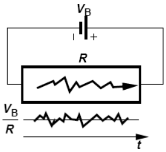
$$= \alpha_1 A \cos\omega t + \frac{\alpha_2 A^2}{2}(1 + \cos 2\omega t) + \frac{\alpha_3 A^3}{4}(3\cos\omega t + \cos 3\omega t) \tag{2}$$

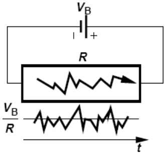
$$= \frac{\alpha_2 A^2}{2} + \left(\alpha_1 A + \frac{3\alpha_3 A^3}{4}\right)\cos\omega t + \frac{\alpha_2 A^2}{2}\cos(2\omega t) + \frac{\alpha_3 A^3}{4}\cos 3\omega t \tag{3}$$

- DC
- Fundamental
- Second harmonic
- Third harmonic
- Even-order harmonics result from $\alpha_j$ with even $j$
- $n^{\text{th}}$ harmonic grows in proportion to $A_n$

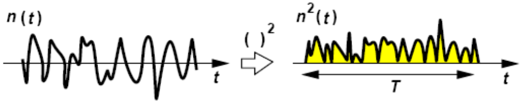
Please have a look at appendix slides (3,4) for circuits examples

# Noise: Noise as a Random Process (♣)




Higher temperature

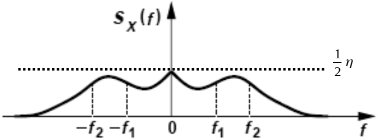



$$P_n = \lim_{T\to\infty} \frac{1}{T} \int_0^T n^2(t)\,\mathrm{d}t \qquad (4)$$

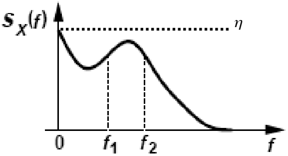
- The average current remains equal to $\frac{V_B}{R}$ but the instantaneous current displays random values
- $T$ must be long enough to accommodate several cycles of the lowest frequency.

# Noise Spectrum: Power Spectral Density (PSD) (♣)

- Two-Sided




- One-Sided




$$\int_{\infty}^{-\infty} S_x^{2-sided}(f)\,\mathrm{d}f = \lim_{T\to\infty} \frac{1}{T}\int_0^T x^2(t)\,\mathrm{d}t \qquad \text{(Parseval)} \qquad (5)$$

- Total area under $S_x(f)$ represents the average power carried by $x(t)$

# Noise Spectrum: Power Spectral Density (PSD) (♣)

## Example of Noise Spectrum

- A resistor of value $R_1$ generates a noise voltage whose *one-sided* PSD is given by

$$S_v(f) = 4kTR_1 \quad [\mathrm{V}^2/\mathrm{Hz}] \qquad (6)$$

- where $k = 1.38 \times 10^{-23}$ J/K denotes the Boltzmann constant and $T$ the absolute temperature. Such a flat PSD is called *white* because, like white light, it contains all frequencies with equal power levels.

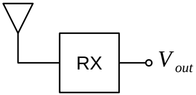
1. The total average power carried by the noise voltage is the area under $S_v(f)$ and it appears to be infinite; an implausible result because the resistor noise arises from the finite ambient heat. In reality, $S_v(f)$ begins to fall at $f > 1$ THz, exhibiting a finite total energy, i.e., thermal noise is not exactly white.
2. The dimension of $S_v(f)$ is voltage squared per unit bandwidth $(\mathrm{V}^2/\mathrm{Hz})$
3. The noise voltage for a 50-Ω resistor in 1 Hz at room temperature ($T = 300$ K) is:

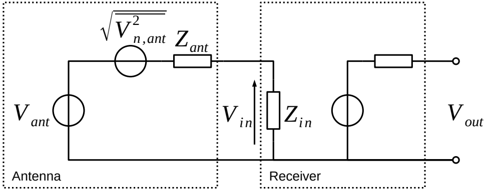
$$\overline{V_n^2} = 8.28 \times 10^{-19}\ \mathrm{V}^2/\mathrm{Hz} \qquad (7)$$

$$\sqrt{\overline{V_n^2}} = 0.91\ \mathrm{nV}/\sqrt{\mathrm{Hz}} \qquad (8)$$

Note: the *two-sided* PSD is: $S_v(f) = 2kTR_1 \quad [\mathrm{V}^2/\mathrm{Hz}]$

# Noise power in receiver load




$$\sqrt{\overline{V_{n,in}^2}} = \frac{Z_{in}}{Z_{in}+Z_{ant}}\sqrt{\overline{V_{n,ant}^2}} \quad [\mathrm{V}/\sqrt{\mathrm{Hz}}] \quad (9)$$

$$\sqrt{\overline{V_{n,in}^2}} \underset{matched}{=} \frac{1}{2}\sqrt{\overline{V_{n,ant}^2}} \quad [\mathrm{V}/\sqrt{\mathrm{Hz}}] \quad (10)$$

$$\overline{V_{n,in}^2} \underset{matched}{=} kTZ_{ant} \quad [\mathrm{V}^2/\mathrm{Hz}] \quad (11)$$

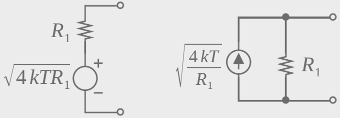
$$P_{n,in} \underset{matched}{=} \frac{\overline{V_{n,in}^2}}{Z_{ant}} = kT \quad [\mathrm{W}/\mathrm{Hz}] \quad (12)$$

**Integrated noise power in the receiver load**

$$P_{n,in,tot} = kTB \quad [\mathrm{W}] \quad (13)$$

# Effect of Transfer Function on Noise/Device Noise (♣)


- Define PSD to allow many of the frequency-domain operations used with deterministic signals to be applied to random signals as well.




- Noise can be modeled by a series voltage source or a parallel current source
- Polarity of the sources is unimportant but must be kept same throughout the calculations

# Noise Figure : Definition (♣)

$$NF = \frac{SNR_{in}}{SNR_{out}} \tag{14}$$

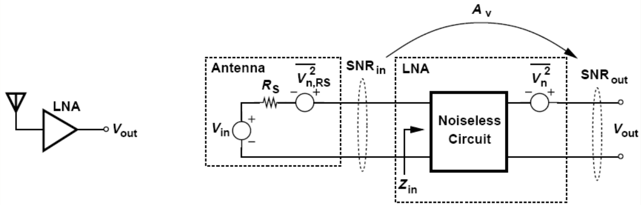
$$NF_{|\mathrm{dB}} = 10\log_{10}\left(\frac{SNR_{in}}{SNR_{out}}\right) \tag{15}$$

- Depends on not only the noise of the circuit under consideration but the SNR provided by the preceding stage
- If the input signal contains no noise, $NF = \infty$

# Noise Figure : Exercise (♣)

## Self-exercise

- Calculation of Noise Figure




- Solution

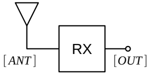
$$\alpha = \frac{Z_{in}}{Z_{in} + R_S} \tag{16}$$

$$NF = 1 + \frac{\overline{V_n^2}}{|\alpha|^2 A_v^2} \cdot \frac{1}{\overline{V_{RS}^2}} \tag{17}$$

- NF must be specified with respect to a source impedance-typically 50 Ω
- What is the word used to says that $Z_{in} = R_S$?

# Sensitivity (♣)

- The **sensitivity** is defined as the minimum signal level that a receiver can detect with **acceptable quality**(†).




$$SNR_{ant} = \frac{P_{sig,ant}}{P_{n,ant}} \quad (18) \qquad SNR_{out} = \frac{P_{sig,out}}{P_{n,out}} \quad (19) \qquad NF = \frac{SNR_{in}}{SNR_{out}} \quad (20)$$

$$P_{sig,ant} = SNR_{ant} \cdot P_{n,ant} = SNR_{out} \cdot NF \cdot \overbrace{P_{n,ant}}^{kTB} \quad (21)$$

$$P_{sen} := P_{sig,ant,min} = SNR_{out,min} \cdot NF \cdot k \cdot T \cdot B \quad (\ddagger) \quad (22)$$

$$P_{sen|\mathrm{dBm}} = SNR_{out,min|\mathrm{dB}} + NF_{\mathrm{dB}} + 10\log_{10}(kT) + 10\log 10(B) - 10\log_{10}(1\,\mathrm{mW}) \quad (23)$$

$$= SNR_{out,min|\mathrm{dB}} + NF_{\mathrm{dB}} + 10\log 10(B) \underbrace{-174}_{\mathrm{dBm/Hz}} \quad (24)$$

† : Arbitrary! ; ‡ : $T = 290$ K

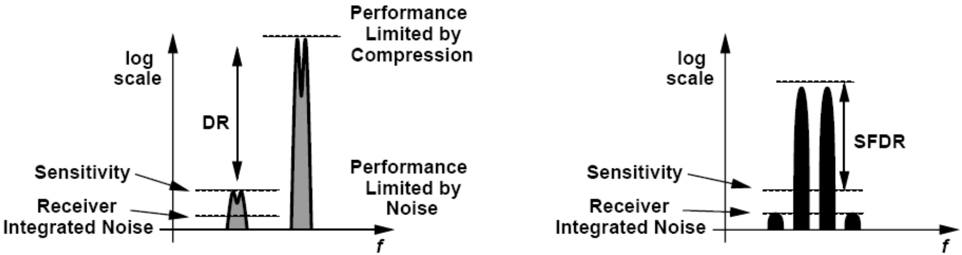
# Sensitivity (♣)

## Self-exercise

- A GSM receiver requires a minimum SNR of 12 dB and has a channel bandwidth of 200 kHz. A wireless LAN receiver, on the other hand, specifies a minimum SNR of 23 dB and has a channel bandwidth of 20 MHz.
  - Compute the sensitivities of these two systems if both have an NF of 7 dB.

- Solution:
  - For the GSM receiver, $P_{sen} = -102\,\mathbf{dBm}$, whereas for the wireless LAN system, $P_{sen} = -71\,\mathbf{dBm}$.
    Does this mean that the latter is inferior? No, the latter employs a much wider bandwidth and a more efficient modulation to accommodate a data rate of $54\,\mathbf{Mb/s}$. The GSM system handles a data rate of only $270\,\mathbf{kb/s}$. In other words, specifying the sensitivity of a receiver without the data rate is not really meaningful.

# Dynamic Range (♣)



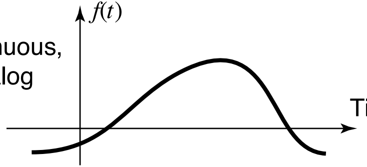

## Dynamic Range

Maximum tolerable desired signal power divided by the minimum tolerable desired signal power

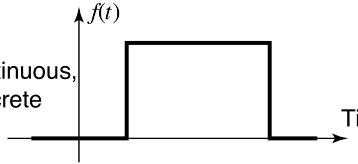
# Outline

# Nature of the message

- Analogue
  - Ex: Voice, music, position,...




- Digital
  - Ex : piano notes, text (consisting of letters), RVB pixels



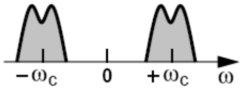

- The information is given by the variation of a physical parameter = MODULATION
  - Light, mecanical, electromagnetic waves

# Journey of the Signal (♣)

Baseband modulation

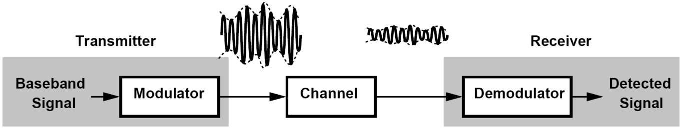

$$x(t) = a(t) \quad (25)$$

Passband modulation




$$x(t) = a(t)\cos(\omega_c t + \theta(t)) \quad (26)$$

- For bandpass modulation : several parameters of a sinusoidal carrier can be modulated according to the baseband signal.

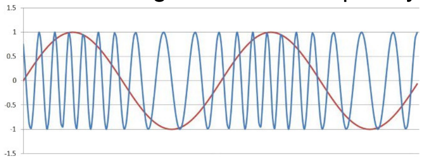



- A simple communication system consists of a modulator/transmitter, a channel, and a receiver/demodulator

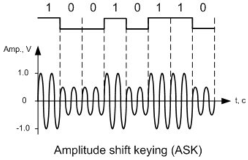
# Usual physical modulated parameters

- Analogue modulation
  - Amplitude

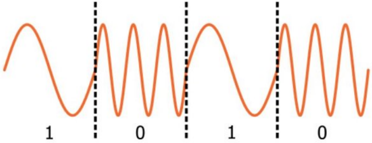
  - Angle : Phase/Frequency




- Digital modulation
  - Amplitude




  - Angle : Phase/Frequency

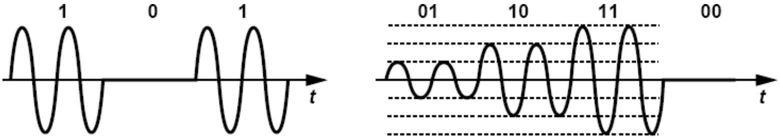


# Modulation keywords summary

- Baseband
- Analog
- Amplitude

- Bandpass or Modulated carrier
- Digital
- Frequency / Phase

# Important aspects of modulations (♣)

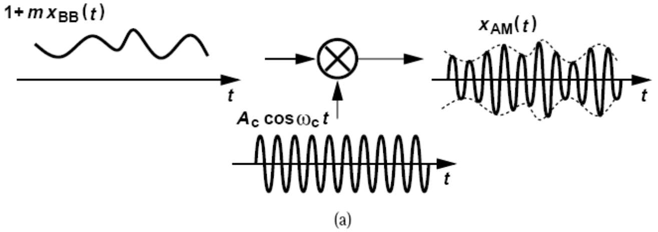


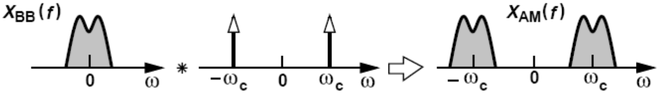

2-level    4-level

- Detectability
  - the quality of the demodulated signal for a given amount of channel attenuation and receiver noise
- Bandwidth Efficiency
  - the bandwidth occupied by the modulated carrier for a given information rate in the baseband signal
- Power Efficiency
  - the type of power amplifier (PA) that can be used in the transmitter

# Outline

# Analog Modulation: 1st case : Amplitude Modulation




$$x_{AM}(t) = A_c\,[1 + m x_{BB}(t)] \cos\omega_c t \tag{27}$$

- $m$ is called the *modulation index*

# Amplitude Modulation – Spectrum (♣)

$$p(t) = a(t) \cdot \cos(\omega_0 t) \qquad (28)$$

$$\overset{\circ}{p}(f) = \left\{ \overset{\circ}{a} * \frac{1}{2} [\delta_{f_0} + \delta_{-f_0}] \right\}(f) \qquad (29)$$

$$\overset{\circ}{p}(f) = \frac{1}{2} \left[ \overset{\circ}{a}(f - f_0) + \overset{\circ}{a}(f + f_0) \right] \qquad (30)$$

1. $a(t)$ being real-valued, its TF has Hermitian symmetry

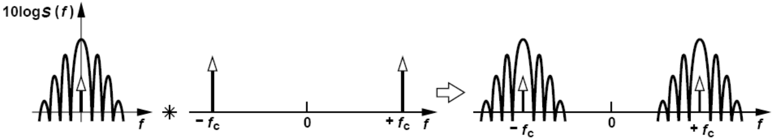
   $$\overset{\circ}{a}(-f) = \overset{\circ}{a}^{*}(f) \qquad (31)$$

   The spectrum of the carrier $p(t)$ is therefore symmetrical around the frequencies $\pm f_0$, with two bands lateral: an upper lateral band (ULB) and a lower lateral band (LLB).
2. A spectral ray at the carrier frequency is present if there is a DC component in $a(t)$ (embedded carrier).

# Spectrum of (Binary) Amplitude Modulation (♣)

- Consider the product of a random binary sequence, toggling between 0 and 1, and a sinusoidal carrier
- The spectrum of a random binary sequence with equal probabilities of ONEs and ZEROs is given by

$$S(f) = T_b \left( \frac{\sin \pi f T_b}{\pi f T_b} \right)^2 + \frac{1}{2} \delta(f) \tag{32}$$




- Multiplication by a sinusoid in the time domain shifts this spectrum to a center frequency of $\pm f_c$

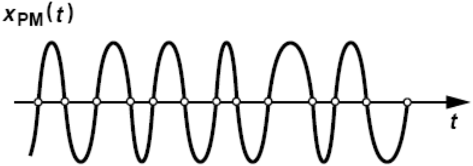
# Analog Modulation: 2nd case : Phase & Frequency Modulation (angle modulation) (♣)

- Phase Modulation
  - Amplitude is constant and the excess phase is linearly proportional to the baseband signal

$$x_{PM}(t) = A_c \cos\left[\omega_c t + m x_{BB}(t)\right] \quad (33)$$

- Frequency Modulation
  - the excess frequency is linearly proportional to the baseband signal

$$x_{FM}(t) = A_c \cos\left[\omega_c t + m \int_{-\infty}^{t} x_{BB}(\tau)\, d\tau\right] \quad (34)$$



# Spectrum of angle modulation

- Much more complicated than AM

$$p(t) = A_0 \cos\left[\omega_0 t + 2\pi f \int_0^t s_u(\theta)\, d\theta\right] \quad (35)$$

- Closed form only possible for special cases
  - Sinus modulation
  - Small amplitude angle modulation
  - FSK

# Spectrum of angle modulation : Sinus modulated case

- Sinus modulation case

$$p(t) = A_0 \cos\left[\omega_0 t + \beta \sin(\Omega t)\right] \tag{36}$$

We use the associated complex exponential carrier:

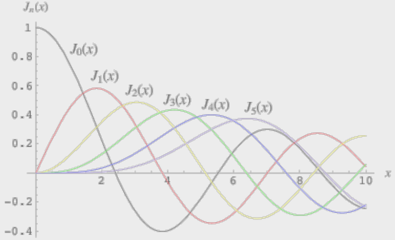
$$p(t) = \Re\left\{A_0 e^{j[\omega_0 t + \beta \sin(\Omega t)]}\right\} \tag{37}$$

$$p(t) = \Re\left[A_0 e^{j\omega_0 t} \sum_{n \in \mathbb{Z}} J_n(\beta) e^{jn\Omega t}\right] \tag{38}$$

with the use of the Fourier series expansion of $e^{j\beta \sin(\Omega t)}$ ; $J_n(\beta)$ is the Bessel functions of the first kind of order $n$. The modulated carrier $p(t)$ can be written:

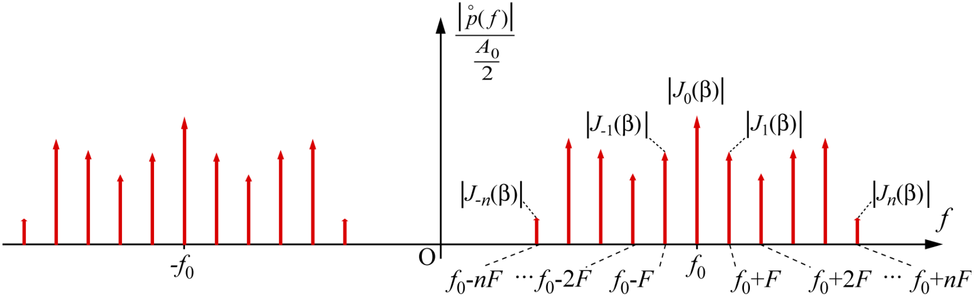
$$p(t) = A_0 \sum_{n \in \mathbb{Z}} J_n(\beta) \cos\left[(\omega_0 + n\Omega) t\right] \tag{39}$$

Bessel Function of the First Kind



# Spectrum of angle modulation : Sinus modulated case

$$p(t) = A_0 \sum_{n\in\mathbb{Z}} J_n(\beta) \cos\left[(\omega_0 + n\Omega)\, t\right] \tag{39}$$



# Outline

# Outline

Communication theory : Digital modulations

- Baseband
- Passband
- Cellular communication considerations (mobile systems and multiple access)
- Hardware considerations
- Wireless standards
- Transceivers architectures

# Outline

# Analog modulation vs Digital modulation ? (Definition)

## Definition

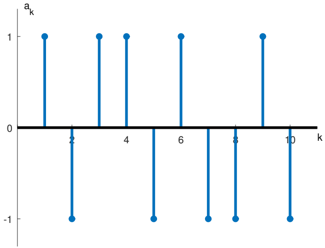
- Analog modulation
  - The modulating signal is analog (continuous in time and amplitude)
- Digital modulation
  - The modulating signal is digital (discrete in time and amplitude)

## Note

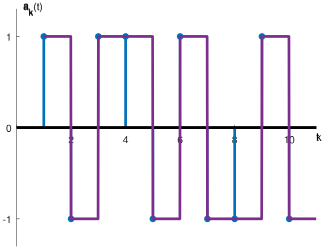
Though "smoothed" (=filtered) digital signals tend to be analog signals, the carried information remains coded on discrete states, hence the word "digital" is kept for these signals.

# Elementary binary transmission

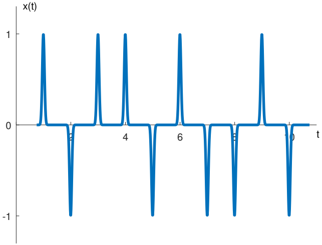
- Ideal binary pulses




- Binary square pulses
  - (baseband) ASK




- Particular shaped pulses
  - (baseband) "Filtered" ASK



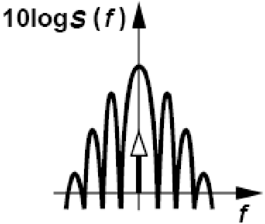
# Outline

# Impairments : Channel BW limitation a.k.a. ISI

## Spectrum of (baseband) squared ASK (♣)

- The spectrum of a random binary sequence with equal probabilities of ONEs and ZEROs is given by

$$S(f) = T_b \left( \frac{\sin \pi f T_b}{\pi f T_b} \right)^2 + \frac{1}{2}\delta(f) \tag{32}$$



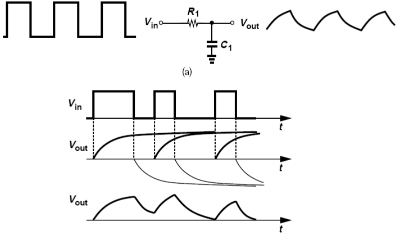

Infinite bandwidth !

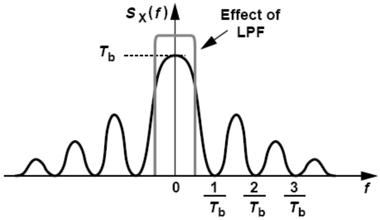
# Impairments : Channel BW limitation a.k.a. ISI

## Intersymbol Interference : definition (♣)

### Channel bandwidth limitation

But channel and electronics are bandlimited systems



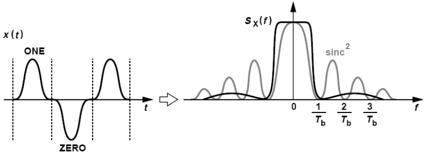

- Each bit level is corrupted by decaying tails created by previous bits.
  - **Intersymbol Interference**
- Low pass filtering creates ISI (except. . . see next slide)

# Impairments : Channel BW limitation a.k.a. ISI

## Pulse Shaping (♣)

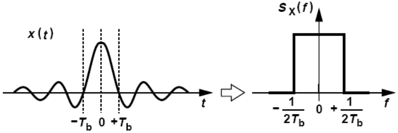
- Special pulse shapes (a.k.a filters) are designed to minimise ISI
  - They reduce the initial bandwidth

### Ex : Sinc

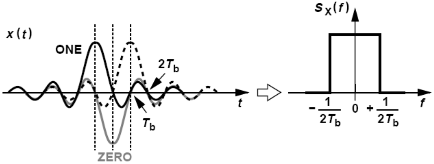



- Baseband pulse is designed to occupy a limited bandwidth.

### Single pulse



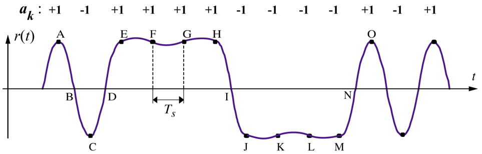

### Random pulse sequence

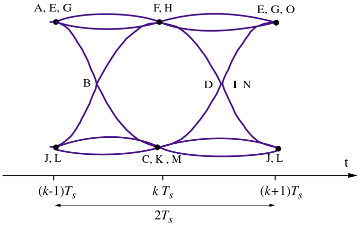


# Impairments : Channel BW limitation a.k.a. ISI

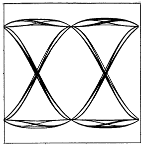
## ISI characterization : eye diagram

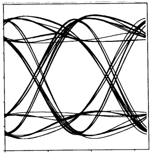
Empirical measure of the quality of the received digital baseband signal



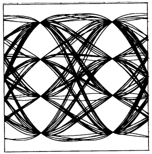

- Binary modulation





No ISI With ISI

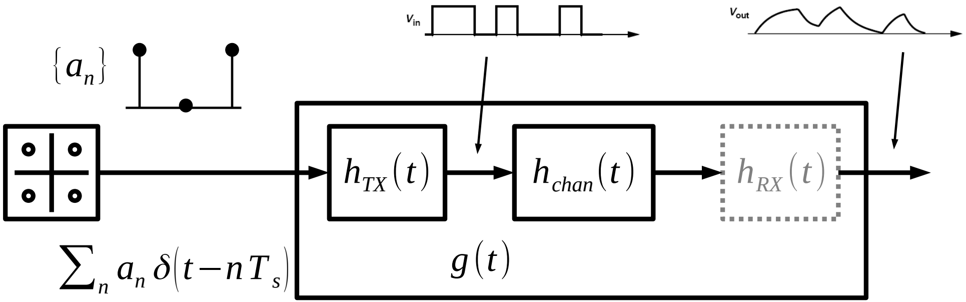
- M-ary modulation



The wider the "eye" opens, the better the signal quality is.

# Impairments : Channel BW limitation a.k.a. ISI

## Nyquist criterion for ISI Cancellation (time domain)




- Effects of ISI could be completely nullified if, at every sampling instant, the response due to all symbols except the current symbol is made equal to zero

# Impairments : Channel BW limitation a.k.a. ISI

## Receiver input and ISI cancellation formula

Output of the overall channel:

$$y(t) = \sum_{n\in\mathbb{Z}} a_n g(t - nT_s) \quad (40)$$

$$= a_k g(t - kT_s) + \sum_{n\in x, n\neq k} a_n g(t - nT_s) \quad (41)$$

Signal value at the sampling instant $kT_s$:

$$y(kT_s) = a_k g(0) + \sum_{n\in\mathbb{Z}, n\neq k} a_n g[(k-n)T_s] \quad (42)$$

- $a_k g(0)$: magnitude of the expected pulse
- $\sum_{n\in\mathbb{Z}, n\neq k} a_n g[(k-n)T_s]$ perturbation term due to ISI

Let's find the condition to cancel the ISI term:

$$\forall a_n \sum_{n\in\mathbb{Z}, n\neq k} a_n g[(k-n)T_s] = 0$$

$$\Rightarrow \forall a_{k-p} \sum_{p\in\mathbb{Z}^*} a_{k-p} g(pT_s) = 0 \quad (43)$$

### ISI cancellation

We deduce a *necessary and sufficient condition* for canceling the ISI:

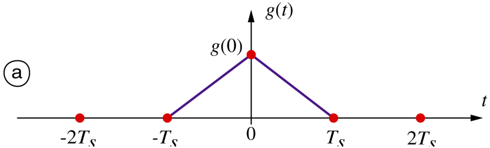
$$\boxed{\begin{cases} \forall p \neq 0, g(pT_s) = 0 \\ g(0) \neq 0 \end{cases}} \quad (44)$$

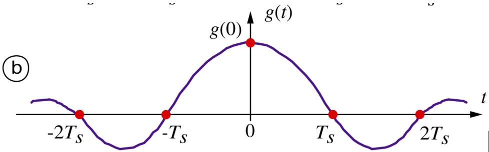
# Impairments : Channel BW limitation a.k.a. ISI

## Nyquist criterion for ISI Cancellation (time domain)

$$\begin{cases} \forall p \neq 0, g\left(pT_s\right) = 0 \\ g\left(0\right) \neq 0 \end{cases} \quad (44) \quad \Leftrightarrow \quad g(t) \cdot \text{Ш}_{T_s} = g(0) \cdot \delta(t) \quad (45)$$

### Examples



# Impairments : Channel BW limitation a.k.a. ISI

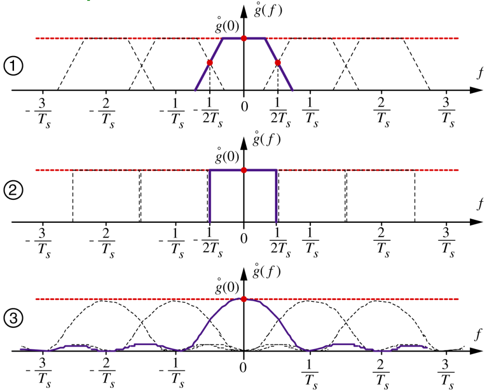
## Nyquist criterion for ISI Cancellation (frequency domain)

Applying Fourier transform on Eq. (45) yields:

**Nyquist criterion for ISI Cancellation (freq domain)**

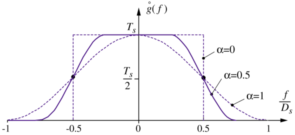
$$\sum_{k\in\mathbb{Z}} \mathring{g}\left(f - \frac{k}{T_s}\right) = T_s \cdot g(0) \qquad (46)$$

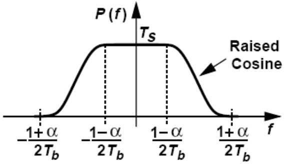
### Examples



## Practical pulse shaping filter : Raised Cosine Filter

- Frequency response




$$\overset{\circ}{g}(f) = \begin{cases} \frac{T_s}{2}\left(1-\sin\left[\frac{\pi}{2\alpha}\left(2\left|f\right|T_s-1\right)\right]\right), & (1-\alpha)\frac{D_s}{2} \leq |f| \leq (1+\alpha)\frac{D_s}{2} \\ T_s, & 0 \leq |f| \leq (1-\alpha)\frac{D_s}{2} \\ 0, & \text{elsewhere} \end{cases} \tag{47}$$

- $\alpha$: roll-off factor, typical values are in the range of $0.3 \sim 0.5$
- The 3dB bandwidth is the same ! The smaller $\alpha$, the steeper the response.

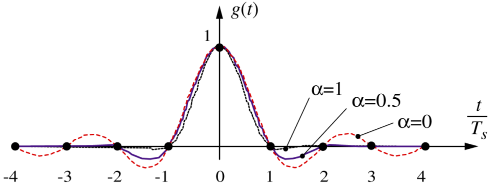
- Theoretical frequency support (*Null-to-null* **total bandwidth**) :

$$BW_{RaisedCos} = \frac{1+\alpha}{T_s} \tag{48}$$

# Impairments : Channel BW limitation a.k.a. ISI

## Practical pulse shaping filter : Raised Cosine Filter

- Time domain response



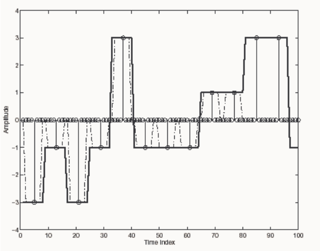

$$g(t) = \frac{\sin\left(\frac{\pi t}{T_s}\right)}{\frac{\pi t}{T_s}} \times \frac{\cos\left(\frac{\pi \alpha t}{T_s}\right)}{1 - \left(\frac{2\alpha t}{T_s}\right)^2} \tag{49}$$

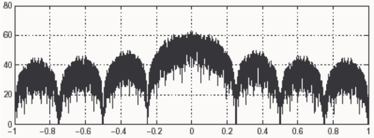
- Instants cancellation are the same !
- The smaller $\alpha$, the higher the overshoot

# Impairments : Channel BW limitation a.k.a. ISI

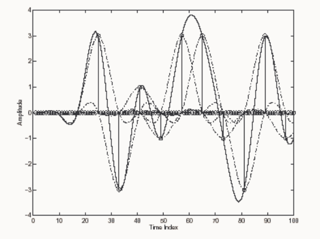
## Signal spectrum after pulse shaping (♦)

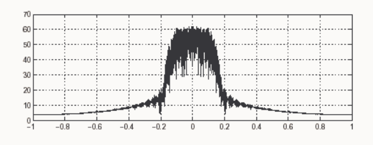
- Before pulse-shaping



- After pulse-shaping

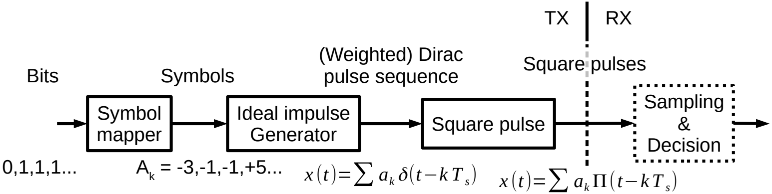


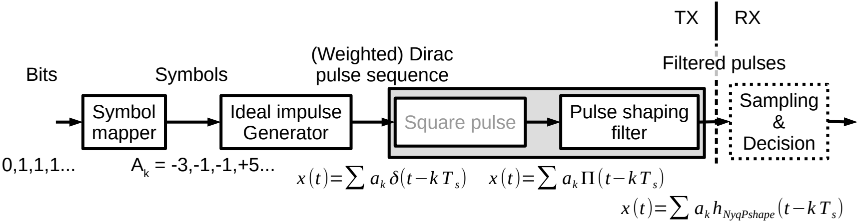
# Impairments : Channel BW limitation a.k.a. ISI

## Conclusion : Transmitter model

- Naive transmitter model



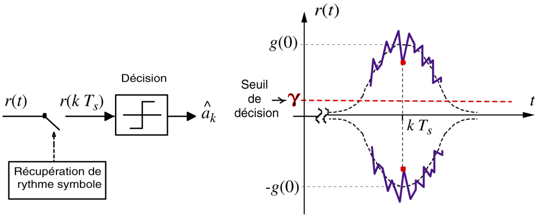

- Now, the transmitter exhibits a **pulse shaping filter**



# Outline

# Impairments : Channel Noise

## Error probability (for binary symbols)




- **Symbol probability error :**

$$P_s\,(e) \triangleq P(\hat{a}_k \neq a_k) \qquad (50)$$

  - Binary symbols : $P_B(e) = P_S(e)$

- **Hypothesis**
  - random sequence of symbols $a_k(\omega)$ *independent and stationary with* $k$,
  - no ISI (*Nyquist channel*),
  - additive noise $b(t, \omega) = B$ random stationary centered, independent of the signal, with probability density $f_B(b)$.

$\omega$ denotes here a sample in the sense of the probability theory

# Impairments : Channel Noise

## Error probability (for binary symbols)

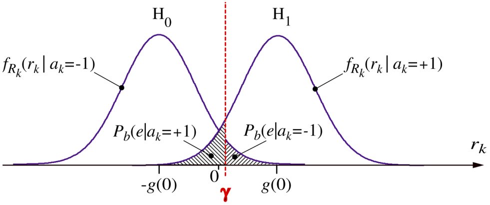
The value sampled at times multiple of $T_s$ is doubly random:

$$r(kT_s, \omega) = R_k = a_k(\omega) \cdot g(0) + b(kT_s, \omega) \tag{51}$$

- random signal term : $a_k(\omega) \cdot g(0)$, with $a_k(\omega) \sim \{\alpha_1, \alpha_2\} = \{-1, +1\}$
- random noise term : $b(kT_s, \omega) = B_k$

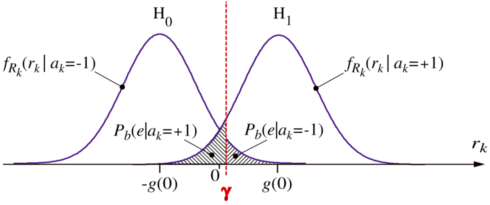
- Conditionnal probability of $R_k$

$$f_{R_k}(r_k | a_k = \alpha_i) = f_B(r_k - \alpha_i \cdot g(0)) \tag{52}$$



# Impairments : Channel Noise

## Error probability (for binary symbols)




- The decision rule is as follows:

$$\begin{cases} \text{if } r \geq \gamma, & \text{Symbol 1 was sent} \\ \text{if } r < \gamma, & \text{Symbol 0 was sent} \end{cases} \tag{53}$$

- Fundamental question : what is the **best value** of $\gamma$ ?

- The best value of $\gamma$ is the one that *minimizes the probability of error*.
- Let's compute the error probability (law of total probability):

$$\begin{aligned} P_b\left(e\right) &= P_b(e, a_k = -1) + P_b(e, a_k = +1) && (54) \\ &= P\left(-1\right) \cdot P_b(e|a_k = -1) + P\left(+1\right) \cdot P_b(e|a_k = +1) && (55) \\ &= P\left(-1\right) \int_{\gamma}^{+\infty} f_{R_k}(r_k|a_k = -1)\,\mathrm{d}r_k + P\left(+1\right) \int_{-\infty}^{\gamma} f_{R_k}(r_k|a_k = +1)\,\mathrm{d}r_k && (56) \end{aligned}$$

# Impairments : Channel Noise

## Error probability (for binary symbols)

- The minimum of the error probability is obtained when:

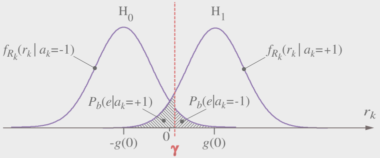
$$\begin{cases} \frac{\mathrm{d}P_b(e)}{\mathrm{d}\gamma} & = -P\left(-1\right) \cdot f_{R_k}(\gamma | a_k = -1) + P\left(+1\right) \cdot f_{R_k}(\gamma | a_k = +1) = 0 \\ \frac{\mathrm{d}^2 P_b(e)}{\mathrm{d}\gamma^2} & = -P\left(-1\right) \left. \frac{\mathrm{d}f_{R_k}(r_k | a_k = -1)}{\mathrm{d}r_k} \right|_{r_k = \gamma} + P\left(+1\right) \left. \frac{\mathrm{d}f_{R_k}(r_k | a_k = +1)}{\mathrm{d}r_k} \right|_{r_k = \gamma} > 0 \end{cases} \tag{57}$$

- Which yields the following condition for the optimal threshold $\gamma_0$:

$$\frac{f_{R_k}(\gamma_0 | a_k = +1)}{f_{R_k}(\gamma_0 | a_k = -1)} = \frac{P(-1)}{P(+1)} \tag{58}$$

# Impairments : Channel Noise
## Error probability (for binary symbols)

- Recall the channel is AWGN




$$f_{R_k}(r_k|a_k = \alpha_i) = \frac{1}{\sigma_b\sqrt{2\pi}} e^{-\frac{(r_k - \alpha_i g(0))^2}{2\sigma_b^2}} \quad (59)$$

- So:

$$\frac{f_{R_k}(\gamma_0|a_k = +1)}{f_{R_k}(\gamma_0|a_k = -1)} = \frac{P(-1)}{P(+1)} = e^{\frac{2\gamma_0 g(0)}{\sigma_b^2}} \quad (60)$$

- So the optimal threshold $\gamma_0$ is :

$$\gamma_0 = \frac{\sigma_b^2}{2g(0)} \ln \frac{P(-1)}{P(+1)} \quad (61)$$

  - The optimal decision threshold tends to move away from the most probable symbol.

# Impairments : Channel Noise

## Error probability (for equiprobable binary symbols)

- For equiprobable symbols: $P(-1) = P(+1) = 1/2$, then $\gamma_0 = 0$.
  - Reuse Eq. (55) and (56):

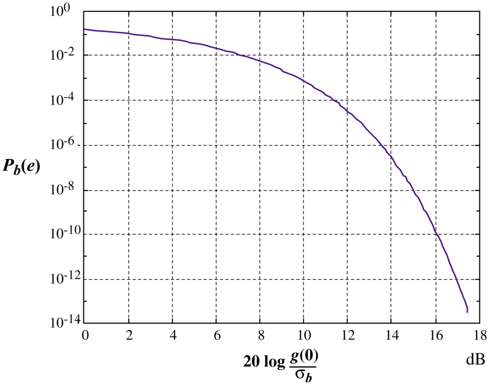
$$P_b(e) = P_b(e|a_k = +1) \tag{62}$$

$$= P_b(e|a_k = -1) \tag{63}$$

$$= \int_{\gamma_0}^{\infty} f_{R_k}(r_k|a_k = -1)\, \mathrm{d}r_k \tag{64}$$

$$= \frac{1}{\sqrt{2\pi}} \int_{\frac{g(0)}{\sigma_b}}^{\infty} \mathrm{e}^{-\frac{z^2}{2}}\, \mathrm{d}z \tag{65}$$

$$= Q\left[\frac{g(0)}{\sigma_b}\right] \tag{66}$$



# Impairments : Channel Noise
## Error probability (for equiprobable binary symbols)

### Exercise

Compute the average duration time with only one error for an error probability of $1 \times 10^{-3}$ for a bit rate of $100\,\mathrm{Mbits/s}$

# Impairments : Channel Noise

## Error probability (for equiprobable binary symbols)

### Exercise

Compute the average duration time with only one error for an error probability of $1 \times 10^{-3}$ for a bit rate of $100\,\mathrm{Mbits/s}$

- Solution

$$P_b = \frac{N_{\text{error}}}{N_{\text{bits}}} = 1 \times 10^{-3}\ \text{err/bit} \tag{67}$$

$$D_R = \frac{N_{\text{bits}}}{\Delta T} = 100\ \text{Mbits/s} \tag{68}$$

$$\Delta T = \frac{N_{\text{error}}}{P_b \cdot D_R} = \frac{1\ \text{error}}{1 \times 10^{-3}\ \text{error/bit} \times 100\ \text{Mbits/s}} = 1 \times 10^{-5}\ \text{s} \tag{69}$$

# Impairments : Channel Noise

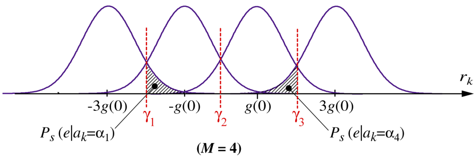
## Error probability (for M-ary symbols)

$$a_k(\omega) \sim \{\alpha_1, \alpha_2, \ldots, \alpha_n\} = \{\pm 1, \pm 3, \ldots, \pm(M-1)\} \quad (70)$$

Here, the bit error probability is not anymore the symbol error probability :

$$\frac{P_s(e)}{\log_2 M} \leq P_b(e) \leq P_s(e) \quad (71)$$

$$f_{R_k}(r_k | a_k = \alpha_i) = f_B(r_k - \alpha_i \cdot g(0)) \quad (72)$$



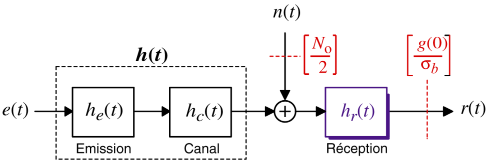

$$P_s(e) = 2\frac{M-1}{M} Q\left(\frac{g(0)}{\sigma_b}\right) \quad (73)$$

$$P_b(e) = 2 \times \frac{M-1}{M \cdot \log_2 M} Q\left(\frac{g(0)}{\sigma_b}\right) \quad (74)$$

(Lower bound (Gray coding))

# Outline

# Impairments : Optimal RX – Matched filter
## Output SNR




- For a given transmitter and propagation channel, the output SNR is a function of the characteristics of the reception filter.

$$g(0) = \int_{\mathbb{R}} \overset{\circ}{h}(f) \cdot \overset{\circ}{h}_r(f)\, \mathrm{d}f \qquad (75)$$

$$\sigma_b^2 = \frac{N_0}{2} \int_{\mathbb{R}} |h_r(f)|^2\, \mathrm{d}f \qquad (76)$$

- Minimize $P(e) = K \cdot Q\left(\frac{g(0)}{\sigma_b}\right)$

$\Longrightarrow$ maximize $\frac{g(0)}{\sigma_b}$

$\Longrightarrow$ optimize $g(\cdot)$

$$\Rightarrow \left[\frac{g(0)}{\sigma_b}\right]^2 = \frac{2}{N_0} \frac{\left(\int_{\mathbb{R}} \overset{\circ}{h}(f) \cdot \overset{\circ}{h}_r(f)\, \mathrm{d}f\right)^2}{\int_{\mathbb{R}} |h_r(f)|^2\, \mathrm{d}f} \qquad (77)$$

# Impairments : Optimal RX – Matched filter

## Maximum output SNR and optimal filter

- Using Schwarz inequality we show that:

$$\left[\frac{g(0)}{\sigma_b}\right]^2 \leq \frac{2}{N_0}\int_{\mathbb{R}} \left|\overset{\circ}{h}(f)\right|^2 df \qquad (78)$$

- Which is maximum for:

$$\overset{\circ}{h}_r(f) = \lambda \cdot \overset{\circ}{h}^{*}(f) \qquad (79)$$

- Which yields the following time domain condition:

$$\boxed{\boldsymbol{h_r(t) = \lambda \cdot h^*(-t)}} \qquad \heartsuit \qquad (80)$$

  - This is what is called the ***matched filter***

- Let us write $E_h$ the energy of the channel+TX filter $h$:

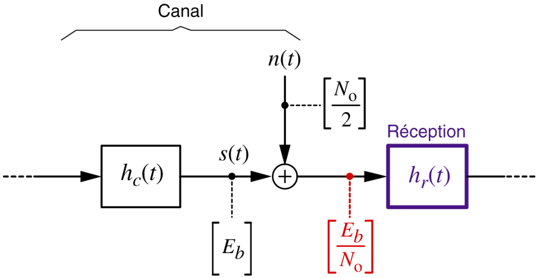
$$\frac{g(0)}{\sigma_b} \leq \sqrt{\frac{2E_h}{N_0}} \qquad (81)$$

- The value of the maximum SNR does not depend on the shape of the received pulse but only on its energy.

# Impairments : Optimal RX – Matched filter

## Normalizing the SNR to the bit energy

- In order to be able to carry out performance comparisons according to the values of $M$, the average energy per bit $E_b$ is used instead of $E_h$:




$$E_b \triangleq P_s T_b = \frac{P_s}{D_b} \qquad (82)$$

- $P_s$ : average useful signal power at the input of the receiver
- $T_b$ : bit period

# Impairments : Optimal RX – Matched filter

## Error probability as function of bit energy and noise PSD

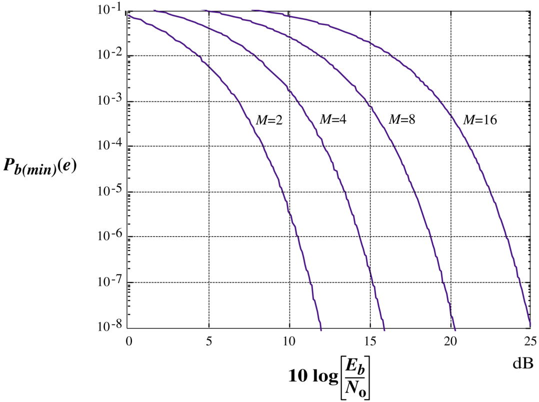
$$P_s = \int_{\mathbb{R}} \mathring{\gamma}_s(f)\, \mathrm{d}f = \frac{\sigma_a^2}{T_s} \int_{\mathbb{R}} \left| \mathring{h}(f) \right|^2 \mathrm{d}f \qquad (83)$$

$$= \frac{M^2 - 1}{3\, T_s} E_h \qquad (84)$$

Hence:

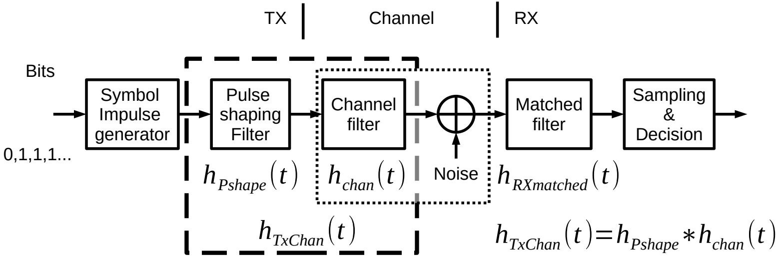
$$E_b = \frac{M^2 - 1}{3 \log_2 M} E_h \qquad (85)$$

$$P_{b(\min)}(e) = 2 \frac{M - 1}{M \log_2 M} Q\left( \sqrt{\frac{6 \log_2 M}{M^2 - 1} \frac{E_b}{N_0}} \right) \qquad (86)$$



# Impairments : Optimal RX – Matched filter

## Conclusion : Transceiver model (optimal receiver)




### Optimal transceiver ♡

The TX and RX filters $h_{Pshape}(t)$ and $h_{RXmatched}(t)$ are set so that:

- $h_{tot}(t) = h_{Pshape} * h_{chan} * h_{RXmatched}(t)$ is an *ISI Nyquist* filter
- $h_{RXmatched} = h^{*}_{TxChan}(-t)$ is the *matched filter* of $h_{TxChan}$

# Outline

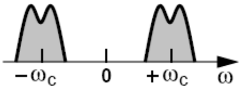
# Outline

# Journey of the Signal (recall) (♣)

Baseband modulation

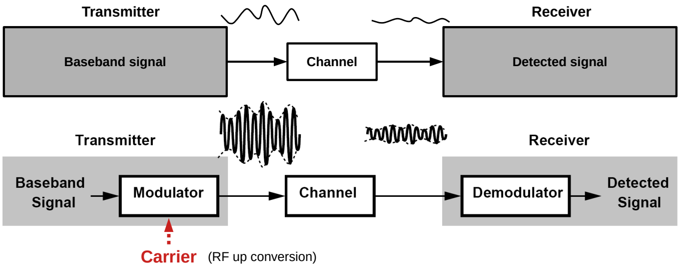

$$x(t) = a(t) \tag{87}$$

Passband modulation

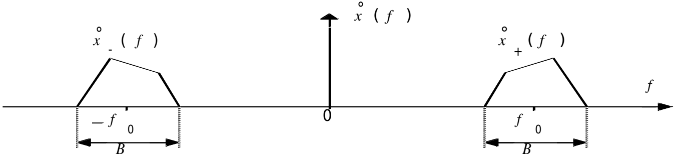



$$x(t) = a(t)\cos(\omega_c t + \theta(t)) \tag{88}$$

- For bandpass modulation : several parameters of a sinusoidal carrier can be modulated according to the baseband signal.



# Bandpass Model




$$\overset{\circ}{x}(f) = \overset{\circ}{x}_{-}(f) + \overset{\circ}{x}_{+}(f) \quad (89)$$

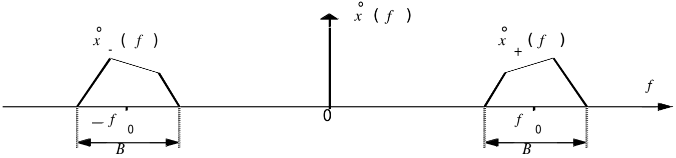
$$\overset{\circ}{x}_{+}(f) = \overset{\circ}{x}(f) \cdot u(f) \quad (90)$$

$$\overset{\circ}{x}_{-}(f) = \overset{\circ}{x}(f) \cdot u(-f) \quad (91)$$

**Self-exercise**

What is the property if $x$ is a real-valued signal ?

# Bandpass Model




$$\overset{\circ}{x}(f) = \overset{\circ}{x}_-(f) + \overset{\circ}{x}_+(f) \tag{89}$$

$$\overset{\circ}{x}_+(f) = \overset{\circ}{x}(f) \cdot u(f) \tag{90}$$

$$\overset{\circ}{x}_-(f) = \overset{\circ}{x}(f) \cdot u(-f) \tag{91}$$

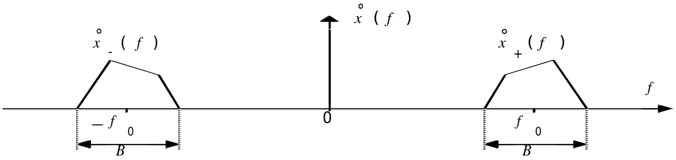
## Self-exercise

What is the property if $x$ is a real-valued signal ?

- Solution (already seen) : $\overset{\circ}{x}(f)$ is *Hermitian*

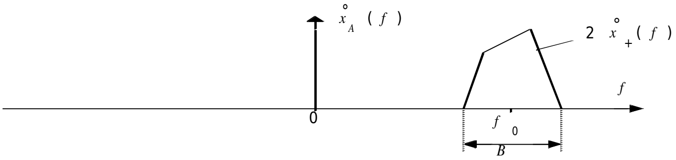
$$\overset{\circ}{x}(-f) = \overset{\circ}{x}^*(f) \tag{92}$$

# Analytic signal (visual definition)



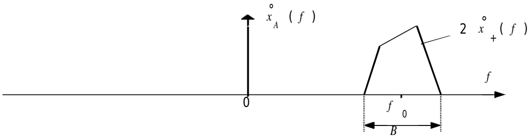

- Analytic signal

$$\overset{\circ}{x}_A(f) \triangleq 2\overset{\circ}{x}_+(f) \tag{93}$$



# Analytic signal (formal definition)

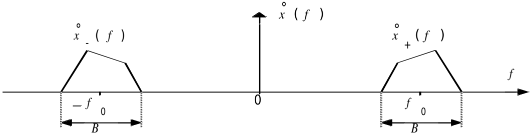
- When $x(t)$ is real-valued, then $x_A(t)$ can be written:

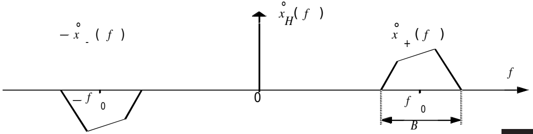



$$x_A(t) = x(t) + jx_H(t) \qquad (94)$$

  - where $x_H(t)$ is the Hilbert transform of $x(t)$

$$\mathring{x}_H(f) = -j\,\mathrm{sgn}(f)\cdot\mathring{x}(f) \qquad (95)$$



# Analytic signal (exercise)

## Analytic signal of $\cos(\omega_0 t)$

$$x(t) = \cos(\omega t) \quad \longleftrightarrow \quad X(\omega) = \frac{1}{2}\left(\delta(\omega+\omega_0) + \delta(\omega-\omega_0)\right) \tag{96}$$

# Analytic signal (exercise)

## Analytic signal of $\cos(\omega_0 t)$

$$x(t) = \cos(\omega t) \quad \longleftrightarrow \quad X(\omega) = \frac{1}{2}\left(\delta(\omega+\omega_0) + \delta(\omega-\omega_0)\right) \tag{96}$$

$$X_H(\omega) = -j\,\mathrm{sgn}(\omega)\cdot H(\omega) \tag{97}$$

$$= (-j)\cdot(-1)\cdot\frac{1}{2}\delta(\omega+\omega_0) + (-j)\cdot(+1)\cdot\frac{1}{2}\delta(\omega-\omega_0) \tag{98}$$

$$= -j\cdot\frac{1}{2}\left(\delta(\omega-\omega_0) - \delta(\omega+\omega_0)\right) \tag{99}$$

$$\Longrightarrow \quad x_H(t) = -j\cdot\frac{1}{2}\left(\mathrm{e}^{j\omega_0 t} - \mathrm{e}^{-j\omega_0 t}\right) = \sin(\omega_0 t) \tag{100}$$

$$x_A(t) = x(t) + jx_H(t) \tag{101}$$

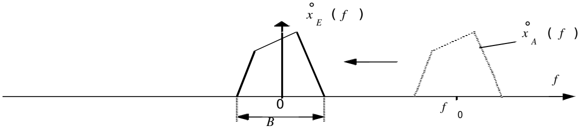
$$= \cos(\omega t) + j\sin(\omega_0 t) \tag{102}$$

$$= \mathrm{e}^{j\omega t} \tag{103}$$

# Complex baseband envelope

- Shift analytic signal to DC

$$\overset{\circ}{x}_E(f) \triangleq \overset{\circ}{x}_A(f - f_0) = 2\overset{\circ}{x}_+(f - f_0) \tag{104}$$

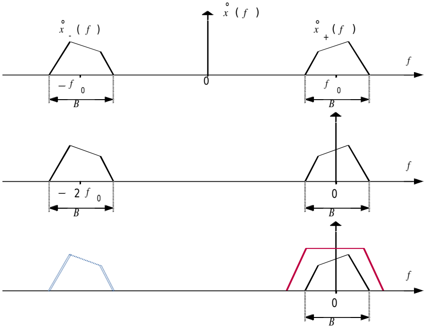


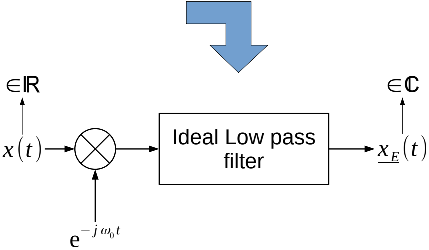

## Fundamental formula

$$x(t) = \Re\left\{x_E(t) \cdot e^{j2\pi f_0 t}\right\} \tag{105}$$

$x_E(t)$ contains all the information of $x(t)$ ; **the carrier frequency can be ignored !**

# From RF to complex baseband envelope (signal processing model)




## Note

The actual hardware implementation is slightly different (cf. IQ transceivers)

# Complex baseband envelope : resulting RF signals

- $x_E(t)$ is a complex signal and it can be developped as:

- Cartesian components

$$\underline{x}_E(t) = i(t) + jq(t) \tag{106}$$

- Then :

$$x(t) = \Re\left\{x_E(t) \cdot \mathrm{e}^{j\omega_0 t}\right\} \tag{107}$$

$$= i(t)\cos(\omega_0 t) - q(t)\sin(\omega_0 t) \tag{108}$$

- Polar components

$$\underline{x}_E(t) = a(t)\,\mathrm{e}^{j\varphi(t)} \tag{109}$$

- Then :

$$x(t) = a(t)\cos(\omega_0 t + \varphi(t)) \tag{110}$$

# Complex baseband envelope : Exercise

**Exercise**

Write both decomposition of $x_E(t)$ when $x(t)$ is a (real-valued) AM signal

# Complex baseband envelope : Exercise

## Exercise

Write both decomposition of $x_E(t)$ when $x(t)$ is a (real-valued) AM signal

- Solution

$$x_{AM}(t) = A(t)\cos(\omega_0 t + \Phi) = \Re\left\{A(t)e^{j\Phi}e^{\omega_0 t}\right\} \quad (111)$$

Cartesian:

$$x_E(t) = A(t)\left(\cos(\Phi) + j\sin(\Phi)\right) \quad (112)$$

Polar:

$$x_E(t) = A(t)e^{j\Phi} \quad (113)$$

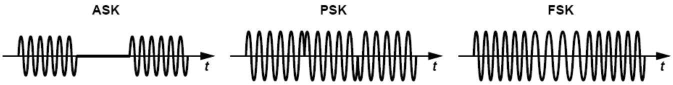
# Conclusion

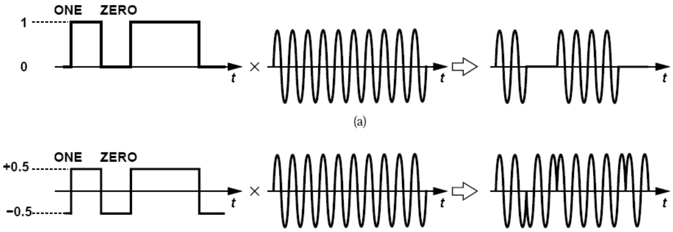
- Bandpass modulation is :
  - A (complex-valued) baseband modulation *mixed with* a (complex-valued) carrier
- Recall baseband pulse shaping and error probabilities of previous (real-valued) baseband modulation
  - Apply exactly the same on the (complex-valued) baseband model !

# Outline

# Basic bandpass digital modulation: ASK, PSK, FSK (♣)

"Amplitude Shift Keying", "Phase Shift Keying", and "Frequency Shift Keying"



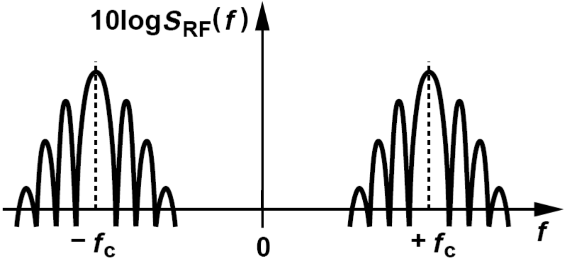




(a)

(b)

- Signals look analog but data is digital

# The Spectrum of PSK and ASK Signal (recall) (♣)




$$x_{PSK}(t) = x_{BB}(t)\cos\omega_c t \tag{114}$$

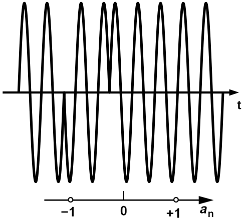
- The upconversion operation shifts the spectrum to $\pm f_c$
- Spectrum of ASK is similar but with impulses at $\pm f_c$
- Same as baseband transmission : the spectrum must be constrained

# Signal constellation concept and waveforms (♣)

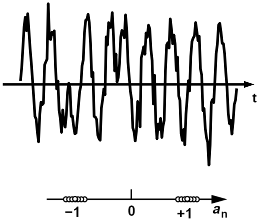
- Simple example : Binary PSK and ASK

$$x_{PSK}(t) = a_n \cos\omega_c t, \quad a_n = \pm 1 \quad (115) \qquad x_{ASK}(t) = a_n \cos\omega_c t, \quad a_n = \{0, 1\} \quad (116)$$

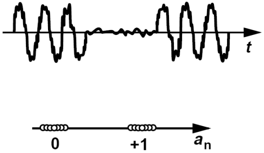
- Ideal




- Noisy

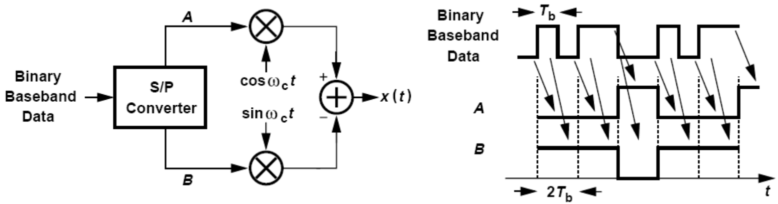



- Noise corrupts the amplitude for both ZEROs and ONEs.




- The **constellation** is the graphical representation of the symbols
  - When the transmission is noisy, the received symbols form cloud points around the ideal position

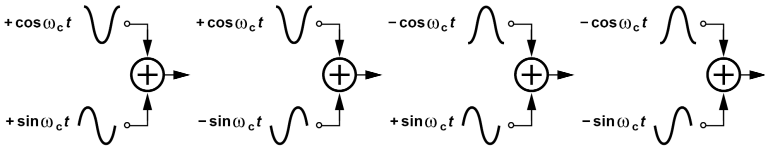
# Quadrature Modulation (♣)



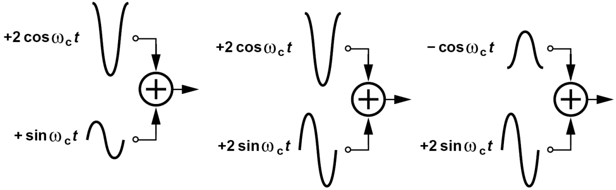

$$x(t) = b_{2m} A_c \cos \omega_c t + b_{2m+1} A_c \sin \omega_c t \qquad (117)$$

- QPSK halves the occupied bandwidth
  - In general, that is the case for all quadrature modulations. . .
- Pulses appear at *A* and *B* are called *symbols* rather than *bits*

# Quadrature Amplitude Modulation (QAM) (carrier level waveforms) (♣)



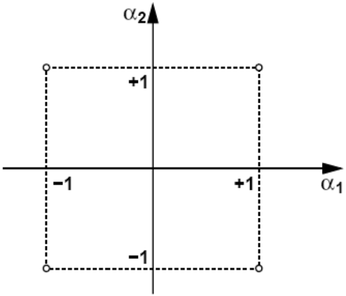

(a)

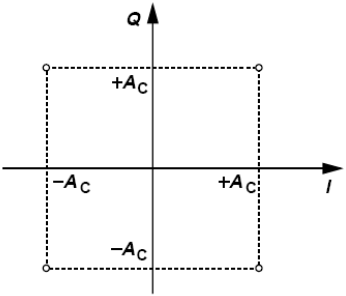



(b)

- *QAM* allows four possible amplitudes for sine and cosine, $\pm 1$, $\pm 2$

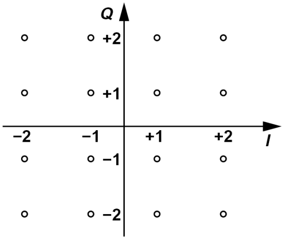
# Quadrature Modulation: Constellation (♣)




- $I$ for in-phase and $Q$ for Quadrature

# Quadrature Modulation: Constellation (♣)

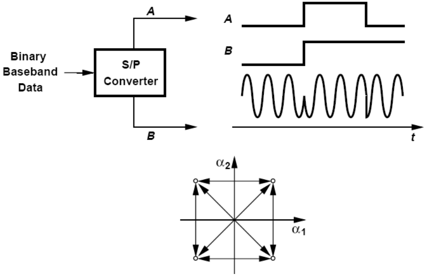


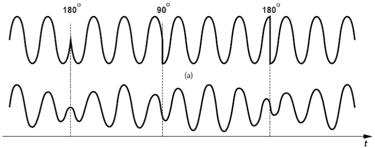

$$x_{16QAM}(t) = \alpha_1 A_c \cos\omega_c t - \alpha_2 A_c \sin\omega_c t, \quad \{\alpha_1 = \pm 1, \pm 2; \alpha_2 = \pm 1, \pm 2\} \tag{118}$$

- Saves bandwidth
- Denser constellation: making detection more sensitive to noise
- (HW constraint : Large envelope variation: need highly linear PA)

# Outline

# Important Drawback of PSKs (QPSK, OQPSK,. . . ) (♣)



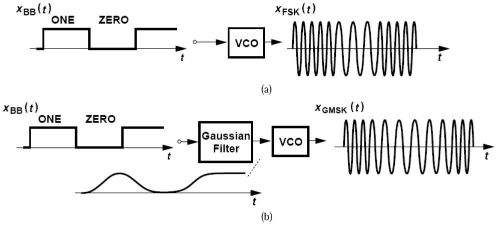




- Can use special constant envelope PA
- Important drawback of QPSK stems from the large phase changes at the end of each symbol.

- With pulse shaping, the output signal amplitude (“envelope”) experiences large changes each time the phase makes a 90 or 180 degree transition.
- Resulting waveform is called a “variable-envelope signal”.
  - therefore, adding pulse shaping requires to change PA (linear PA)

# Solution : filter phase before modulating carrier (♣)

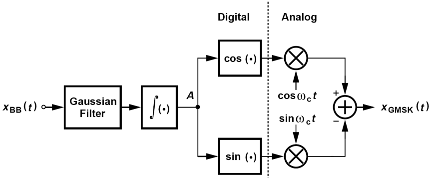
- GMSK and GFSK Modulation




$$x_{GMSK}(t) = A_c \cos\left[\omega_c t + m \int x_{BB} * h(t)\mathrm{d}t\right] \tag{119}$$

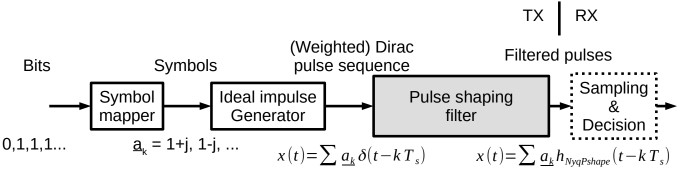
- Was used for GSM

- GMSK Modulator
  - Straightforwardly implemented using IQ modulator

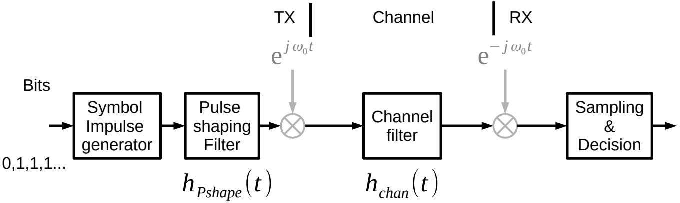


# General case : Pulse shaping filter

- Applies to (complex-valued) baseband signal (exactly the same as previously)




- Noiseless channel

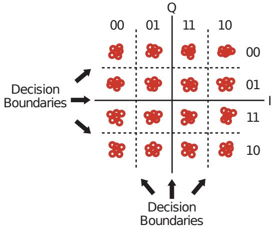



- The ISI properties offered by pulse shaping filter is maintained if exact pulse shape is preserved at the carrier
  - $h_{chan}$ is flat on the transmission band

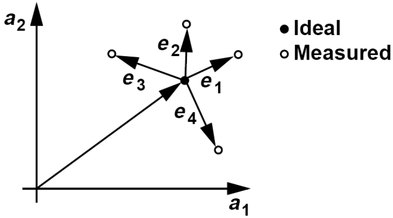
# Outline

# Constellation Diagram with noise (♣) (▲)

- Sampled data values no longer land in exact same location across all sample instants
- Decision boundaries remain fixed




- Derived Metrics : EVM




$$EVM_{rms} = \frac{1}{V_{rms}} \sqrt{\frac{1}{N} \sum_{j=1}^{N} e_j^2} \quad (120)$$

$$EVM_{pow} = \frac{1}{P_{avg}} \cdot \frac{1}{N} \sum_{j=1}^{N} e_j^2 \quad (121)$$

# Error probabilities (▼)

- Requires further assumptions (nearest neighbor approximation)

| Modulation | $P_s(\gamma_s)$ | $P_b(\gamma_b)$ |
|---|---|---|
| BPSK | | $= Q\left(2\sqrt{\gamma_b}\right)$ |
| QPSK | $\approx 2Q\left(\sqrt{\gamma_s}\right)$ | $\approx Q\left(2\sqrt{\gamma_b}\right)$ |
| MPSK | $\approx 2Q\left(\sqrt{2\gamma_s}\sin\left(\frac{\pi}{M}\right)\right)$ | $\approx \frac{2}{\log_2 M} Q\left(\sqrt{2\gamma_b \log_2 M}\sin\left(\frac{\pi}{M}\right)\right)$ |
| M-QAM | $\approx 4Q\left(\sqrt{\frac{3\bar{\gamma}_s}{M-1}}\right)$ | $\approx \frac{4}{\log_2 M} Q\left(\sqrt{\frac{3\bar{\gamma}_b \log_2 M}{M-1}}\right)$ |

$$\gamma_s = E_s/N_0 \quad (122) \qquad \gamma_b = E_b/N_0 \quad (123) \qquad \gamma_b = \frac{\gamma_s}{\log_2 M} \quad (124) \qquad P_b \approx \frac{P_s}{\log_2 M} \quad (125)$$

- Common form :

$$P_s(\gamma_s) \approx \alpha_M Q\left(\sqrt{\beta_M \gamma_s}\right) \quad (126) \qquad P_b(\gamma_b) \approx \hat{\alpha}_M Q\left(\sqrt{\hat{\beta}_M \gamma_b}\right) \quad (127)$$

# Outline

Communication theory : Digital modulations

Baseband

Passband

Introduction

Analysis of some (digital) bandpass modulations

Impairments : Channel bandwidth limitation (TX)

Impairments : Channel Noise (RX analysis)

Impairments : Fading channels

Impairments : Mitigating multipath fading – OFDM

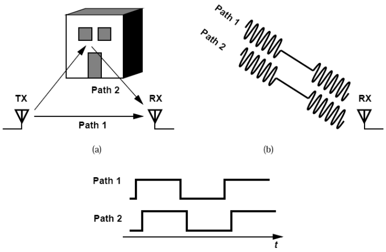
Cellular communication considerations (mobile systems and multiple access)

Hardware considerations

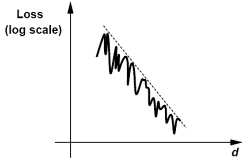
Wireless standards

Transceivers architectures

# Multipath Propagation (♣)




- Multipath Propagation may lead to considerable *intersymbol interference*
- Direct path: signals experience a power loss proportional to the square of the distance
- Reflective path: loss increases with the fourth power of the distance




- Multi-path fading: two signals possibly arriving at the receiver with opposite phases and roughly equal amplitudes, the net received signal may be very small

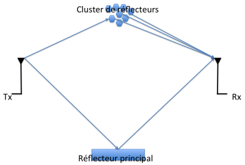
# Radio channel model

- Emitted signal

$$s(t) = \Re\left\{u(t)\cdot e^{j\omega_0 t}\right\} = \Re\left\{a(t)e^{j\varphi(t)}\cdot e^{j\omega_0 t}\right\} = I(t)\cos(\omega_0 t) - Q(t)\sin(\omega_0 t) \quad (128)$$

- Received signal

$$r(t) = \Re\left\{\sum_{n=0}^{N(t)} \alpha_n(t)u(t-\tau_n(t))\cdot e^{j(\omega_0(t-\tau_n(t))+\Phi_n^D)}\right\} \quad (129)$$



# Radio channel model

- Other form:

$$r(t) = \Re\left\{ \mathrm{e}^{j\omega_0 t} \sum_{n=0}^{N(t)} \alpha_n(t) u(t - \tau_n(t)) \cdot \mathrm{e}^{-j\Phi_n(t))} \right\} \tag{130}$$

$$\Phi_n(t) = \omega_0 \tau_n(t) - \Phi_n^D \tag{131}$$

- $n$ corresponds to a path with
  - length

$$L_n \rightarrow \tau_n = \frac{L_n}{c} \tag{132}$$

  - Doppler value

$$\Phi_n^D = \int 2\pi f_n^D(t) \mathrm{d}t, \qquad f_n^D(t) = \frac{v \cos\theta_n(t)}{\lambda} \tag{134}$$

  - Attenuation

$$\alpha_n(t) \tag{133}$$

  - Angle of the movement direction

$$\theta_n(t) \tag{135}$$

# Fading models (time domain definition)
## Delay spread

- Simple definition of *Delay spread* :

$$\overline{\Delta_\tau} \sim \max(\tau_n) - \min(\tau_n) \tag{136}$$

- More accurate :

$$\overline{\Delta_\tau} = \mathrm{RMS}\left\{\tau_n\right\} \tag{137}$$

- Symbol period : $T_s$
- Signal bandwidth : $B_S \approx T_s^{-1}$

### Definition of selective fading models

When :

$$\overline{\Delta_\tau} \ll T_s : \text{narrowband fading model} \tag{138}$$

$$\overline{\Delta_\tau} \gg T_s : \text{wideband fading model} \tag{139}$$

# Fading models (time domain interpretation)
## Delay spread

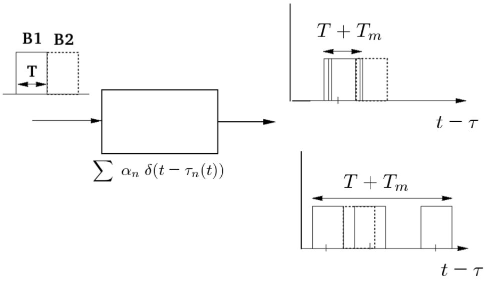
- Narrowband fading model: $\overline{\Delta_\tau} \ll T_s$
- Then we can consider that

$$\tau_i \ll T_s \qquad (140)$$
$$u(t-\tau_i) \approx u(t) \qquad (141)$$

- Then :

$$r(t) \approx \Re\left\{ u(t)\cdot e^{j\omega_0 t} \underbrace{\left( \sum_{n=0}^{N(t)} \alpha_n(t)\cdot e^{j\Phi_n^D} \right)}_{A(t)} \right\} \qquad (142)$$

- Wideband fading model: $\overline{\Delta_\tau} \gg T_s$




$\rightsquigarrow$ ISI

# Fading models (frequency domain interpretation)

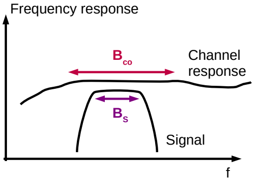
## Coherence band

- Coherence band (definition)

$$B_{co} \approx \frac{1}{\Delta_\tau} \tag{143}$$

- Narrowband fading model

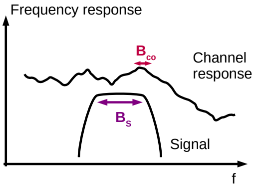
$$B_{co} \gg B_S \tag{144}$$




"Flat fading"

- Wideband fading model

$$B_{co} \ll B_S \tag{145}$$




"Frequency Selective fading"

## Doppler effect

- The *Doppler spread* reflects the variability of the channel over time
- Simple definition of *Doppler spread*:

$$\overline{\Delta_f^D} \sim \max(f_n^D) - \min(f_n^D) \tag{146}$$

- More accurate:

$$\overline{\Delta_f^D} = \mathrm{RMS}\left\{f_n^D\right\} \tag{147}$$

### Definition of variable fading models

When :

$$\overline{\Delta_f^D} \ll B_s : \text{slow fading model} \tag{148}$$

$$\overline{\Delta_f^D} \gg B_s : \text{fast fading model} \tag{149}$$

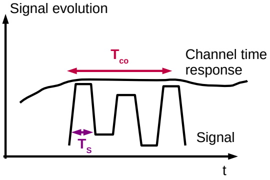
# Doppler effect – time domain interpretation
# Coherence time

- Coherence time (definition)

$$T_{co} \approx \frac{1}{\Delta_f^D} \tag{150}$$

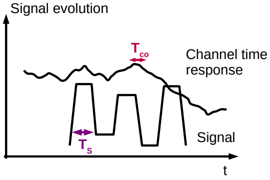
- Slow fading model

$$T_{co} \gg T_S \tag{151}$$

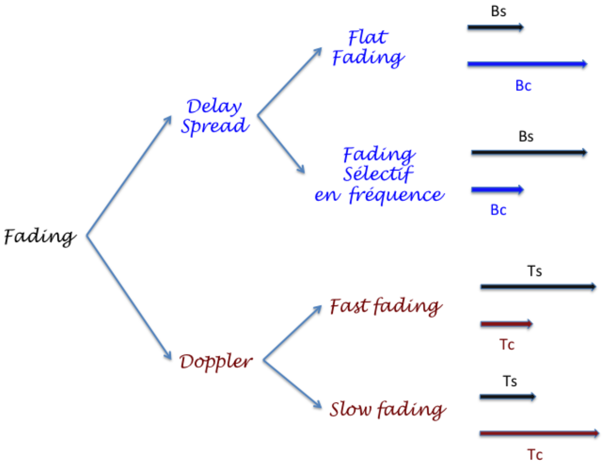



- Fast fading model

$$T_{co} \ll T_S \tag{152}$$



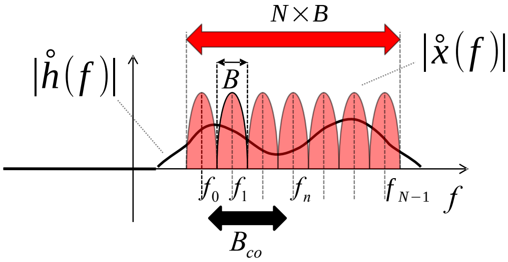
# Summary of RF channel models

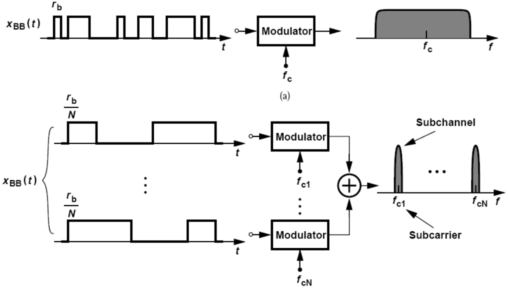


# Outline

# Multiple Carrier Principle (♣)

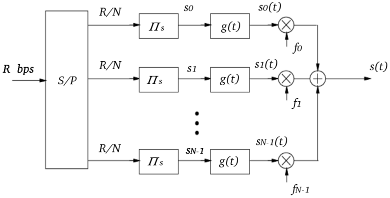
- Transmit $N$ Symbols in parallel during a period $T_s$ and band $B_s = 1/T_s$ in such a way that each symbol experiences almost a flat channel $B_s \ll B_{co}$ ($\tau_{max} \ll T_s$) while the total bandwidth $N \times B$ allows *High Capacity*

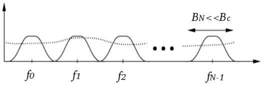



  - Here, the baseband data is first demultiplexed by a factor of $N$. The $N$ streams are then impressed on $N$ different carrier frequencies.

# A multi-carrier modulator implementation




- A bit stream with bit rate $D$ bit/s is divided into $N$ streams with bit rate $D' = D/N$.
- Each stream is modulated by an $M$-state modulator (eg M-QAM).
- Each stream is shaped by a Nyquist filter $g(t)$ (roll-off $\beta$)
- Then each stream is transposed around the frequency $f_i$.

- During the MC symbol period, the output is :

$$s(t) = \sum_{i=0}^{N-1} \alpha_i g(t) \cos(2\pi f_i t + \phi_i) \qquad (153)$$

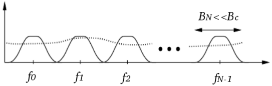
- The MC symbol period is :

$$T_{MC} = \frac{1+\beta}{B_N} = \frac{(1+\beta)N}{B_S} \qquad (155)$$

# Spectral efficiency and orthogonality

- This system is spectrally inefficient

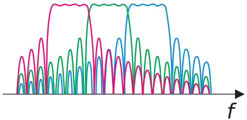
$$B_{tot} = N \times B_N = \frac{(1+\beta)\cdot N}{T_{MC}} \qquad (156)$$




- The total bandwidth can be reduced by "bringing the carriers closer" to each other making the signal carriers **overlapping**. This can be done thanks to the orthogonality of the set

$$\{\cos\left(2\pi\left(f_0 + iB_N\right)t + \Phi_i\right)\}, \quad i = 0, 1, \ldots, N-1 \qquad (157)$$

on the interval $[0, T_{MC}]$




- Now, carriers are separated by

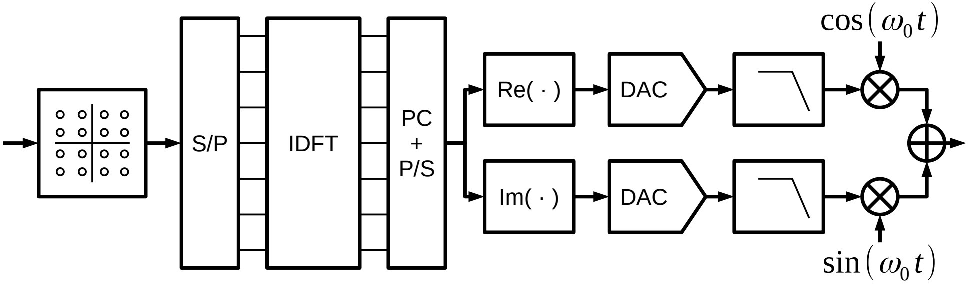
$$\Delta_{f,\mathrm{ortho}} = \frac{1}{T_{MC}} \quad \left( < B_N = \frac{1+\beta}{T_{MC}} \right) \qquad (158)$$

$$B_{\mathrm{tot}} \approx N \times \Delta_{f,\mathrm{ortho}} = \frac{N}{T_{MC}} \qquad (159)$$

N. Sambo et al., https://ieeexplore.ieee.org/stamp/stamp.jsp?tp=&arnumber=7045405&isnumber=7045380

# Practical implementation of OFDM

- The previously depicted OFDM transmitter is very complex as it requires *N* carrier frequencies and modulators (i.e. *N* oscillators and mixers).
- In practice, the subchannel modulations are performed in the digital baseband and subsequently converted to analog form.



# Outline

Communication theory : Digital modulations

- Baseband
- Passband
- Cellular communication considerations (mobile systems and multiple access)
- Hardware considerations
- Wireless standards
- Transceivers architectures

# Mobile RF Communications: Cellular System (♣)

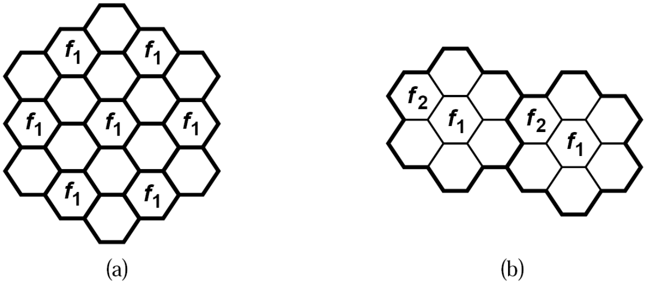



Figure 3.42. (a) Simple cellular system, (b) 7-cell reuse pattern.

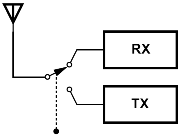
- Immediate neighbors cannot utilize same frequency
- The mobile units in each cell are served by a base station, and all of the base stations are controlled by a “mobile telephone switching office” (MTSO)

# Summary of Multiple Access Methods

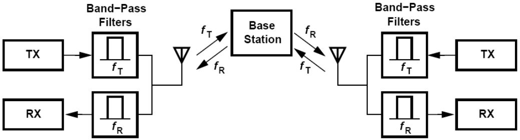
- TDMA
- FDMA
  - OFDMA
- CDMA
  - WCDMA

# Time and Frequency Division Duplexing (♣)

- TDD: same frequency band is utilized for both transmit and receive paths but the system transmits for half of the time and receives for the other half.



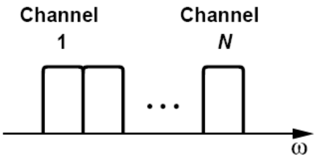

- FDD: employ two different frequency bands for the transmit and receive paths.



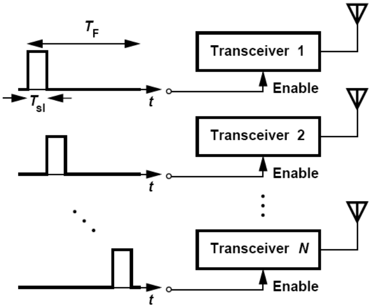

Have a look at appendix slide (11) for a quick comparison between FDD and TDD.

# Frequency-Division / Time-Division Multiple Access (♣)

- FDMA: available frequency band can be partitioned into many channels, each of which is assigned to one user.

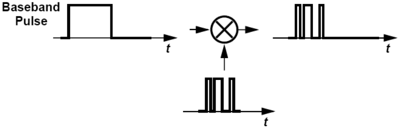


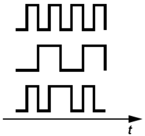

- TDMA: same band is available to each user but at different times




Have a look at appendix slide (12) for a quick comparison between TDMA and FDMA.

# Code-Division Multiple Access: Direct-Sequence CDMA (♣)

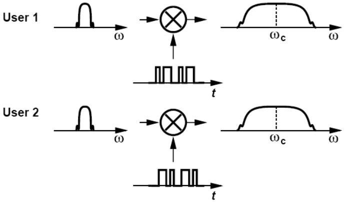


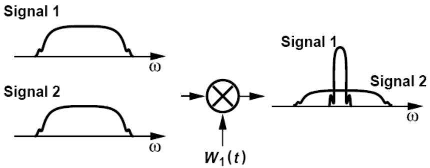

- Walsh's recursive equation

$$W_{2n} = \begin{bmatrix} W_n & W_n \\ W_n & \overline{W_n} \end{bmatrix} \qquad (160)$$

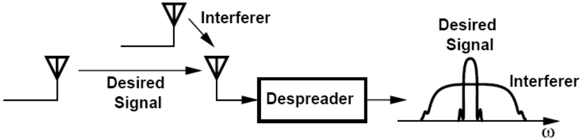
- CDMA allows the widened spectra of many users to fall in the same frequency band



# Direct-Sequence CDMA: decoding (♣)




- Desired signal is “despread”; Unwanted signal remains spread




- Near/Far Effect: one high-power transmitter can virtually halt communications among others: Requires Power Control

# Outline

Communication theory : Digital modulations

- Baseband
- Passband
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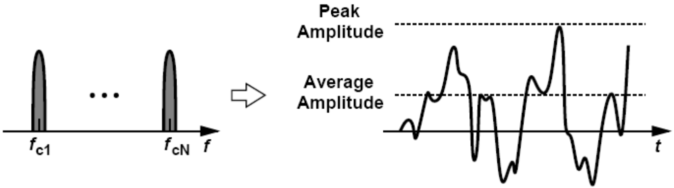
- Hardware considerations
- Wireless standards
- Transceivers architectures

# Peak-to-Average Ratio (♣)




$$\mathrm{PAR} = \frac{\max\left\{x^2(t)\right\}}{\overline{x^2(t)}} \tag{161}$$

- Large PAR: pulse shaping in the baseband, amplitude modulation schemes such as QAM, orthogonal frequency division multiplexing

# Spectral Regrowth: Constant vs. Variable Envelope (♣)

- Constant Envelope

$$x(t) = A_c \cos\left[\omega_c t + \varphi(t)\right] \tag{162}$$

- Let's consider a 3rd order nonlinearity

$$y(t) = \alpha_3 x^3(t) \tag{163}$$

$$= \alpha_3 A_c^3 \cos^3\left[\omega_c t + \varphi(t)\right] \tag{164}$$

$$= \frac{\alpha_3 A_c^3}{4} \cos\left[3\omega_c t + 3\varphi(t)\right] + \frac{\alpha_3 A_c^3}{4} \cos\left[\omega_c t + \varphi(t)\right] \tag{165}$$

- Shape of the spectrum in the vicinity of $\omega_c$ remains unchanged

- Variable Envelope

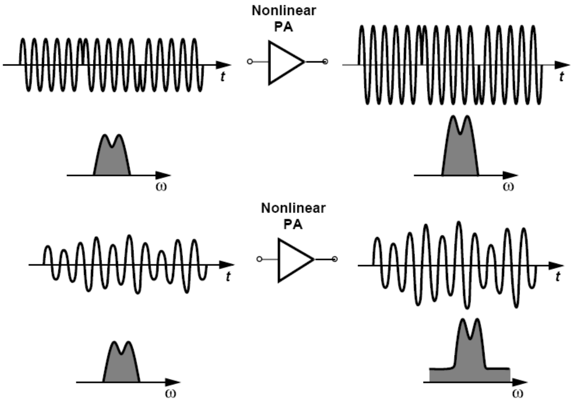
$$x(t) = x_I(t) \cos\omega_c t - x_Q(t) \sin\omega_c t \tag{166}$$

$$y(t) = \alpha_3 \left[x_I(t) \cos\omega_c t - x_Q(t) \sin\omega_c t\right]^3 + \cdots \tag{167}$$

$$= \alpha_3 x_I^3(t) \frac{\cos 3\omega_c t + 3\cos\omega_c t}{4} - \alpha_3 x_Q^3(t) \frac{-\cos 3\omega_c t + 3\sin\omega_c t}{4} \cdots \tag{168}$$

- Spectrum "grows" when a variable-envelope signal passes through a nonlinear system.

# Spectral Regrowth: An Illustration




- Constant Envelope: Shape of Spectrum unchanged
- Variable Envelope: Spectrum grows

# IQ mismatch Signal Constellation (♣)

- Due to circuit nonidealities, one of the carrier phases in a QPSK modulator suffers from a small phase error (“mismatch”) of $\theta$

$$x(t) = \alpha_1 A_c \cos(\omega_c t + \theta) + \alpha_2 A_c \sin \omega_c t, \qquad \{\alpha_1, \alpha_2\} = \{\pm 1, \pm 1\} \tag{169}$$

- The signal constellation at the output of this modulator is :

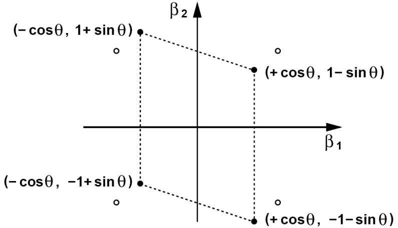
$$x(t) = \alpha_1 \cos\theta A_c \cos\omega_c t + (\alpha_2 - \alpha_1 \sin\theta) A_c \sin\omega_c t \tag{170}$$

$$\beta_1 = +\cos\theta, \qquad \beta_2 = 1 - \sin\theta, \tag{171}$$

$$\beta_1 = +\cos\theta, \qquad \beta_2 = -1 - \sin\theta, \tag{172}$$

$$\beta_1 = -\cos\theta, \qquad \beta_2 = 1 + \sin\theta, \tag{173}$$

$$\beta_1 = -\cos\theta, \qquad \beta_2 = -1 + \sin\theta \tag{174}$$




Further discussion on slide 143

# Outline

Communication theory : Digital modulations

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- Wireless standards
- Transceivers architectures

# Outline

Communication theory : Digital modulations

- Baseband
- Passband
- Cellular communication considerations (mobile systems and multiple access)
- Hardware considerations
- Wireless standards
  - Common specifications
  - W-CDMA example
- Transceivers architectures

# Wireless Standards: Common Specifications (♣)

- Frequency Bands and Channelization:
  - Each standard performs communication in an allocated frequency band
- Data Rates:
  - The standard specifies the data rates that must be supported
- Antenna Duplexing Method:
  - Most cellular phone systems incorporate FDD and other standards employ TDD
- Type of Modulation:
  - Each standard specifies the modulation scheme.

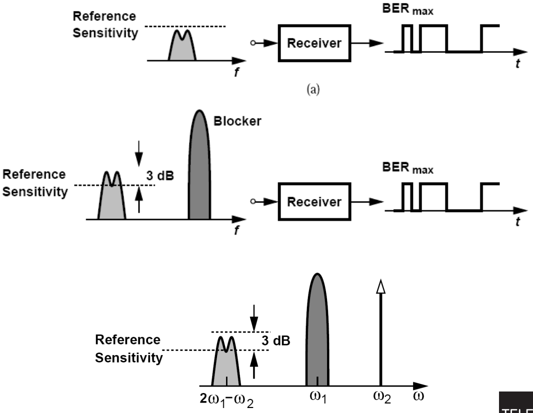
# Wireless Standards: Common Specifications (♣)

- TX output power:
  - The standard specifies the power levels that the TX must produce
- TX EVM and Spectral Mask:
  - The signal transmitted by the TX must satisfy several requirements like EVM and spectral mask
- RX Sensitivity:
  - The standard specifies the acceptable receiver sensitivity, usually in terms of maximum BER
- RX Input Level Range:
  - The standard specifies the desired signal range that the receiver must handle with acceptable noise or distortion

# Wireless Standards: Common Specifications (♣)

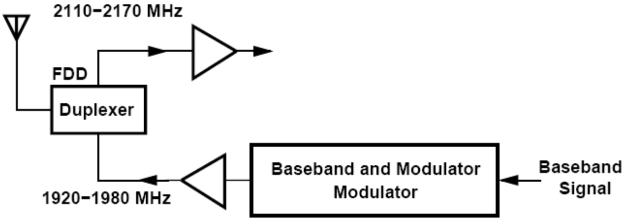
- RX Tolerance to Blockers:
  - The standard specifies the largest interferer that the RX must tolerate while receiving a small desired signal.

- Many standards also stipulate an intermodulation test



# Outline

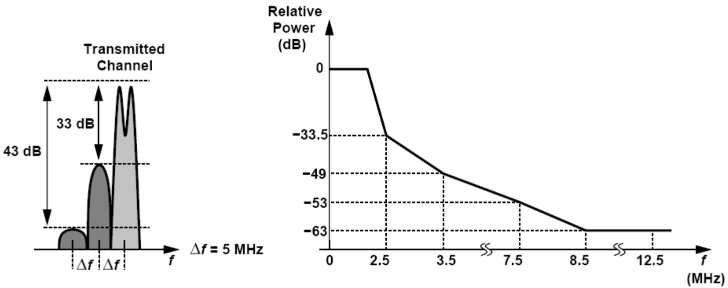
# Wideband CDMA: Air Interface (♣)




- BPSK for uplink, QPSK for downlink
- IMT-2000: total bandwidth 60 MHz,
- Each channel can accommodate a data rate 384 kb/s in a (spread) bandwidth of 3.84 MHz; but, with “guard bands” included, the channel spacing is 5 MHz.

CDMA : Code division multiple access – IMT-2000: International Mobile Telecommunications-2000

# Wideband CDMA: Transmitter Requirements (♣)

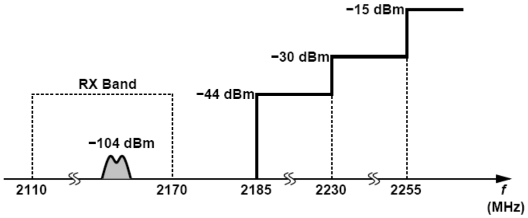



(a) (b)

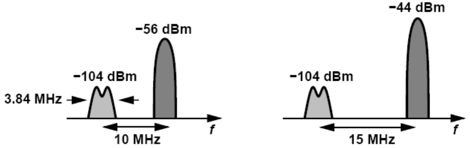
- Output power: $-49\,\mathrm{dBm}$ to $24\,\mathrm{dBm}$.
- Adjacent and alternate adjacent channel power $33\,\mathrm{dB}$ and $43\,\mathrm{dB}$ below main channel.

# Wideband CDMA: Receiver Requirements (♣)

- Blocking mask using a tone:

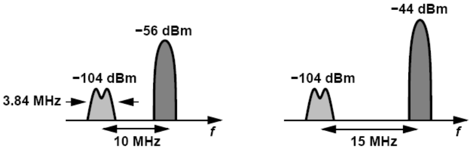



- Blocking test using a modulated interferer:




– Reference sensitivity: $-107\,\mathrm{dBm}$. Sinusoidal test for only out-of-band blocking

# Example of Wideband CDMA Receiver Requirements (♣)




- Estimate the required $P_{1\mathrm{dB}}$ of a WCDMA receiver satisfying the in-band test of figure.

- To avoid compression, $P_{1\mathrm{dB}}$ must be 4 to 5 dB higher than the blocker level, i.e., $P_{1\mathrm{dB}} \approx -40\,\mathrm{dBm}$.
- More precisely:
  - Consider $x(t) = A_1(1 + m\cos\omega_{m_1}t)\cos\omega_1 t + A_2(1 + m\cos\omega_{m_2}t)\cos\omega_2 t$ and expand $y(t) = \alpha_1 x(t) + \alpha_3 x^3(t)$ to get $\cos\omega_1 t$ amplitude, derive $A_1$ as a function of $A_2$.
  - We show that the input compression point must exceed $A_2$ $(= -44\,\mathrm{dBm})$ by about $1\,\mathrm{dB}$.

Have a look at appendix slides (14, 15, 16) for the complete derivation.

# Outline

Communication theory : Digital modulations

- Baseband
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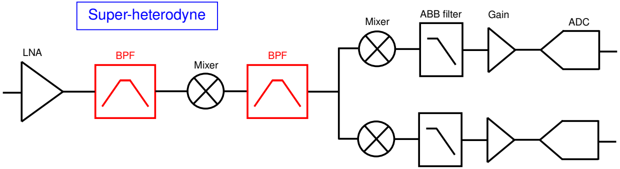
- Wireless standards
- Transceivers architectures

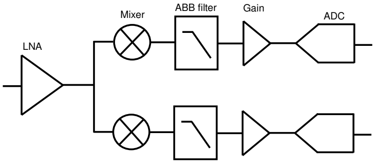
# Outline

# Receiver architectures (♠)

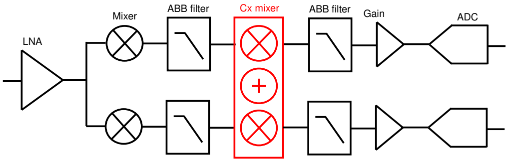
Super-heterodyne




Zero-IF

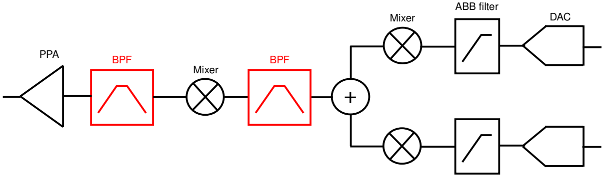



Low-IF

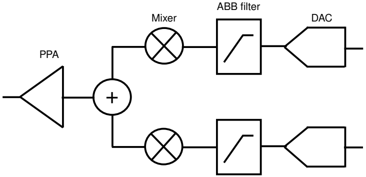


# Transmitter architectures (♠)

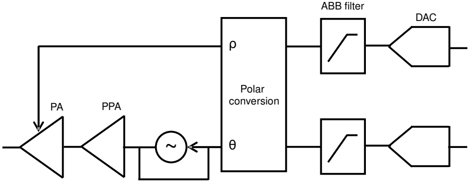
Dual-conversion




Zero-IF

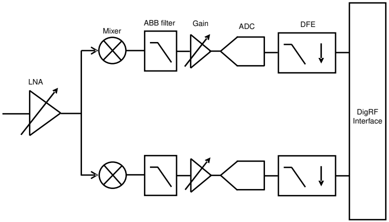



Polar modulation



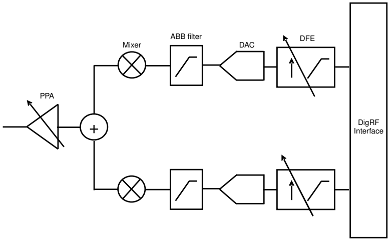
# Modern receiver (♠)
## Zero-IF RX

- Low-noise amplifier (LNA)
  - Gain
  - Noise performance
  - Linearity performance
- IQ Down-conversion Mixer
  - Frequency conversion
  - Gain
- IQ Analog baseband low-pass filter
  - Adjacent and blocker attenuation
  - Noise performance
  - Linearity performance
- IQ Analog-to-digital converter (ADC)
  - Digital conversion
  - Noise performance
  - THD performance
- IQ Digital front-end (RxDFE)
  - Down-sampling
  - Anti-alias filtering
  - Digital channel selection



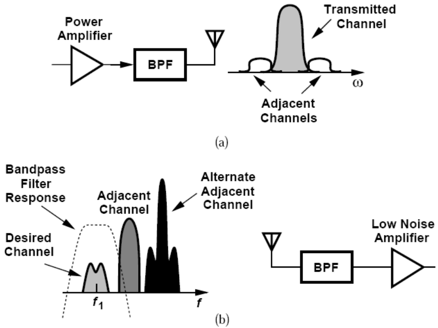
# Modern transmitter (♠)
## Zero-IF TX

- IQ Digital front-end (TxDFE)
  - Up-sampling
  - Anti-alias filtering
  - Noise filtering
  - Gain
- IQ Digital-to-Analog converter (DAC)
  - Analog conversion
  - Noise performance
- IQ Analog baseband low-pass filter
  - Anti-alias filtering
  - Noise performance
  - Spectral re-growth performance
- IQ Up-conversion Mixer and IQ modulator
  - Frequency conversion
  - Gain
- Pre-power amplifier (PPA)
  - Gain
  - Spectral re-growth performance
  - Pre-power amplification



# Outline

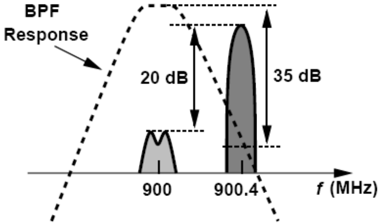
# General Considerations: Narrow Channel Bandwidth (♣)




- Narrow channel bandwidth impacts the RF design of the transceiver.

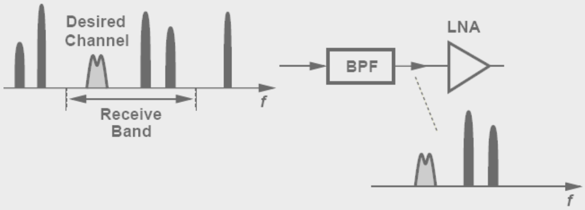
Recall Transceiver Specifications Lab for a practical example

# Can We Simply Filter the Interferers to Relax the Receiver Linearity Requirement? (♣)



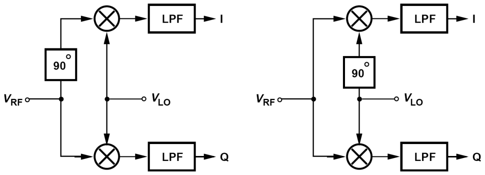

- First, the filter must provide a very high Q
- Second, the filter would need a variable, yet precise center frequency

# Channel Selection and Band Selection (♣)

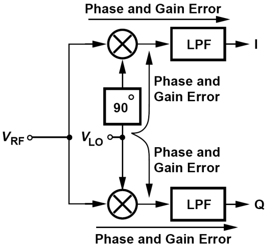



- All of the stages in the receiver chain that precede channel-selection filtering must be sufficiently linear
- Channel selection must be deferred to some other point where center frequency is lower and hence required Q is more reasonable
- Most receiver front ends do incorporate a “band-select” filter

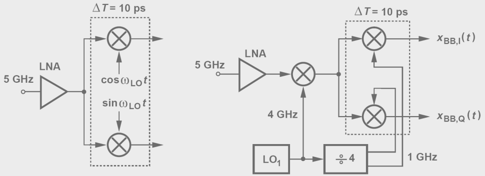
# I/Q Mismatch: Sources (♣)




- Separation into quadrature phases can be accomplished by shifting either the RF signal or the LO waveform by 90°.
- Errors in the 90° phase shift circuit and mismatches between the quadrature mixers result in imbalances in the amplitudes and phases of the baseband I and Q outputs.



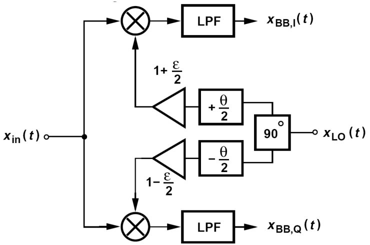
# I/Q Mismatch in Direct-Conversion Receivers And Heterodyne Topologies (♣)




- Quadrature mismatches tend to be larger in direct-conversion receivers than in heterodyne topologies.
- This occurs because
  1. the propagation of a higher frequency ($f_{in}$) through quadrature mixers experiences greater mismatches;
  2. the quadrature phases of the LO itself suffer from greater mismatches at higher frequencies;

# Effect of I/Q Mismatch (♣)

- Let us lump all of the gain and phase mismatches shown below:



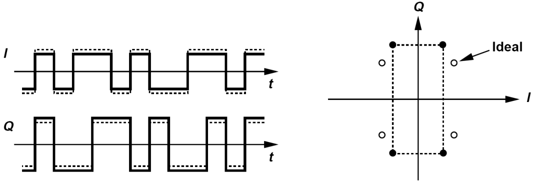

$$\begin{cases} x_{LO,I}(t) & = 2\left(1+\frac{\epsilon}{2}\right)\cos\left(\omega_c t+\frac{\theta}{2}\right) \\ x_{LO,Q}(t) & = 2\left(1-\frac{\epsilon}{2}\right)\cos\left(\omega_c t-\frac{\theta}{2}\right) \end{cases} \tag{175}$$

$$\begin{cases} x_{BB,I}(t) & = a\left(1+\frac{\epsilon}{2}\right)\cos\frac{\theta}{2} - b\left(1+\frac{\epsilon}{2}\right)\sin\frac{\theta}{2} \\ x_{BB,Q}(t) & = -a\left(1-\frac{\epsilon}{2}\right)\sin\frac{\theta}{2} + b\left(1-\frac{\epsilon}{2}\right)\cos\frac{\theta}{2} \end{cases} \tag{176}$$

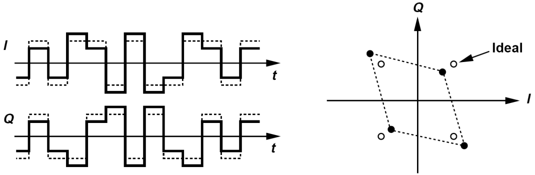
# Effect of I/Q Mismatch (♣)

- We now examine the results for two special cases:

1. $\epsilon \neq 0$, $\theta = 0$ : the quadrature baseband symbols are scaled differently in amplitude



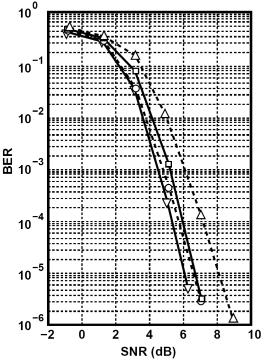

2. $\epsilon = 0$, $\theta \neq 0$ : each baseband output is corrupted by a fraction of the data symbols in the other output



# Requirement of I/Q Mismatch (♣)

- For complex signal waveforms such as OFDM with QAM, the maximum tolerable I/Q mismatch can be obtained by simulations

- The bit error rate is plotted for different combinations of gain and phase mismatches, providing the maximum mismatch values that affect the performance negligibly.
- For example, in a system employing OFDM with 128 subchannels and QPSK modulation in each subchannel, we observe that gain/phase mismatches below 0.6 dB/6° have negligible effect.



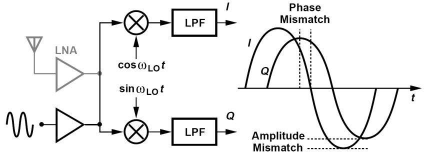

Effect of I/Q mismatch on an OFDM signal with QPSK modulation.
(∇: no imbalance;
○ : $\theta = 6°$, $\epsilon = 0.6$ dB;
□ : $\theta = 10°$, $\epsilon = 0.8$ dB;
△ : $\theta = 16°$, $\epsilon = 1.4$ dB.)

---

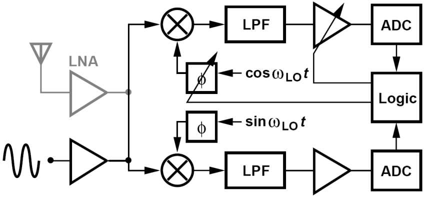
Also discussed for RX in slide 158

# Computation and Correction I/Q Mismatch (♣)

- In many high performance systems, the quadrature phase and gain must be calibrated – either at power-up or continuously.




- Calibration at power-up can be performed by applying an RF tone at the input of the quadrature mixers and observing the baseband sinusoids in the analog or digital domain
- With the mismatches known, the received signal constellation is corrected before detection.



# Outline

# Outline

From Wireless Standard Requirements to System Level Spec

Introduction

Example : From 3GPP requirements to RX System specifications WCDMA Band I (TS 25.101 Release 9)

From 3GPP requirements to TX System specifications WCDMA Band I (TS 25.101 Release 9)

# General Considerations: Units in RF Design (♣)

$$A_V|_{dB} = 20 \log \frac{V_{out}}{V_{in}} \qquad (177)$$

$$A_P|_{dB} = 10 \log \frac{P_{out}}{P_{in}} \qquad (178)$$

## Note

- This relationship between Power and Voltage only holds when the input and output impedance are equal

## Further derivation

$$A_P|_{dB} = 10 \log \frac{\frac{V_{out}^2}{R_0}}{\frac{V_{in}^2}{R_0}} \qquad (179)$$

$$= 20 \log \frac{V_{out}}{V_{in}} \qquad (180)$$

$$= A_V|_{dB} \qquad (181)$$

## dBm definition

$$P_{sig}|_{\mathrm{dBm}} = 10 \log \left( \frac{P_{sig}}{1\,\mathrm{mW}} \right) \qquad (182)$$

# Example of Units in RF (♣)

## Self-Exercise

An amplifier senses a sinusoidal signal and delivers a power of $0\,\mathrm{dBm}$ to a load resistance of $50\,\Omega$. Determine the peak-to-peak voltage swing across the load.

- Solution:

$$\frac{V_{pp}^2}{8R_L} = 1\,\mathrm{mW} \tag{183}$$

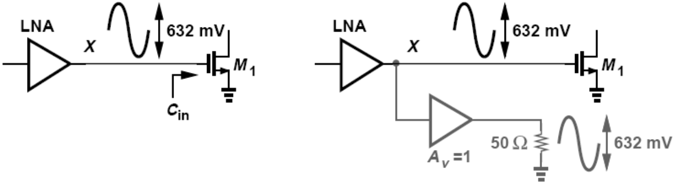
$$V_{pp} = 632\,\mathrm{mV} \tag{184}$$

# Example of Units in RF (♣)

## Example

- A GSM receiver senses a narrowband (modulated) signal having a level of $-100\,\mathrm{dBm}$. If the front-end amplifier provides a voltage gain of $15\,\mathrm{dB}$, calculate the peak-to-peak voltage swing at the output of the amplifier.
  - Since the amplifier output voltage swing is of interest, we first convert the received signal level to voltage. From the previous example, we note that $-100\,\mathrm{dBm}$ is $100\,\mathrm{dB}$ below $632\,\mathrm{mVpp}$. Also, $100\,\mathrm{dB}$ for voltage quantities is equivalent to $10^5$. Thus, $-100\,\mathrm{dBm}$ is equivalent to $6.32\,\mu\mathrm{Vpp}$. This input level is amplified by $15\,\mathrm{dB}$ ($\approx 5.62$), resulting in an output swing of $35.5\,\mu\mathrm{Vpp}$.

# dBm used at Interfaces Without Power Transfer (♣)




- dBm can be used at interfaces that do not necessarily entail power transfer
- We mentally attach an ideal voltage buffer to node X and drive a 50-Ω load. We then say that the signal at node X has a level of 0 dBm, tacitly meaning that if this signal were applied to a 50-Ω load, then it would deliver 1 mW.

## Receiver: Noise Figure (♣)

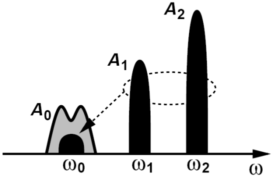
- 11a/g[2] specifies a packet error rate of $10\,\%$. This translates to a bit error rate of $1 \times 10^{-5}$, which in turn necessitates an SNR of $18.3\,\mathrm{dB}$ for 64QAM modulation. Since TX baseband pulse shaping reduces the channel bandwidth to $16.6\,\mathrm{MHz}$, for a sensitivity of $-65\,\mathrm{dBm}$ (at $52\,\mathrm{Mb/s}$), using sensitivity equation (24)[3], we can derive: $\mathrm{NF} = 18.4\,\mathrm{dB}$.
- In practice, signal detection in the digital baseband processor suffers from nonidealities, incurring a “loss” of a few decibels. Moreover, the front-end antenna switch exhibits a loss of around $1\,\mathrm{dB}$.
- Manufacturers typically target an RX noise figure of about $10\,\mathrm{dB}$.

[2] short for IEEE 802.11 a/b/g

[3] $\mathrm{Sensitivity} = -174\,\mathrm{dBm/Hz} + \mathrm{NF} + 10\log\mathrm{BW} + \mathrm{SNR}$

# Receiver: Nonlinearity (♣)

- We represent the desired, adjacent, and alternate channels by $A_0 \cos\omega_0 t$, $A_1 \cos\omega_1 t$, and $A_2 \cos\omega_2 t$, respectively.
- For a third-order nonlinearity of the form $y(t) = \alpha_1 x(t) + \alpha_2 x^2(t) + \alpha_3 x^3(t)$, the desired output is given by $\alpha_1 A_0 \cos\omega_0 t$ and the IM3 component at $\omega_0$ by $3\alpha_3 A_1^2 A_2/4$




- We choose the IM3 corruption to be around $-15$ dB to allow for other nonidealities:

$$20\log\left|\frac{3\alpha_3 A_1^2 A_2}{4\alpha_1 A_0}\right| = -15\,\text{dB} \quad (185)$$

$$20\log\left|\frac{3\alpha_3}{4\alpha_1}\right| = -15\,\text{dB} - 40\log A_1 - 20\log A_2 + 20\log A_0 \quad (186)$$

$$20\log\left|\frac{3\alpha_3}{4\alpha_1}\right| = 79\,\text{dBm} \quad (187)$$

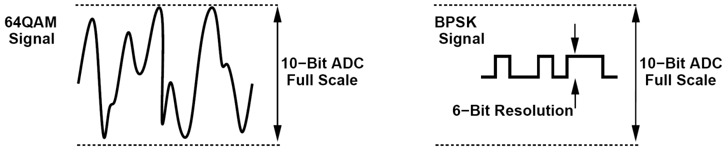
$$\mathrm{IIP}_3|_{\mathrm{dBm}} = 20\log\sqrt{\left|\frac{4\alpha_1}{3\alpha_3}\right|} \quad (188)$$

$$= -39.5\,\text{dBm} \quad (189)$$

# Example of AGC Range: Choice of Gain (♣)

## AGC for ADC full scale for each modulation

- The choice of the gain in the above example guarantees that the signal level reaches the ADC full scale for 64QAM as well as BPSK modulation. Is that necessary?
- No, it is not. The ADC resolution is selected according to the SNR required for 64QAM modulation (and some other factors). For example, a 10-bit ADC exhibits an SNR of about 62 dB, but a BPSK signal can tolerate a much lower SNR and hence need not reach the ADC full scale. In other words, if the BPSK input is amplified by, say, 60 dB rather than 84 dB, then it is digitized with 6 bits of resolution and hence with ample SNR ($\approx$ 38 dB).



# Receiver: I/Q Mismatch (♣)

- The I/Q mismatch study proceeds as follows.
  1. To determine the tolerable mismatch, we must apply in system simulations a 64QAM OFDM signal to a direct-conversion receiver and measure the BER or the EVM. Such simulations are repeated for various combinations of amplitude and phase mismatches, yielding the acceptable performance envelope.
  2. Using circuit simulations and random device mismatch data, we must compute the expected I/Q mismatches in the quadrature LO path and the downconversion mixers.
  3. Based on the results of the first two steps, we must decide whether the “raw” matching is adequate or calibration is necessary.

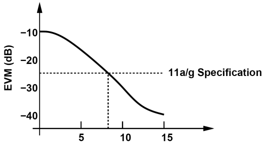
---

Same as TX ; see slide 147 for a plot example.

# Transmitter (♣)

- The transmitter chain must be linear enough to deliver a 64QAM OFDM signal to the antenna with acceptable distortion.

$$g\left(V_{in}\right) = \frac{\alpha V_{in}}{\left[1 + \left(\frac{V_{in}}{V_0}\right)^{2m}\right]^{\frac{1}{2m}}} \qquad (190)$$




- High linearity:
  1. assign most of gain to last PA stage
  2. minimize the number of stages in the TX chain

# Outline

From Wireless Standard Requirements to System Level Spec

- Introduction
- Example : From 3GPP requirements to RX System specifications WCDMA Band I (TS 25.101 Release 9)
- From 3GPP requirements to TX System specifications WCDMA Band I (TS 25.101 Release 9)

# Outline

## From Wireless Standard Requirements to System Level Spec

- Introduction
- Example : From 3GPP requirements to RX System specifications WCDMA Band I (TS 25.101 Release 9)
  - Reference Sensitivity test case
- From 3GPP requirements to TX System specifications WCDMA Band I (TS 25.101 Release 9)

# Reference Sensitivity test case (♠)

- Parameters
  - Channel bandwidth
  - Sensitivity requirement
  - BER requirement

## 7.3.1 Minimum requirement

The BER shall not exceed 0.001 for the parameters specified in Table 7.2.

**Table 7.2: Test parameters for reference sensitivity, minimum requirement.**

| Operating Band | Unit | DPCH_Ec <REFSENS> | <REF$\hat{I}_{or}$> |
|---|---|---|---|
| I | dBm/3.84 MHz | -117 | -106.7 |
| II | dBm/3.84 MHz | -115 | -104.7 |
| III | dBm/3.84 MHz | -114 | -103.7 |
| IV | dBm/3.84 MHz | -117 | -106.7 |
| V | dBm/3.84 MHz | -115 | -104.7 |
| VI | dBm/3.84 MHz | -117 | -106.7 |
| VII | dBm/3.84 MHz | -115 | -104.7 |
| VIII | dBm/3.84 MHz | -114 | -103.7 |
| IX | dBm/3.84 MHz | -116 | -105.7 |
| X | dBm/3.84 MHz | -117 | -106.7 |
| XI | dBm/3.84 MHz | -117 | -106.7 |
| XII | dBm/3.84 MHz | -114 | -103.7 |
| XIII | dBm/3.84 MHz | -114 | -103.7 |
| XIV | dBm/3.84 MHz | -114 | -103.7 |
| XIX | dBm/3.84 MHz | -117 | -106.7 |
| XXI | dBm/3.84 MHz | -117 | -106.7 |

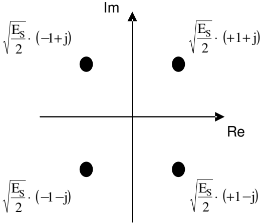
# Standard target SNR (♠)

- Steps to calculate target SNR for sensitivity and low data-rates test cases:
  - Identify the type of modulation
  - Find out what Bit Error Rate (BER) is required
  - Estimate the Signal-to-Noise-Ratio (SNR) needed to be compliant
- As defined in TS 25.101 & 25.213:
  - WCDMA RX basic modulation is Quadrature Phase-Shift Keying (QPSK)
  - « The BER shall not exceed 0.001 ... »

# Standard target SNR (♠)

- A mathematical approach is used to calculate the target SNR:
  - Assume a Gaussian distribution of the channel noise
  - All symbols are equally likely
  - Calculate first Symbol Error Rate (SER) and then deduce BER

$$P(x) = \frac{1}{\sqrt{\pi N_0}} \mathrm{e}^{\frac{x^2}{N_0}} \quad (191)$$



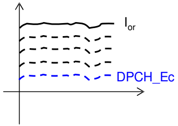

- Recall previous section on BER
- BER of 0.001 is achieved only when $\frac{E_b}{N_0} \geq 7.3\,\mathrm{dB}$

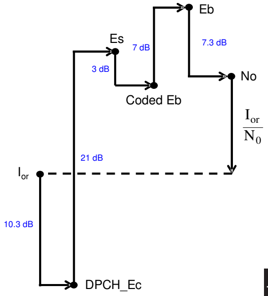
# From $E_b/N_0$ to physical SNR (♠)

- The WCDMA UE RX signal is a spread-spectrum modulation scheme (TS 25.213)
- The spreading factor is equal to 128
- The useful channel $DPCH_{\mathrm{Ec}}$(*) is 10.3 dB below the composite signal $I_{\mathrm{or}}$
- The receiver sees only the composite signal

$$\frac{I_{\mathrm{or}}}{N_0} = I_{\mathrm{or}} + \frac{DPCH_{\mathrm{Ec}}}{I_{\mathrm{or}}} + 10 \times \log_{10}\left(\frac{SF}{M}\right) + BCG - \frac{E_b}{N_0} \quad (192)$$

- The physical SNR is negative !
- Rough estimation for RF band I: $-7.4$ dB




(*): Dedicated Physical Channel ; see appendix slide 24, 25, 26, 27, 28 for other system derivations

# Outline

From Wireless Standard Requirements to System Level Spec

- Introduction
- Example : From 3GPP requirements to RX System specifications WCDMA Band I (TS 25.101 Release 9)
- From 3GPP requirements to TX System specifications WCDMA Band I (TS 25.101 Release 9)

# Outline

From Wireless Standard Requirements to System Level Spec

- Introduction
- Example : From 3GPP requirements to RX System specifications WCDMA Band I (TS 25.101 Release 9)
- From 3GPP requirements to TX System specifications WCDMA Band I (TS 25.101 Release 9)
  - Out-of-band emissions test cases
  - Modulation accuracy test case

# ACLR

- 3GPP defines ACLR as the ratio of the transmitted channel to the adjacent WCDMA channel
  - Spectral re-growth of the transmitter must fulfill the requirements integrated
    - in measurement bandwidth of 3.84 MHz
    - around channels at frequency offset of $\pm 5$ & $\pm 10$ MHz for the 1st and 2nd adjacents respectively
    - However absolute limit is $-50$ dBm

Table 6.11: UE ACLR

| Power Class | Adjacent channel frequency relative to assigned channel frequency | ACLR limit |
|---|---|---|
| 3 | + 5 MHz or - 5 MHz | 33 dB |
| 3 | + 10 MHz or - 10 MHz | 43 dB |
| 4 | + 5 MHz or - 5 MHz | 33 dB |
| 4 | + 10 MHz or -10 MHz | 43 dB |

# ACLR

- One rule: relax PA budget as power range is 20 dB higher than RFIC !!!
- Placing $ACLR_{RFIC}$ above $ACLR_{PAM}$ is a good partition budget
  $\rightarrow ACLR_{RFIC} = ACLR_{PAM} + 6\,\text{dB}$

| ACLR ± 5MHz | Ptx = +24dBm | RF bands | |
|---|---|---|---|
| | | I | % |
| RFIC | dB | 41.0 | 20.0 |
| PA | dB | 35.0 | 80.0 |
| Total | dB | 34.0 | |
| Required in 3GPP | dB | 33 | |
| Margin to 3GPP | dB | 1.0 | |

| **ACLR ± 10MHz** | **Ptx = +24dBm** | **RF bands** | |
|---|---|---|---|
| | | **I** | % |
| **RFIC** | **dB** | 51.0 | 20.0 |
| **PA** | **dB** | 45.0 | 80.0 |
| **Total** | **dB** | 44.0 | |
| **Required in 3GPP** | **dB** | 43 | |
| **Margin to 3GPP** | **dB** | 1.0 | |

---

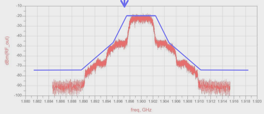
Recall RFIC overview on slide 8

# Spectrum Emission Mask

- Spectral re-growth of transmitter must fulfill the following mask requirements
  - Measurement bandwidth is variable $\rightarrow$ from 30 kHz to 1 MHz
  - Bandwidth range is $2.5 \leq |F_{offset}| \leq 12.5\,\mathrm{MHz}$
- Only System simulation can check SEM requirements

Table 6.10: Spectrum Emission Mask Requirement

| Δf in MHz (Note 1) | Minimum requirement (Note 2) | | Measurement bandwidth |
|---|---|---|---|
| | Relative requirement | Absolute requirement | |
| 2.5 - 3.5 | $\left\{-35-15\cdot\left(\frac{\Delta f}{MHz}-2.5\right)\right\}dBc$ | -71.1 dBm | 30 kHz (Note 3) |
| 3.5 - 7.5 | $\left\{-35-1\cdot\left(\frac{\Delta f}{MHz}-3.5\right)\right\}dBc$ | -55.8 dBm | 1 MHz (Note 4) |
| 7.5 - 8.5 | $\left\{-39-10\cdot\left(\frac{\Delta f}{MHz}-7.5\right)\right\}dBc$ | -55.8 dBm | 1 MHz (Note 4) |
| 8.5 - 12.5 MHz | -49 dBc | -55.8 dBm | 1 MHz (Note 4) |



# Outline

# Modulation quality

- 3GPP defines the EVM requirements in terms of:
  - supported modulations
  - output power of transmitter
  - measurement period 2560 chips $\sim$ 666.7 µs
- 3GPP value is $\sim$ 15%, however state-of-the-art implementation target 10%
  - Budget is split equally between PAM and RFIC as all TX chain blocks contributes in EVM !!!
  - Digital, DAC, ABB filter, IQ modulator, PPA and phase noise from synthesizer

Table 6.15: Parameters for Error Vector Magnitude/Peak Code Domain Error

| Parameter | | Unit | Level |
|---|---|---|---|
| UE Output Power, no 16QAM | | dBm | ≥ -20 |
| UE Output Power, 16QAM | | dBm | ≥ -30 |
| Operating conditions | | | Normal conditions |
| Power control step size | | dB | 1 |
| Measurement period (Note 1) | PRACH | Chips | 3904 |
| | Any DPCH | | From 1280 to 2560 (Note 2) |
| Note 1: Less any 25µs transient periods | | | |
| Note 2: The longest period over which the nominal power remains constant | | | |

| Modulation accuracy EVM | Ptx = +24 dBm | RF bands | |
|---|---|---|---|
| | | I | % |
| RFIC | % | 6 | 50 |
| PA | % | 6 | 50 |
| Total | % | 8.5 | |
| Required in 3GPP | % | 17.5 | |
| Required in state-of-the-art | % | 10 | |
| Margin to 3GPP | dB | 6.0 | |
| Margin to state-of-the-art | dB | 1.5 | |

# Outline

Conclusion

# Conclusion

- RF transceivers in today's state-of-the-art solutions are driven mainly by
  - Power consumption
  - Area
  - Fabrication cost
- RF System engineer responsibility $\rightarrow$ make sure it is in-line with the standard requirements
- System partition is done at two levels
  - Top-level $\rightarrow$ deriving specifications from standard requirements
  - Block-level $\rightarrow$ optimizing the partition between blocks to make circuit design easier
- Wireless communication standards are evolving quickly and so does RF System engineering

# Appendix slides

# Outline

App.: Basic concepts in RF design (short recall)

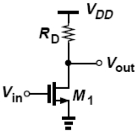
App.: Communication theory : Introduction

App.: From Wireless Standard Requirements to System Level Spec

# Considered nonlinearity class in this lecture (♣)

- Memoryless and Static System
  - Linear : $y(t) = ax(t)$
  - Nonlinear : $y(t) = \alpha_0 + \alpha_1 x(t) + \alpha_2 x^2(t) + \alpha_3 x^3(t) + \ldots$
- The IO characteristic of a memoryless nonlinear system can be approximated with a polynomial

## Example




$$\begin{aligned} V_{out} &= V_{DD} - I_D R_D && (193) \\ &= V_{DD} - \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{in} - V_{TH})^2 R_D && (194) \end{aligned}$$

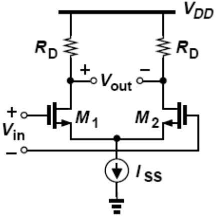
- In this idealized case, the circuit displays only second-order nonlinearity

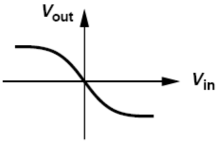
# Considered nonlinearity class in this lecture (♣)

## Example (continued)

- Square-law MOS transistors operating in saturation

$$V_{out} = -\frac{1}{2}\mu_n C_{ox}\frac{W}{L}R_D V_{in}\sqrt{\frac{4I_{SS}}{\mu_n C_{ox}\frac{W}{L}} - V_{in}^2} \qquad (195)$$

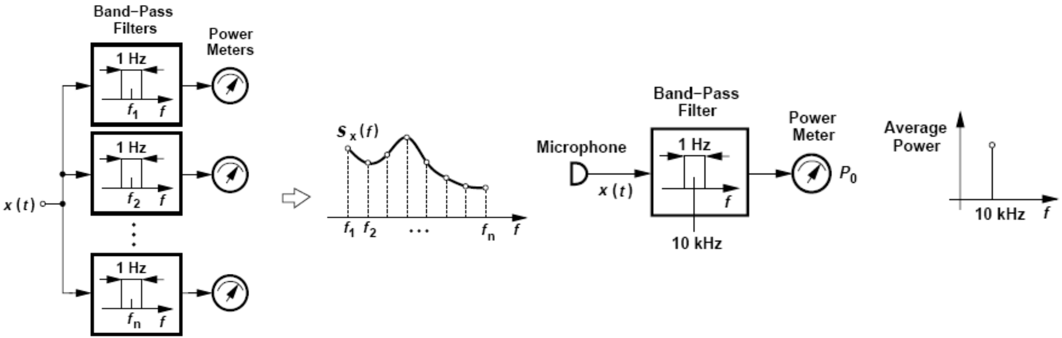



- If the differential input is small
  - It can be shown:

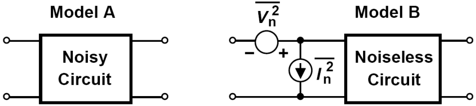
$$V_{out} \approx -\sqrt{\mu_n C_{ox}\frac{W}{L}I_{SS}}R_D V_{in} + \frac{\left(\mu_n C_{ox}\frac{\omega}{2}\right)^{\frac{3}{2}}}{8\sqrt{I_{SS}}}R_D V_{in}^3 \qquad (196)$$

# Measurement of Noise Spectrum (♣)




- To measure signal's frequency content at 10 kHz, we need to filter out the remainder of the spectrum and measure the average power of the 10-kHz component.

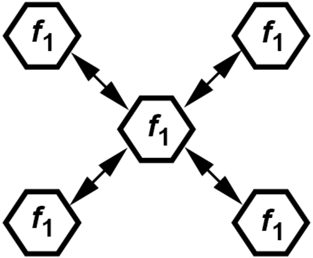
# Representation of Noise in Circuits: Input-Referred Noise (♣)




- Voltage source: short the input port of models A and B and equate their output noise voltage
- Current source: leave the input ports open and equate the output noise voltage

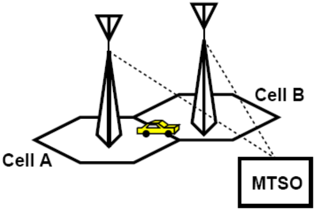
# Outline

# Co-Channel Interference (♣)




- CCI: depends on the ratio of the distance between two co-channel cells to the cell radius, independent of the transmitted power
- Given by the frequency reuse plan, this ratio is approximately equal to 4.6 for the 7-cell pattern.

# Hand-off (♣)




- When a mobile unit roams from cell A to cell B, since adjacent cells do not use the same group of frequencies, the channel must also change.
- Second-generation cellular systems allow the mobile unit to measure the received signal level from different base stations, thus performing hand-off when the path to the second base station has sufficiently low loss

# Diversity & Interleaving (♣)

- Diversity
  - Space Diversity or Antenna Diversity employs two or more antennas spaced apart by a significant fraction of the wavelength so as to achieve a higher probability of receiving a nonfaded signal
  - Frequency Diversity refers to the case where multiple carrier frequencies are used
  - Time Diversity: the data is transmitted or received more than once to overcome short-term fading
- Interleaving
  - Errors occur in clusters of bits
  - To lower the effect of these errors, the baseband bit stream in the transmitter undergoes “interleaving” before modulation

# FDD vs. TDD

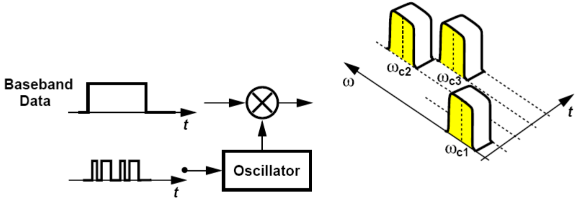
- Features of FDD
  - FDD: components of the transmitted signal that leak into the receive band are attenuated by typically only about 50 dB.
  - FDD: owing to the trade-off between the loss and the quality factor of filters, the loss of the duplexer is typically quite higher than that of a TDD switch.
  - FDD: spectral leakage to adjacent channels in the transmitter output

- Features of TDD
  - TDD: two paths (RX,TX) do not interfere because the transmitter is turned off during reception
  - TDD: allows direct (peer-to-peer) communication between two transceivers
  - TDD: strong signals generated by all of the nearby mobile transmitters fall in the receive band, thus desensitizing the receiver.

# TDMA Features: Compared with FDMA (♣)

- TDMA: power amplifier can be turned off during the time of the frame out of assigned time slot
- TDMA: digitized speech can be compressed in time by a large factor, smaller required bandwidth.
- TDMA: even with FDD, TDMA bursts can be timed so the receive and transmit paths are never enabled simultaneously
- TDMA: more complex due to A/D conversion, digital modulation, time slot and frame synchronization, etc.

# Frequency-Hopping CDMA




- Can be viewed as FDMA with pseudo-random channel allocation.
- Occasional overlap of the spectra raises the probability of error

# Example of Wideband CDMA Receiver Requirements (♣)

- Estimate the required $P_{\mathrm{1dB}}$ of a WCDMA receiver satisfying the in-band test of figure above. (continuation)
  - For a sinusoid $A_1 \cos \omega_1 t$ and an amplitude-modulated blocker $A_2(1 + m \cos \omega_m t) \cos \omega_2 t$, cross modulation appears as

$$y(t) = \left[\alpha_1 A_1 + \frac{3}{2}\alpha_3 A_1 A_2^2 \left(1 + \frac{m^2}{2} + \frac{m^2}{2} \cos 2\omega_m t + 2m \cos \omega_m t\right)\right] \cos \omega_1 t + \cdots \tag{197}$$

## Example of Wideband CDMA Receiver Requirements (♣)

- Estimate the required $P_{\mathrm{1dB}}$ of a WCDMA receiver satisfying the in-band test of figure above. (continuation)
  - For the case at hand, both channels contain modulation and we consider

$$A_1\left(1+m\cos\omega_{m_1}t\right)\cos\omega_1 t \text{ and } A_2\left(1+m\cos\omega_{m_2}t\right)\cos\omega_2 t \text{ and } m^2/2 \ll 2m \tag{198}$$

$$y(t) = \Big[\alpha_1 A_1\left(1+m\cos\omega_{m_1}t\right) + \frac{3}{2}\alpha_3 A_1\left(1+m\cos\omega_{m_1}t\right)A_2^2\times\left(1+2m\cos\omega_{m_2}t\right)\Big]\cos\omega_1 t+\dots \tag{199}$$

$$= \Big[\alpha_1 A_1\left(1+m\cos\omega_{m_1}t\right) + \frac{3}{2}\alpha_3 A_1 A_2^2\left(1+m\cos\omega_{m_1}t+2m\cos\omega_{m_2}t+2m^2\cos\omega_{m1}t\cos\omega_{m_2}t\right)\Big]\cos\omega_1 t+\dots \tag{200}$$

# Example of Wideband CDMA Receiver Requirements (♣)

- Estimate the required $P_{1\mathrm{dB}}$ of a WCDMA receiver satisfying the in-band test of figure above. (continuation)

  - For the corruption to be negligible, the average power of the second term in the square brackets must remain much less than that of the first

$$\frac{\left(\frac{3}{2}\alpha_3 A_1 A_2^2\right)^2 \left(1+m^2+4m^2+4m^4\right)}{(\alpha_1 A_1)^2 \left(1+m^2\right)} \ll 1 \tag{201}$$

  - Setting this ratio to $-15\,\mathrm{dB}$ $(= 0.0316)$ and neglecting the powers of $m$:

$$\frac{\frac{3}{2}|\alpha_3| A_2^2}{|\alpha_1|} = 0.178 \tag{202}$$

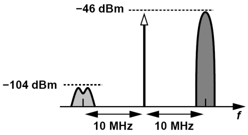
  - Since $A_{1\mathrm{dB}} = \sqrt{0.145|\alpha_1/\alpha_3|}$ (signal amplitude at 1 dB compression):

$$A_{1\mathrm{dB}} = 1.1 A_2 \tag{203}$$

the input compression point must exceed $A_2$ $(= -44\,\mathrm{dBm})$ by about $1\,\mathrm{dB}$. Thus, compression is slightly more dominant than cross modulation in this test.

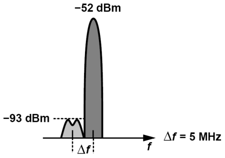
# Wideband CDMA Receiver Requirements: Intermodulation Test & Adjacent Channel Test (♣)

- IMT-2000 intermodulation test:




- A tone and a modulated signal each at $-46\,\mathrm{dBm}$ applied in the adjacent and alternate adjacent channels, desired signal at $-104\,\mathrm{dBm}$

- IMT-2000 receiver adjacent-channel test:



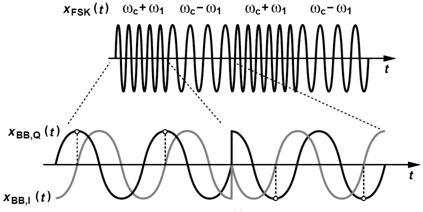

- Desired signal $-93\,\mathrm{dBm}$, adjacent channel $-52\,\mathrm{dBm}$

# Example of I/Q Mismatch (♣)

- An FSK signal is applied to a direct-conversion receiver. Plot the baseband waveforms and determine the effect of I/Q mismatch.
- We express the FSK signal as $x_{FSK}(t) = A_0 \cos[(\omega_c + a\omega_1)t]$, where $a = \pm 1$ represents the binary information; i.e., the frequency of the carrier swings by $+\omega_1$ or $-\omega_1$. Upon multiplication by the quadrature phases of the LO, the signal produces the following baseband components:

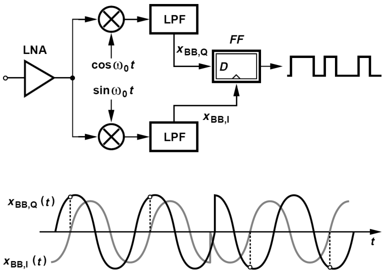
$$\begin{cases} x_{BB,I}(t) & = -A_1 \cos a\omega_1 t \\ x_{BB,Q}(t) & = +A_1 \sin a\omega_1 t \end{cases} \tag{204}$$

- if the carrier frequency is equal to $\omega_c + \omega_1$ (i.e., $a = +1$), then the rising edges of $x_{BB,I}(t)$ coincide with the positive peaks of $x_{BB,Q}(t)$. Conversely, if the carrier frequency is equal to $\omega_c - \omega_1$, then the rising edges of $x_{BB,I}(t)$ coincide with the negative peaks of $x_{BB,Q}(t)$.



# Example of I/Q Mismatch (♣)

- An FSK signal is applied to a direct-conversion receiver. Plot the baseband waveforms and determine the effect of I/Q mismatch.
- Thus, the binary information is detected if $x_{BB,I}(t)$ simply samples $x_{BB,Q}(t)$, e.g., by means of a D flipflop.
- FSK may tolerate large I/Q mismatches: amplitude mismatch proves benign so long as the smaller output does not suffer from degraded SNR, and phase mismatch is tolerable so long as $x_{BB,I}(t)$ samples the correct polarity of $x_{BB,Q}(t)$. Of course, as the phase mismatch approaches 90°, the additive noise in the receive chain introduces errors.



# Outline

# Receiver: Nonlinearity (♣)

- The IP3 corresponding to the 1-dB compression point is satisfied if compression by the desired signal is avoided. The IP3 arising from adjacent channel specifications must be satisfied while the desired signal is only 3 dB above the reference sensitivity.

# Receiver: AGC Range (♣)

- The receiver must automatically control its gain if the received signal level varies considerably.
- The challenge is to realize this gain range while maintaining a noise figure of about $10\,\mathrm{dB}$ and an IIP3 of about $-40\,\mathrm{dBm}$.

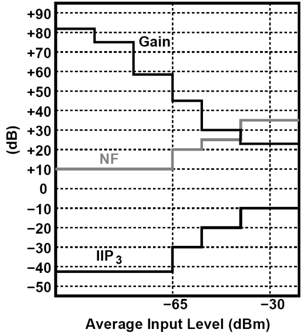
## Example: AGC range of an 11a/g receiver

Determine the AGC range of an 11a/g receiver so as to accommodate the rate-dependent sensitivities.

- the input signal level varies from $-82\,\mathrm{dBm}$ to $-65\,\mathrm{dBm}$, requiring a gain of $86\,\mathrm{dB}$ to $69\,\mathrm{dB}$ so as to reach $1\,\mathrm{Vpp}$ at the ADC input. However, a 64QAM signal exhibits a peak-to-average ratio of about $9\,\mathrm{dB}$; also, baseband pulse shaping to meet the TX mask also creates 1 to $2\,\mathrm{dB}$ of additional envelope variation. Thus, an average input level of $-65\,\mathrm{dBm}$ in fact may occasionally approach a peak of $-65\,\mathrm{dBm} + 11\,\mathrm{dB} = -54\,\mathrm{dBm}$. It is desirable that the ADC digitize this peak without clipping. That is, for a $-65$-$\mathrm{dBm}$ 64QAM input, the RX gain must be around $58\,\mathrm{dB}$.

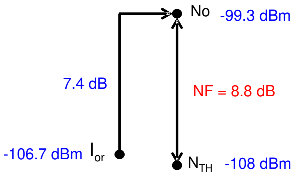
# ACG Range: Required RX Gain Switching and NF and IIP3 Variations (♣)

- The receiver gain range is also determined by the maximum allowable desired input level ($-30\,\mathrm{dBm}$). The baseband ADC preferably avoids clipping the peaks of the waveforms.
- The actual number of steps chosen here depends on the design of the RX building blocks and may need to be quite larger than that depicted



# RFSS noise figure specification (♠)

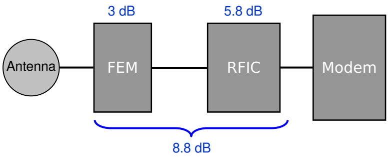
- Steps to calculate NF specification
  - Thermal noise at the antenna: physical absolute minimum noise in the frequency bandwidth of interest
  - The sensitivity power level
  - Physical SNR required to meet 3GPP requirement
- Diagram for NF calculation




RFSS: RF Sub-system (slide 7)

# RFIC noise figure specification (♠)

- Main difference between RFSS and RFIC:
  - FEM contributes largely in the noise figure budget
  - Typical datasheets from vendors show that around 3 dB of power is lost




- WCDMA RFIC NF specification for band I is:

$$NF_{spec} = 5.8\,\mathrm{dB} \tag{205}$$

## From NF to gain and ADC specifications (♠)

- Steps to calculate gain and ADC specifications:
  - ADC full-scale is assumed equal to 1 Vpp
  - NF partition: 1.2 dB to TX leakage degradation, 0.3 dB to ADC, 4 dB to RX and 0.3 dB of margin

$$NF_{TX} = NF_{spec} - (NF_{RX} + NF_{ADC}) - 0.3\,\text{dB} \qquad (206)$$

  - We know that WCDMA bandwidth is 3.84 MHz $\rightarrow N_{th} = -108\,\text{dBm}$
  - ADC power consumption is fixed by the platform engineer → limited range of possible DR
  - Using equation $N_{ADC} = FS_{ADC} - DR \leq N_{th} + G_{RX} + NF_{RX} - 11$ we deduce:

$$G_{RX,spec} = 35\,\text{dBV}_{\text{RMS}}/\text{dBm} \quad (207) \qquad DR_{spec} = 74\,\text{dB} \quad (208)$$

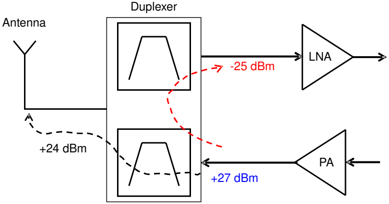
- The specified DR corresponds to $ENOB = 11.5\,\text{bits}$ → feasible in current state-of-the-art design

As indicated in the title slide, the content of this page was produced by R. Mina and some elements can only be explained by the original author. These elements are currently under review for making it self-explanatory.

## Calculating TX leakage at RX input (♠)

- Steps to calculate TX leakage:
  - Maximum output power at antenna $\rightarrow$ 24 dBm ($\approx$ 250 mW TS 25.101 Release 9)
  - Transmit insertion loss inside duplexer $\rightarrow$ 3 dB (FEM vendors data)
  - Receive to transmit isolation inside duplexer $\rightarrow$ 52 dB (FEM vendors data)

$$P_{TX,leak} = P_{TX,ant} + IL_{TX} - ISO_{TX-RX} \tag{209}$$



# From NF to IIP2 specification (♠)

- Steps to calculate IIP2 specification:
  - Determine TX leakage power at LNA input $\rightarrow -25\,\mathrm{dBm}$
  - Calculate overall mitigation factor for specific TX modulation $\rightarrow -9\,\mathrm{dB}$ (simulation done in ADS)
  - Calculate RX noise at antenna $\rightarrow N_{RX,ant} = N_{th} + IL_{RX} + NF_{RX} + NF_{ADC}$
  - Assume $1.2\,\mathrm{dB}$ NF degradation is split equally between self & reciprocal mixing $\rightarrow 8\,\mathrm{dB}$ margin to $N_{RX,ant}$

$$\begin{aligned} IIP_{2,spec} &= 2P_{in,2} - 6\,\mathrm{dB} + M_f - (N_{RX,ant} - 8\,\mathrm{dB} - IL_{RX}) && (210) \\ &= 47\,\mathrm{dBm} && (211) \end{aligned}$$