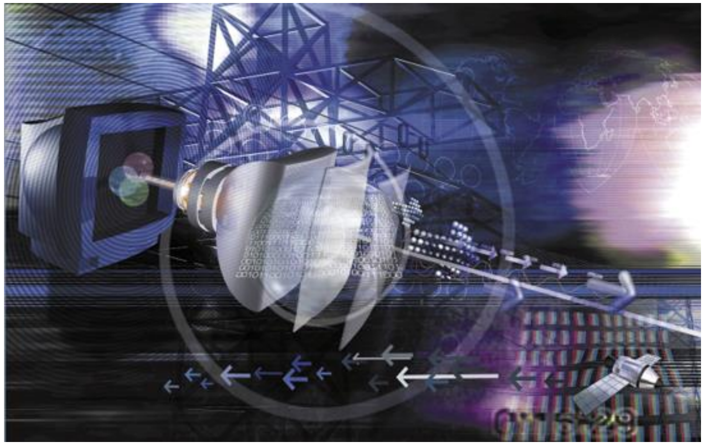
Communication Circuits Design

Academic year 2018/2019 – Semester 2 – Week 4
Lecture 4.1-2: RF Receivers & Transceivers

Francesco Fioranelli, Wasim Ahmad, Faisal Tariq

Outline



- **RF receivers: main topologies**
- **Key metrics of quality**
- **Brief info on transmitters**
- **Phase noise recall**
- **Real world examples on chip and on systems**

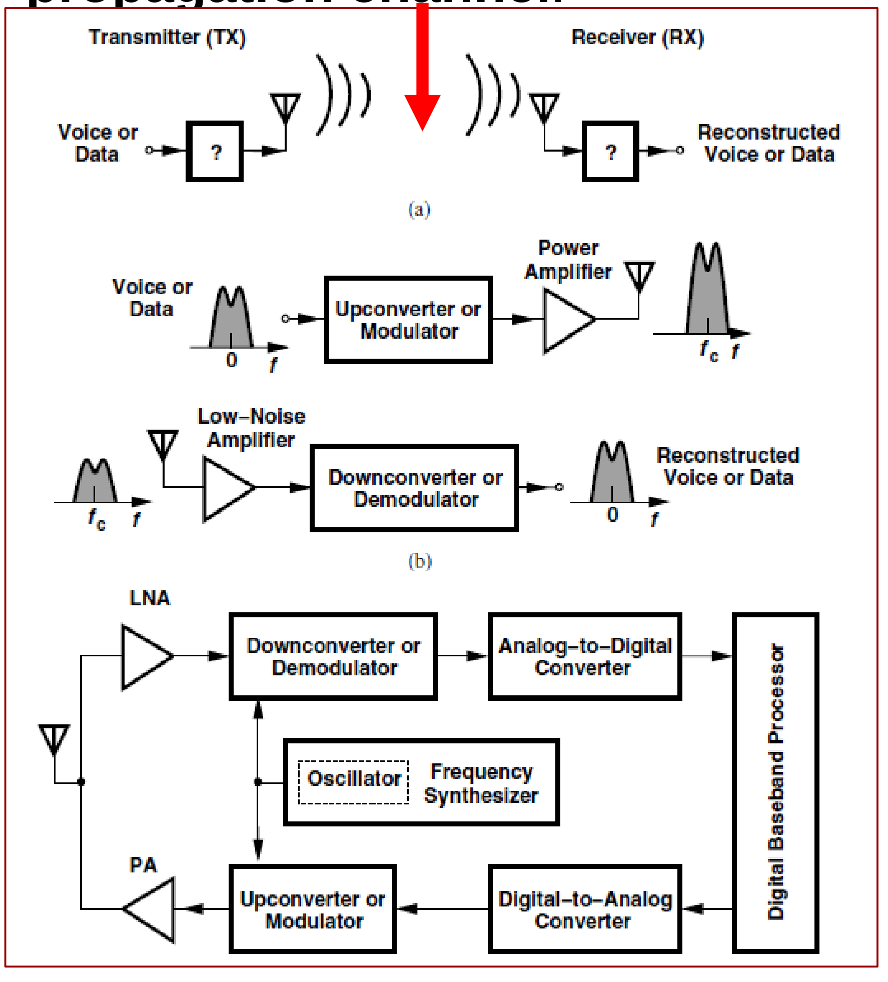
References:

•*R. Sobot, "Wireless Communication Electronics", Springer, at UoG Library online – Chapter 13 (mostly taken from this)*

•*B. Razavi, "RF Microelectronics", Prentice Hall, 2nd ed. – Chapter 4 (note that this is very detailed, beyond what is expected in this course, but nevertheless interesting)*

Communication systems

Let us look again at a generic communication system made of a **transmitter** and a **receiver**, with the data travelling through a **propagation channel**.



<u>Transmitter and receiver</u> are made of the individual blocks we have already studied (mixers, oscillators, LC tuning & matching networks, amplifiers, antennas).

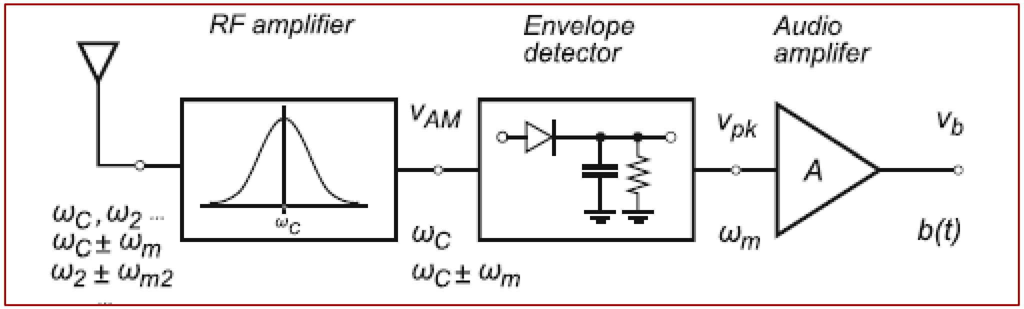
And the **<u>channel</u>**? Modelling the effects of the channel on the communication signals can be a course on its own! Channels can be copper wires, optical fibres, wireless channel (air).

For our simplified view in this course we just *consider the channel as an attenuator block reducing the level of the useful signal of xdB*. Recall Friis formula for ***dependence on λ***

TRF receiver

The oldest and simplest implementation of radio receiver is the **Tuned Radio Frequency (TRF) receiver**.

TRF = antenna -> RF amplifier -> envelope detector -> audio amplifier



Each signal (each radio station) needs its RF amplifier tuned to its carrier frequency ωc, and equally any subsequent stage or block. **Issues**:

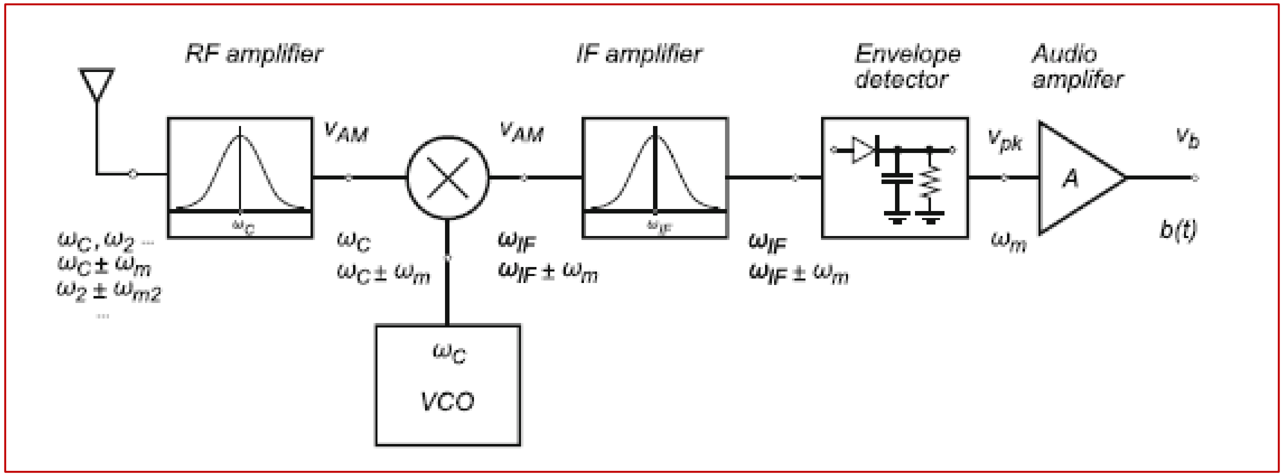
- Envelope detector works only at relatively low HF frequencies (tens of MHz max), so ωc must be low
- RF amplifier gain changes with frequency, so different signals are amplified with different gain values
- Q factor of RF amplifier typically low, hence many frequency components will enter the envelope detector reducing the overall SNR of the output -> as the number of signals increases (more radio users), this problem of *selectivity* becomes worse

TRF to heterodyne

If there are many signals being broadcast at the same time, the Q factor to separate them from each other may become too difficult/expensive to achieve and maintain. You need new receiver topologies.

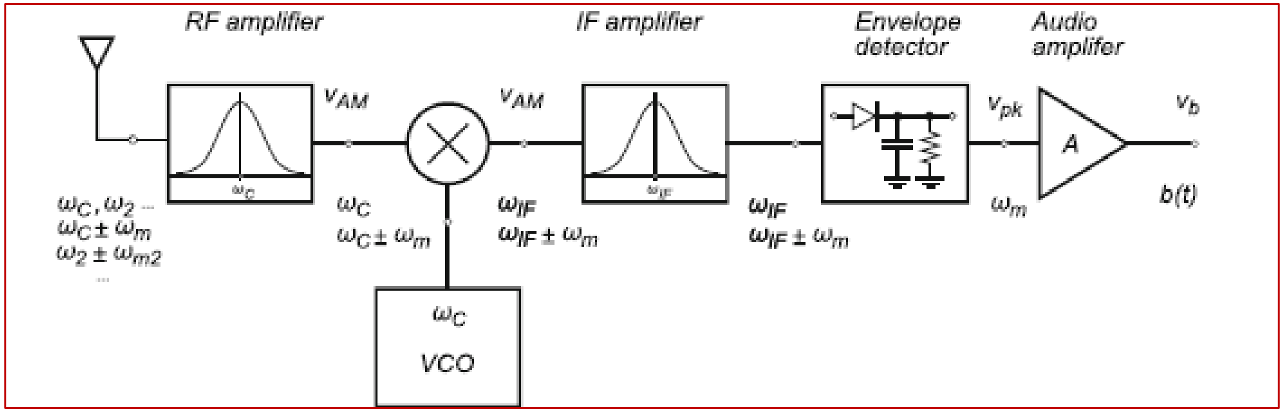
Most modern RF receivers are **heterodyne receivers**, where the received RF signal is down-converted via mixing with an LO to an IF frequency signal before being detected. This down-conversion process can also be done in two steps (2 LO and mixing stages), often referred to as **super-heterodyne receivers**.

Advantage: the IF stage does not change for different RF signals, but all of them are down-converted to the same IF by simultaneous tuning of RF amplifier and LO (voltage controlled oscillator).



Heterodyne receivers & metrics

Before we move on, can you make sure you understand how the *spectrum* of the different signals in this receiver is and the effect of mixing?
If not, this is the moment to ask.



Receivers can be characterised by some common metrics:

- **Selectivity**: accounts for the minimum separation between desired RF signal and closer interference signal so that the first can be still safely received
- **Sensitivity**: minimum amplitude of desired RF signal that can be still safely received with an acceptable SNR
- **Dynamic Range**: amplitude ratio of strongest and weakest RF signals that the receiver can still safely receive

Metrics and non-linearity

Note the expression "*safely receive*" in the metrics – many modern communication systems are **digital** today, which means that the received, down-converted IF signal is digitised to generate a vector of bits. A quality metric in this case is also the ***BER*** (bit error rate) for a given application (how many bits were received wrongly with respect to those transmitted).

As in this course we work mainly on analogue systems, we focus our metrics on the level/amplitude of the signals.

Before providing some formulae to quantify these metrics, we need to look at the ***non-linear effects*** that any receiver will cause (remember, you need non-linear elements to make mixers and amplifiers).
In general if we consider the receiver as a single block with an input x(t) and output y(t), a non linear relation is such that the output depends on the input through square, cubic, fourth-power elements and so on

$$y(t) = a_0 + a_1 x(t) + a_2 x^2(t) + a_3 x^3(t) + \cdots$$

Harmonic distortion

The first effect of non-linearity is **harmonic distortion**. Assume that a single tone $x(t) = Bcos(\omega t)$ is used as input and that a_0 coefficient is zero.

$$\boxed{\begin{aligned} y(t) &= a_1 B\cos\omega t + a_2 B^2\cos^2\omega t + a_3 B^3\cos^3\omega t + \cdots \\ &= a_1 B\cos\omega t + \frac{a_2 B^2}{2}(1+\cos 2\omega t) + \frac{a_3 B^3}{4}(3\cos\omega t + \cos 3\omega t) + \cdots \\ &= \frac{a_2 B^2}{2} + \left(a_1 B + \frac{3a_3 B^3}{4}\right)\cos\omega t + \frac{a_2 B^2}{2}\cos 2\omega t + \frac{a_3 B^3}{4}\cos 3\omega t + \cdots \\ &= b_0 + b_1\cos\omega t + b_2\cos 2\omega t + b_3\cos 3\omega t + \cdots, \end{aligned}}$$

Key point: the output y(t) contains higher-order harmonics of the original input signal, 2ω, 3ω, that did not exist in the input but have been created by the non-linear effects.

Typically this effect is measured by Total Harmonic Distortion THD, which depends on the amplitude of the individual harmonic components with respect to the first (desired) harmonic.

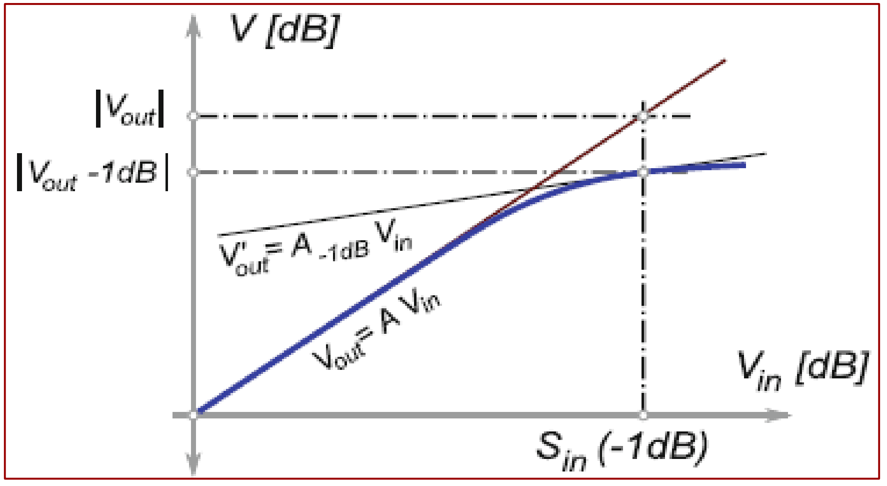
$$THD = \sqrt{D_2^2 + D_3^2 + \cdots} \text{ where } D_2 = \frac{b_2}{b_1}, D_3 = \frac{b_3}{b_1}, \ldots$$

Gain compression

This property is typical of non-linear amplifier circuits. At first the relation between output and input power is linear through the gain A so that $P_{out} = AP_{in}$. However, the output cannot grow indefinitely (for example limitations due to finite DC power supply) and at some point will stop growing.

The **1dB compression point** is the input signal power in dB corresponding to the gain where the output signal power is 1dB lower than the ideal linear model.

<u>Experimentally</u>, one can draw the graph below, the blue curve, and find the 1dB point with respect to the ideal model in red.



Gain compression

Analytically, one can go back to the formula seen for harmonic distortion and compare the amplitude coefficients of the desired harmonic ω for the ideal linear case ($a_1 B$), and for the actual non-linearity case ($a_1 B + \frac{3a_3 B^3}{4}$).

$$\begin{aligned}
y(t) &= a_1 B\cos\omega t + a_2 B^2\cos^2\omega t + a_3 B^3\cos^3\omega t + \cdots \\
&= a_1 B\cos\omega t + \frac{a_2 B^2}{2}(1+\cos 2\omega t) + \frac{a_3 B^3}{4}(3\cos\omega t + \cos 3\omega t) + \cdots \\
&= \frac{a_2 B^2}{2} + \left(a_1 B + \frac{3a_3 B^3}{4}\right)\cos\omega t + \frac{a_2 B^2}{2}\cos 2\omega t + \frac{a_3 B^3}{4}\cos 3\omega t + \cdots \\
&= b_0 + b_1\cos\omega t + b_2\cos 2\omega t + b_3\cos 3\omega t + \cdots,
\end{aligned}$$

If the ratio of the non-linear / linear cases is calculated in dB and imposed to be -1dB (1dB below), one gets the relations below

$$-1\,\mathrm{dB} = 20\log\left(1 + \frac{3a_3 B^2}{4a_1}\right) \rightarrow B(-1\,\mathrm{dB}) = \sqrt{0.145\left|\frac{a_1}{a_3}\right|} \rightarrow S_{\mathrm{in}}(-1\,\mathrm{dB}) = 20\log[B(-1\,\mathrm{dB})]\ \ \mathrm{dB}.$$

Essentially the 1dB compression point of the first harmonic (a_1) is influenced by the third harmonic (a_3), which is an interesting result. To use this formula in an experiment, you need to know or measure the value of the coefficients a_1 a_3

Intermodulation

If we consider two different signals ωa and ωb at the input of a non-linear receiver, we expect to see harmonics of each input frequency, but there will also be frequencies from **inter-modulation.**

Analytically, if the input is tone $x(t) = B_1 cos(\omega_a t) + B_2 cos(\omega_b t)$ then

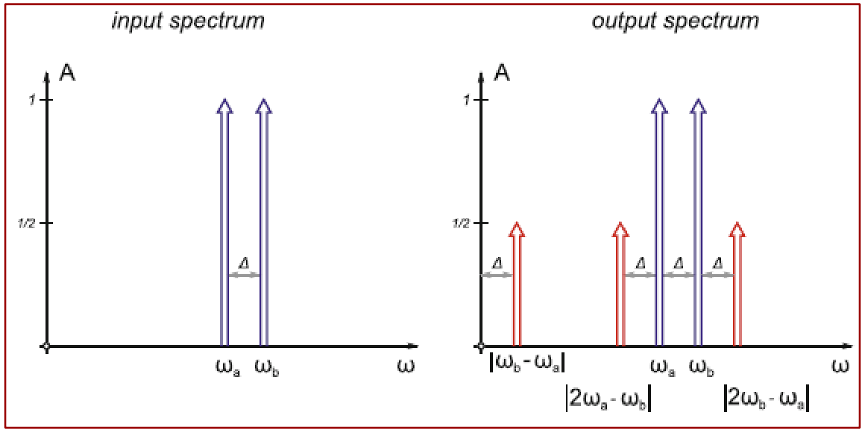
$$y(t) = a_1 \left(B_1 \cos\omega_a t + B_2 \cos\omega_b t\right) + a_2 \left(B_1 \cos\omega_a t + B_2 \cos\omega_b t\right)^2 + a_3 \left(B_1 \cos\omega_a t + B_2 \cos\omega_b t\right)^3 + \cdots$$

$$\begin{aligned}
y(t) = {} & \frac{a_2(B_1^2 + B_2^2)}{2} && \textit{(DC term)} \\
& + \left(a_1 B_1 + \frac{3}{4} a_3 B_1^3 + \frac{3}{2} a_3 B_1 B_2^2\right) \cos\omega_a t && \textit{(fundamental terms)} \\
& + \left(a_1 B_2 + \frac{3}{4} a_3 B_2^3 + \frac{3}{2} a_3 B_2 B_1^2\right) \cos\omega_b t \\
& + \frac{a_2}{2}\left(B_1^2 \cos 2\omega_a t + B_2^2 \cos 2\omega_b t\right) && \textit{(second-order terms)} \\
& + a_2 B_1 B_2 \left[\cos(\omega_a + \omega_b)t + \cos|\omega_a - \omega_b|t\right] \\
& + \frac{a_3}{4}\left(B_1^3 \cos 3\omega_a t + B_2^3 \cos 3\omega_b t\right) && \textit{(third-order terms)} \\
& + \frac{3a_3}{4}\left\{B_1^2 B_2 \left[\cos(2\omega_a + \omega_b)t + \cos(2\omega_a - \omega_b)t\right]\right. \\
& \qquad \left. + B_1 B_2^2 \left[\cos(2\omega_b + \omega_a)t + \cos(2\omega_b - \omega_a)t\right]\right\},
\end{aligned}$$

Besides the expected harmonics of the input frequencies, there are sum&difference components (which you know from mixers' analysis) and 3rd order components $2\omega_a \pm \omega_b$ and $2\omega_b \pm \omega_a$

Intermodulation

These 3rd order components can be a problem if the original input tones ωa and ωb are close to each other, as $2\omega_a - \omega_b \cong \omega_a$ -> so they are very close to the original signals and hard to filter out. Graphically:



As for harmonic distortion, we can try to compare the power of the 3rd order components relative to the fundamental tones ωa and ωb.

Assuming a "*two-tests tone*" where in $x(t) = B_1 cos(\omega_a t) + B_2 cos(\omega_b t)$ the two ωa and ωb are close to each other and $B_1 = B_2 = B$ is small, one can demonstrate that

- Coefficient for the fundamental tones is $a_1 B$
- Coefficient for the 3rd order signals is $\frac{3a_3B^3}{4}$

Intermodulation

If now we impose the two coefficients to be the same, we can derive an expression for the corresponding input signal level B. If then we calculate this in dB, we call this **third-order intercept point (IIP3 or OIP3)**.

$$\boxed{a_1 B = \frac{3B^3 a_3}{4}} \longrightarrow \boxed{B(IIP3) = \sqrt{\frac{4}{3}\left|\frac{a_1}{a_3}\right|}}$$

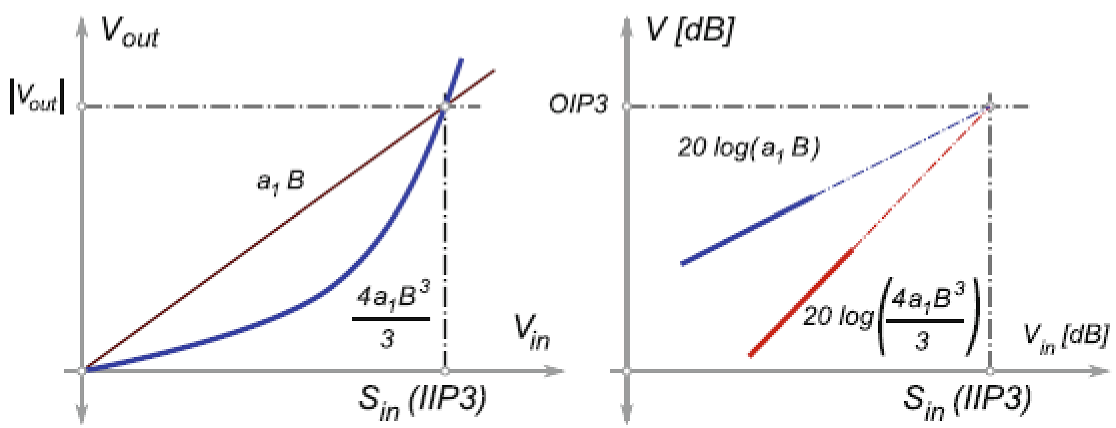
Comparing this with the 1dB compression point formula, one can demonstrate that

$$\boxed{B(-1\,\mathrm{dB}) = \sqrt{0.145\left|\frac{a_1}{a_3}\right|}}$$

$$\boxed{B(-1\,\mathrm{dB}) = \sqrt{\frac{4}{3}\left|\frac{a_1}{a_3}\right|0.11} = IIP3 - 9.6\,\mathrm{dB}}$$

The graphical interpretation of this approach to find IIP3 is given below

Fig. 13.8 Third-order intercept point extrapolation



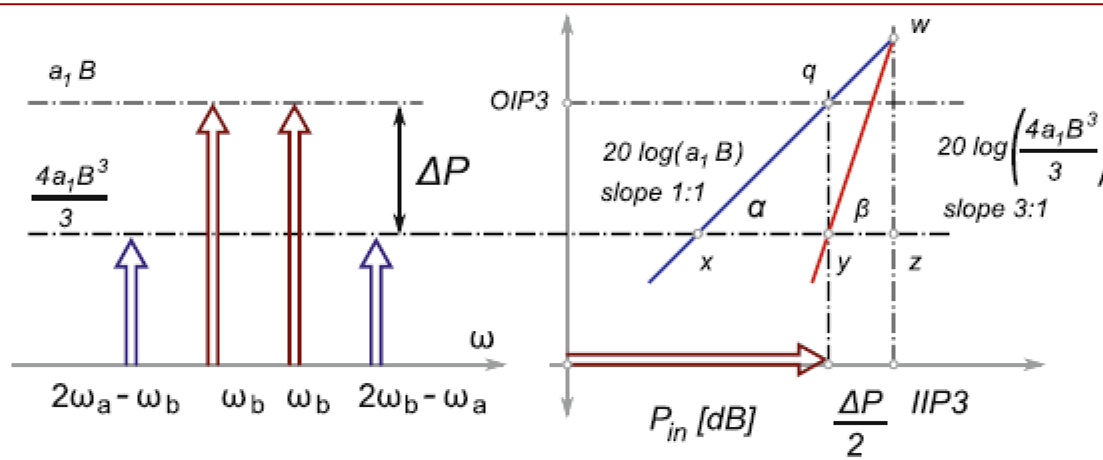
Intermodulation

Note that we obtained the expression for the IIP3 under a small input signals hypothesis (small B). If this does not hold, the analytic expression is no longer valid and you need to <u>measure</u> the relative output power difference ΔP with a spectrum analyser (difference between fundamental and third order tones).

Using geometrical properties in the graph below we can demonstrate that

$$IIP3 = P_{in} + \frac{\Delta P}{2}\ [dB]$$

Fig. 13.9 Graphical solution for third-order intercept point



Note that in all our analysis we have ignored for simplicity the 2nd order tones, but a detailed analysis of intermodulation goes beyond this course.

Image frequency

We discussed the issue of **image/ghost frequencies** for mixers. Let us recap.

If we have an audio signal at f_m=1kHz embedded on a carrier frequency f_C=10MHz, and this is down-converted by a heterodyne receiver with VCO tuned at 9.999MHz, we would get

$$f_1 = f_C + f_{VCO} = 10\,\text{MHz} + 9.999\,\text{MHz} = 19.999\,\text{MHz},$$

$$f_2 = f_C - f_{VCO} = 10\,\text{MHz} - 9.999\,\text{MHz} = 1\,\text{kHz},$$

Here f2 is the desired signal that we want to keep, and f1 needs to be discarded using a low-pass filter.

However, let us suppose that there is another RF signal at frequency f_{ghost}

$$f_{\text{ghost}} = f_C - 2f_m = 10\,\text{MHz} + 2 \times 1\,\text{kHz} = 9.998\,\text{MHz},$$

Because it is close to the previous carrier (10 MHz) it will get through the RF amplifier's tuning circuit and enter the mixer, generating

$$f_3 = f_{VCO} + f_{\text{ghost}} = 9.999\,\text{MHz} + 9.998\,\text{MHz} = 19.997\,\text{MHz}$$

$$f_4 = f_{VCO} - f_{\text{ghost}} = 9.999\,\text{MHz} - 9.998\,\text{MHz} = 1\,\text{kHz}.$$

The result is that you can obtain an undesired IF signal (f4) due to the presence of the ghost/image RF signal. This f4 will overlap and distort the desired f2 signal.

Image frequency

How to deal with ghost/image frequencies? The key point is to avoid that they reach the mixer stage

- Increasing the quality factor of the image-rejection filter (see slides from week 2)
- Imposing by national/international standards a minimum distance between any two neighbouring radio-transmitting frequencies, OR declaring some ranges of frequencies forbidden for transmission
  - You may have heard or GSM, 3G, 4G, 5G, WiFi, Bluetooth – these are all international standards that will dictate in which range of frequencies you can transmit and receive
- Using a super-heterodyne receiver where there are two down-conversion stages (i.e. two pairs of tuned VCO and mixers) to further separate desired and ghost tones

An example of standard

In the UK the national authority for allocation of frequency spectrum is OFCOM. OFCOM decides where each different application or user can transmit. You may play with the interactive map below to see what/who is transmitting at different frequency bands http://static.ofcom.org.uk/static/spectrum/map.html

If in your future job you design a receiver (or transmitter) you need to know the relevant standards…telecommunication engineers also work for the authority itself, or for international standard organisations (ITU, ETSI,…).

Receivers specifications: DR

Apart from dictating where you can or cannot transmit/receive, the telecommunication standards also set some specifications that receivers are compulsory to have.

In slide 6 we mentioned the **dynamic range DR**, the ratio of the largest and smallest values that the system is capable of processing. This is typically measured in dB and is a dimensionless number.

For example if the smallest amplitude is 1mV and the largest is 1V, the DR is $DR = 20\log_{10}(1/_{0.001}) = 60\ dB$

Typical receivers have DR of 100dB or more (as you can guess, the wider the DR the higher the cost).

The ***upper limit of the DR*** is typically set by the non-linearity in the circuits, often measured by the 1dB compression point or the IIP3 (3^{rd} order intercept point). As often this is limited by the available DC power supply, a simplistic way to raise this limit can be to design the circuit to operate with increased power supply level.

RXs specifications: Noise Floor

The ***lower limit of the DR*** is related to the noise, as we need to identify what is the minimum receivable signal level against the background of noise.

The power of the thermal noise Pn in the system can be represented by $\boldsymbol{P_n = kT\Delta f}$ [Watt] where k is Boltzmann constant, Δf is the bandwidth of the system, and T the temperature in the environment (in Kelvin).

If we normalise the Pn per unit of frequency and assume a room temperature of T=290 K, then

$P_n = kT = 1.38 \times 10^{-23} \times 290\ [{}^{W}/_{Hz}]$ → **P_n= -174 dBm = -204 dBWatt** (note the logarithmic scale)

This number is typically referred to as the "**noise floor**" of the environment. One can still reduce this number by cooling the environment (reducing T) or minimising the bandwidth used in the system (reducing the parameter Δf, if this is possible).

RXs specifications: sensitivity

In a real RF receivers, the signal will occupy a certain bandwidth Δf so the actual noise floor is $P_n = -174dBm + 10\log_{10}(\Delta f)\ \ [dBm]$

An RF signal arriving at the initial point of your receiver (an antenna if we think of a wireless comms system) will travel through the whole receiver. The components of the receiver will add some noise, their internal thermal noise, and this is summarised by the **Noise Figure NF** parameter of the receiver. So the true noise floor is

$$\boldsymbol{P_n = -174dBm + 10\log_{10}(\Delta f) + NF}\ [\boldsymbol{dBm}]$$

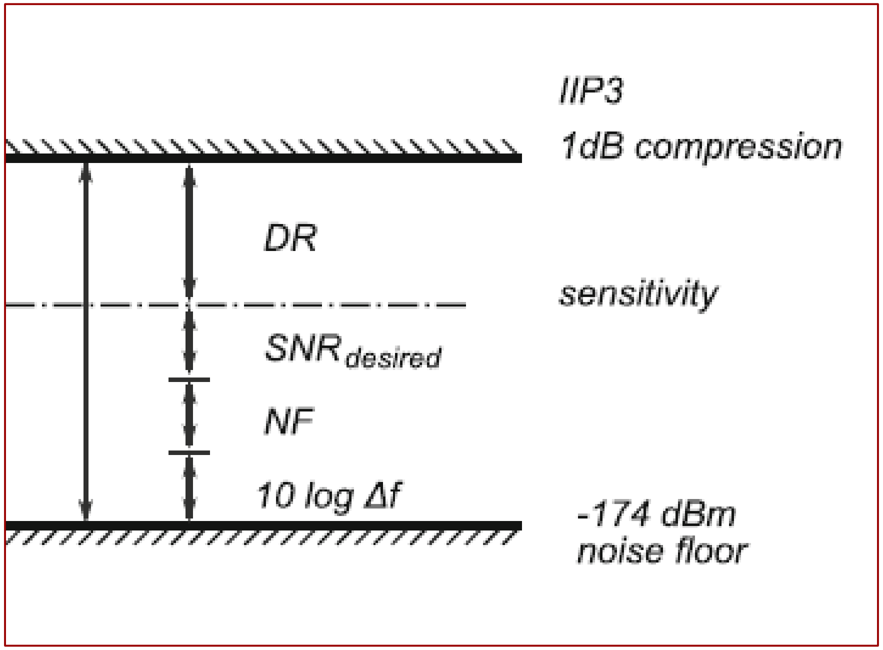
We previously defined the **sensitivity S** of the receiver as the minimum received signal level that the receiver can process with a certain Signal to Noise Ratio (SNR), so that $\boldsymbol{S_n = P_n + SNR_{desired}}$

The ideal dynamic range DR is defined as the difference between the 1dB compression point and the sensitivity $\boldsymbol{DR = 1dB_{point} - S_n}\ [\boldsymbol{dB}]$

This is somewhat an optimistic result, often adjusted for about 30%, so that the realistic DR is about 2/3 of the ideal number.

RXs specs: sensitivity & DR

The graph below summarises the key metrics seen in the previous slides for RF receivers.



Straightforward example. *Determine the sensitivity of a receiver at room temperature whose noise figure is 5dB and desired SNR 10dB if the useful bandwidth is 1 MHz.*

Result S= -99dBm

More on noise

We mentioned **NF noise figure** of a system (receiver for example) as a measure of *how much noise the system adds* to the input and this noise shows itself at the output. The NF is the logarithm of the noise factor F. The formula below can be written where A is the power gain of a generic communication component

$$NF = 10log_{10}(F) = 10log_{10}\left(\frac{SNR_{in}}{SNR_{out}}\right) = 10log_{10}\left(\frac{Psig_{in}}{Psig_{out}}\frac{Pn_{out}}{Pn_{in}}\right) = 10log_{10}\left(\frac{Pn_{out}}{APn_{in}}\right)$$

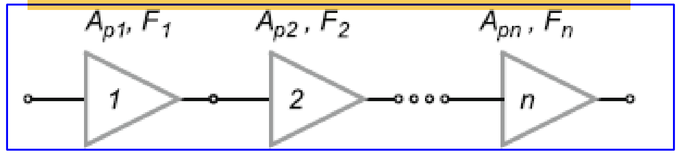
Ideally NF should be 0, but every electronic component adds noise, including amplifiers, so NF>0.

It can be demonstrated that the noise contribution from an electronic component/bloc is $\boldsymbol{P_{na} = (F-1)kT\Delta f}$ where k is Boltzmann's constant, T is the temperature in Kelvin degrees, and Δf the bandwidth of the signal

Let us consider an amplifier. Any noise signal P_{ni} at its input will be amplified at the output by gain A to generate output P_{no}, but the amplifier will also add its internal noise P_{na}. So the total output noise power P_1 for 1 stage amplifier is $P_1 = P_{no} + P_{na} = AP_{ni} + (F-1)kT\Delta f$

More on noise – Multiple stages

If we have a multiple stage amplifier, its total (all stages) output noise power can be written as $\boldsymbol{P_{no,tot} = F_{tot}A_{tot}kT\Delta f}$ – How do we evaluate the total noise factor F_{tot}?



Let us consider just 2 stages. The noise output after the 1st one is

$$P_{no,1} = F_1 A_1 kT\Delta f$$

This is the input to the second stage so this is amplified, but the second stage will also add (and amplify) its internal noise hence

$$P_2 = A_2 P_{no,1} + P_{int,2} = A_2 F_1 A_1 kT\Delta f + A_2(F_2 - 1)kT\Delta f$$

This can be rewritten as $P_2 = A_2 A_1 kT\Delta f\left(F_1 + \frac{F_2 - 1}{A_1}\right)$

Comparing with the initial formula we can write that $A_{tot} = A_2 A_1$ and $F_{tot} = F_1 + \frac{F_2 - 1}{A_1}$

This can be generalised to **Friis formula for the NF of cascaded network** made of multiple stages one after the other (note F is not in dB!)

$$F_{(tot)} = F_1 + \frac{F_2 - 1}{A_{P1}} + \frac{F_3 - 1}{A_{P1}A_{P2}} + \cdots + \frac{F_n - 1}{A_{P1}A_{P2}\cdots A_{P(n-1)}}.$$

More on noise – Filter circuits

We have said earlier (slide 19) that the thermal noise power can be estimated by $\boldsymbol{P_n = kT\Delta f}$ [Watt] where k is Boltzmann constant, Δf is the bandwidth of the system, and T the temperature in the environment (in Kelvin).

However, circuits which acts as filters will restrict the total available bandwidth, essentially filtering/reducing also the noise. In this case we need to modify the relation to $\boldsymbol{P_n = kT\Delta f_{eff}}$ where $\boldsymbol{\Delta f_{eff}}$ is an "*effective bandwidth*". Here we note two important cases (*theory is in chapter 3 of Sobot's book if you want more details*).

-For RC LPF circuits $P_n = kT\Delta f_{eff}$ where $\boldsymbol{\Delta f_{eff} = \frac{1}{4RC}}$ essentially the bandwidth depends on the values of the components R and C

-For RLC BPF circuits $P_n = kT\Delta f_{eff}$ where $\boldsymbol{\Delta f_{eff} = \frac{\pi}{2} B_{3dB}}$ where the effective bandwidth is related to the 3dB bandwidth of the band-pass filter through a π/2 coeffiecient.

Summary

So far we have discussed the following topics about RF receivers:

- Briefly the role of **channel** in a communication system (in very simple terms, just an attenuation)
- Two main **topologies** of receiver: TRF vs (super)heterodyne receivers, with motivation of why using the latter
- **Non-linear effects** in receivers: harmonic distortion, gain compression and 1dB point, intermodulation and IIP3, image/ghost frequency
- **Receiver specifications** (metrics) to meet for communication standards: dynamic range, noise floor, sensitivity
- Some additional information on **noise**, in particular the noise of a multiple stages system and the noise of a RLC band-pass circuit.

Transmitters

Most **transmitters** architectures are similar to the receivers' topologies but with operations performed in the reversed order.
For example, if at the RX we have antenna->LNA->downconversion mixer->amplifier, at the TX we can have data modulator->upconversion mixer->amplifier->antenna

Although outside the scope of this course, you may want to recall that typically the data you want to transmit are modulated/encoded in a certain format requiring **quadrature baseband signals**.
Quadrature -> complex data, with I/Q components (phase & quadrature)
Baseband -> their spectrum is at low frequency, close to DC 0 Hz
For example a GMSK (Gaussian Minimum Shift Keying) RF signal used for GSM mobile phone or Bluetooth looks like

$$x_{GMSK}(t) = Acos(\omega_c t)cos\Phi - A(sin\omega_c)tsin\Phi$$

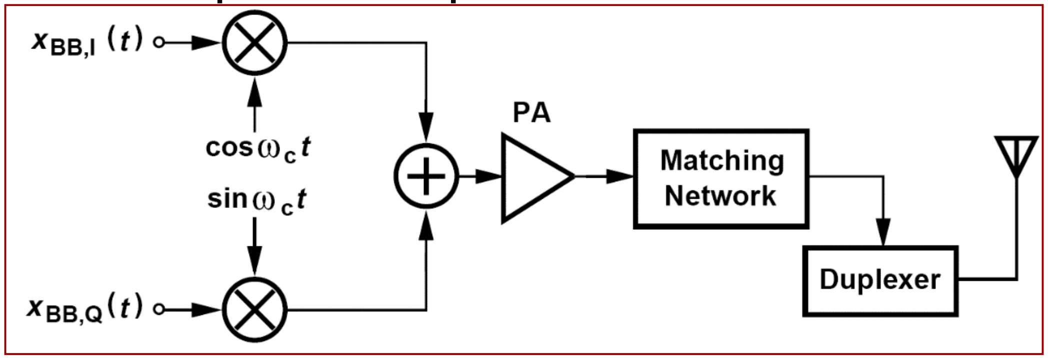
Where ω_c is the RF carrier of the signal, and cosΦ and sinΦ are the I/Q baseband components produced by the baseband data modulator.

Transmitters: direct conversion

So a transmitter's job is typically to generate the baseband components in its digital part, convert them into analogue waveforms, and up-convert them into the required format and RF frequency band required for a specific application.

The easiest architecture is shown below. This is a **direct conversion transmitter** where the original I/Q baseband components are up-converted to the carrier frequency ωc by mixing, combined, amplified, and matched to the antenna for transmission. Note that the duplexer is a device that allows simultaneous transmission and reception with the same antenna (ignore that for now). The conversion is "direct" because the signal "jumps" straight from Baseband to RF frequency.

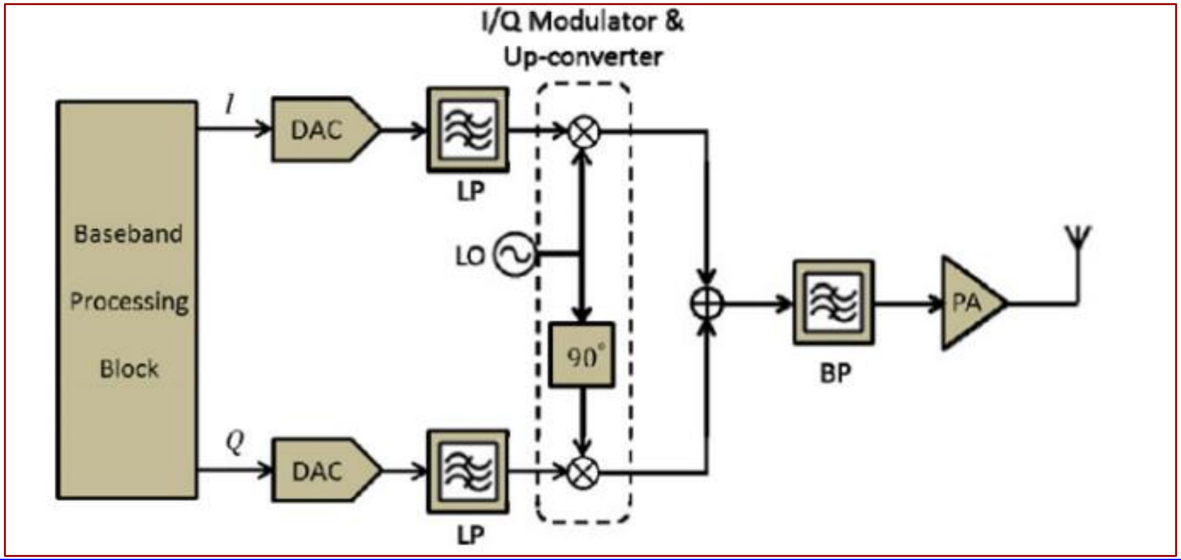
<u>Question: can you anticipate the spectrum at baseband and at RF?</u>



B. Razavi, "RF Microelectronics", Prentice Hall, 2nd ed. – Chapter 4

Transmitters: direct conversion

A slightly different implementation is shown below.
DAC-> digital to analogue converters; **LP**-> low-pass filters
LO-> local oscillator (note the 90 degrees shift to generate cosine/sine as input to the two mixers)
BP->band-pass filter; **PA**->power amplifier connected to the antenna



The DACs convert the digital in-phase (I) and quadrature (Q) baseband signals to analog signals, which feed the quadrature modulator for direct up-conversion. Analog signals are directly up-converted to the RF with I and Q carriers. The block diagram of a direct conversion transmitter is shown in Figure 6.2. The bandpass filter after the signal summation is used to suppress the out-of-band signals produced by the harmonic distortion of the carrier. The RF signal power is increased by the PA for transmission through the antenna.

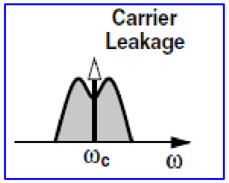
A. Mohammadi, F. M. Ghannouchi, *"RF Transceiver Design for MIMO Wireless Communications", Springer – Chapter 6 – Online at Glasgow University Library*

TXs: direct to superheterodyne

Direct conversion transmitters have some advantages:

- Compact and simple solution (no IF signals components)
- “Clean” spectral output (only the desired signal around the carrier frequency and its harmonics, but no spurious intermodulation components)

However they also suffer a few problems:



- LO leakage into RF (the RF output has a fraction of the unmodulated carrier superimposed to the desired signal)
- I/Q mismatch (unequal gains on the I and Q branches, leading to distortion of the resulting RF signal)
- VCO in the LO pulling (as the PA works at the same frequency of the LO, the coupling effect can be such to “pull” the frequency of the LO, modulating & changing this resulting in increased phase noise).
- Need of a mixer capable to work from baseband to RF

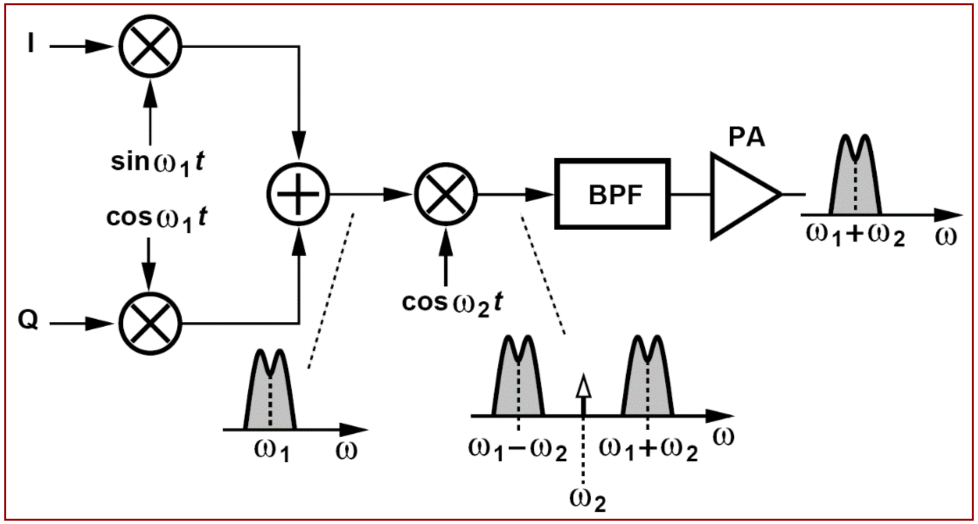
As a result, superheterodyne transmitters can be preferable

Transmitters: superheterodyne

The main point of **superheterodyne transmitters** is that the frequency up-conversion is performed in two stages, first from baseband to a carrier ω_1, then up to ω_2. The final useful signal is at $\omega_1 \pm \omega_2$, far from the frequency of any LO (to avoid LO leakage and LO pulling).

As the initial I/Q up-conversion happens at much lower frequency than for direct conversion, the problem of I/Q mismatch/unbalance is less significant.

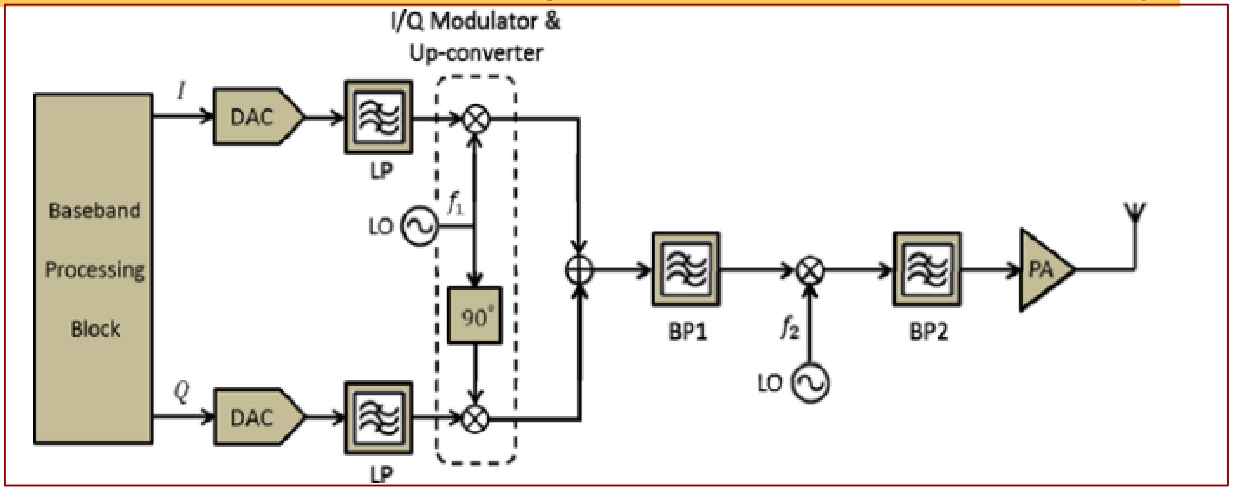
An example of block diagram shown below.

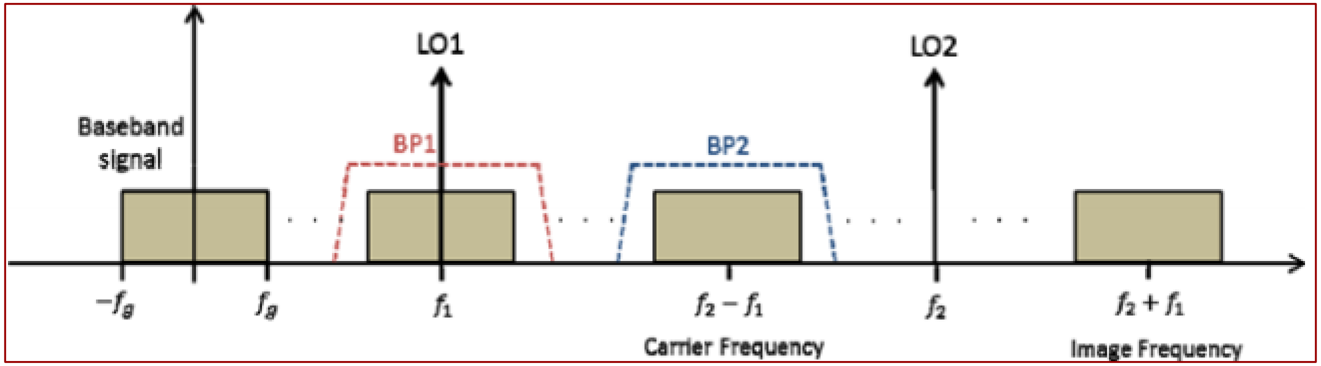


B. Razavi, "RF Microelectronics", Prentice Hall, 2nd ed. – Chapter 4

Transmitters: superheterodyne

An alternative implementation below. The BP1 removes the spurious out-of-band contributions from the first mixing stage, whereas the second BP2 removes the unwanted frequency component within f1±f2. Once again, the main rationale of this topology is the increased frequency separation between desired & undesired components for better filitering.

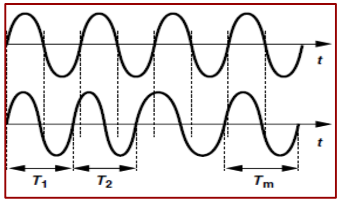


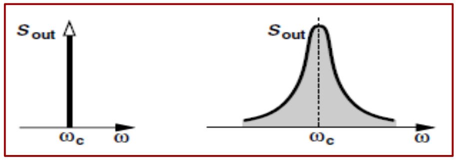


A. Mohammadi, F. M. Ghannouchi, *"RF Transceiver Design for MIMO Wireless Communications", Springer – Chapter 6 – Online at Glasgow University Library*

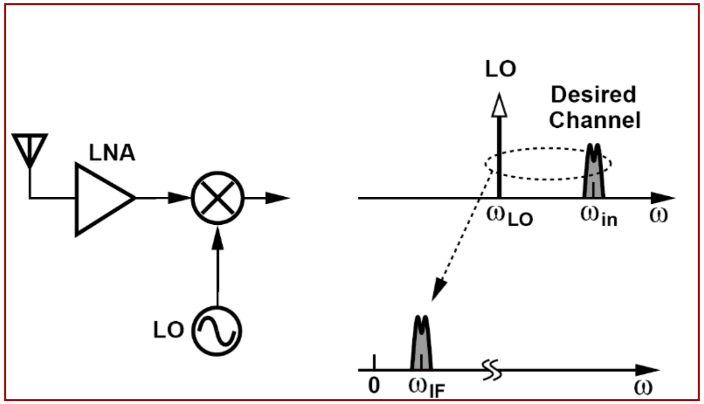
More on Phase Noise at RX

We mentioned phase noise as a characteristics of non-ideal oscillators, whereby their generated frequency is not stable over time. This produces "jitter" in the time domain (observable with an eye diagram) and spectral broadening in the frequency domain.





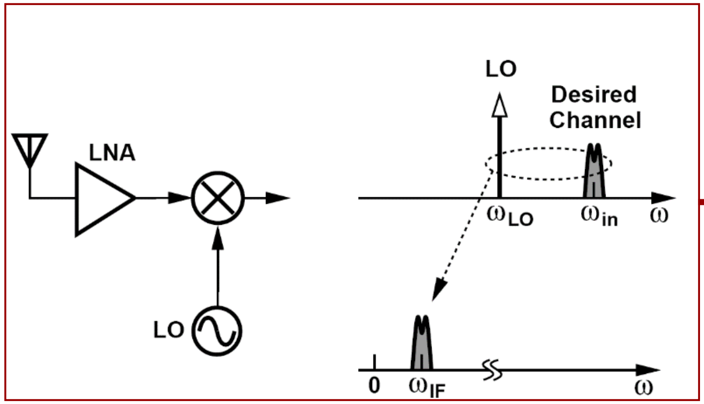
We now want to observe phase noise effect in the RF domain. We assume to have a receiver with frequency down-conversion (left). Ideally the desired signal centred at an RF frequency ω_{in} is down-converted to the IF frequency $\omega_{IF} = \omega_{in} - \omega_{LO}$ (right)

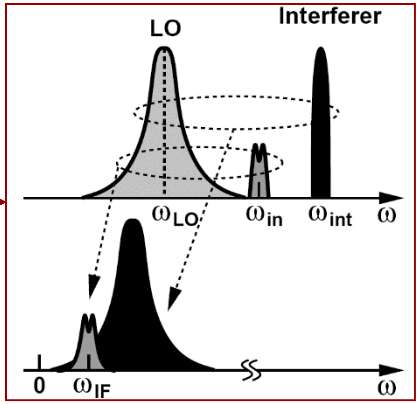


More on Phase Noise at RX

However, the LO will have some phase noise and it is likely that in the RF channel there are interferers, i.e. undesired signals (which could well be other signals from competing users, other TV channels, other mobile phone users,…).
The resulting effect is shown on the right-hand side, where the convolution of desired signal + interferer with the noisy LO spectrum results in a broadened down-converted interferer at IF, whose “phase noise skirt” can corrupt the desired signal. This phenomenon is called **reciprocal mixing**.

To reduce this effect, LO phase noise must be small enough to produce negligible corruption on the desired signal in the presence of interference.



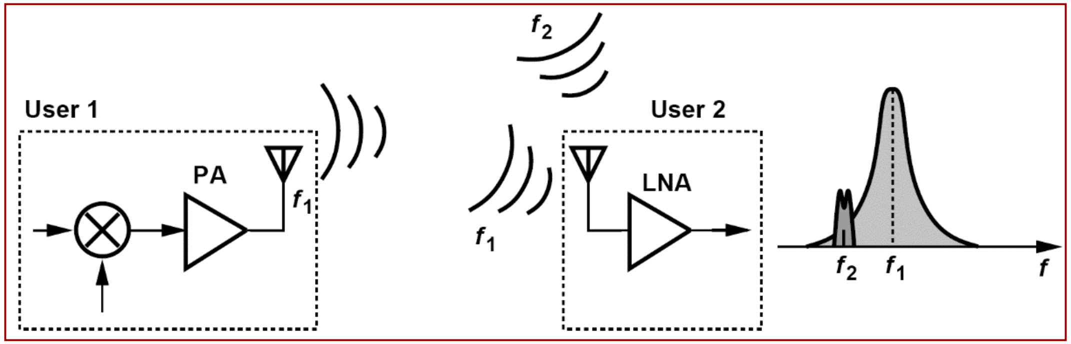


B. Razavi, “RF Microelectronics”, Prentice Hall, 2nd ed. – Chapter 8.7

More on Phase Noise at TX

Phase noise also has <u>effects at the transmitters</u>. For example, if two users are located close to each other, one might transmit a strong signal at a frequency f_1 and the other receiving a desired weak signal at frequency f_2 but also the strong signal at f_1.

If f_1 and f_2 are close to each other, then the phase noise skirt of f_1 will mask/corrupt the desired signal at f_2 for the second user's receiver, and this happens even before the down-conversion stage.



B. Razavi, "RF Microelectronics", Prentice Hall, 2nd ed. – Chapter 8.7

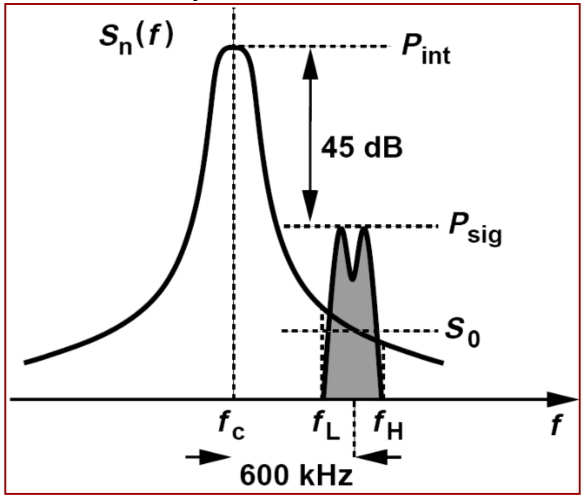
These few slides showed a couple of practical simple cases of when phase noise is relevant. We can now look at an example.

More on Phase Noise: example

A GSM receiver (each channel is 200 kHz wide between f_H and f_L) must withstand an interferer located 3 channels away from the desired one, and 45dB higher. Estimate the maximum tolerable phase noise of the LO if the corruption due to reciprocal mixing must remain 15 dB below the desired signal.

The total noise power introduced by the interferer in the desired channel is equal to

$$P_{n,tot} = \int_{f_L}^{f_H} S_n(f)df$$



For simplicity, we assume $S_n(f)$ is relatively flat in this bandwidth and equal to S_0, so the total phase noise power contribution is a rectangle $S_0(f_H - f_L)$. The ratio of the useful signal power P_{sig} and this noise power has to be at least 15dB, so

$$10\log_{10}(\frac{P_{sig}}{S_0(f_H - f_L)}) = 15\ dB$$

Since the interferer is convolved with the LO phase noise S_0, it must be normalized to the power of the interference P_{int} which we know is 45dB stronger than the useful signal P_{sig}

This means $10\log_{10}(\frac{P_{int}}{P_{sig}}) = 45dB$; combining the two $10\,log_{10}\left(\frac{P_{sig}}{S_0(f_H-f_L)}\right) + 10\,log_{10}(\frac{P_{int}}{P_{sig}}) = 15dB + 45dB$

If we multiply both terms by -1 and use logarithmic properties, we can write

$$10\,log_{10}\left(\frac{S_0}{P_{int}}\right) + 10\,log_{10}(f_H - f_L) = -\,15dB - 45dB$$

As the channel is 200 kHz, then $\mathbf{10\,log_{10}\left(\frac{S_0}{P_{int}}\right) = -113dBc/Hz}$

<u>Note this compares the phase noise S_0 from the LO with the interference level P_{int}</u>

RF Transceivers – On chip

We have discussed receiver and transmitters. You may come across this term **transceivers**, which typically indicates a unique system/device capable of both transmitting and receiving.

One aspect to be well aware of is that, although we tend to study transmitters/receivers/transceivers as a block diagram, they can be integrated into a single, compact device. Sometimes people call them ***transceivers on chip***. We want to quickly look at some examples.

AD9364 Analog Devices

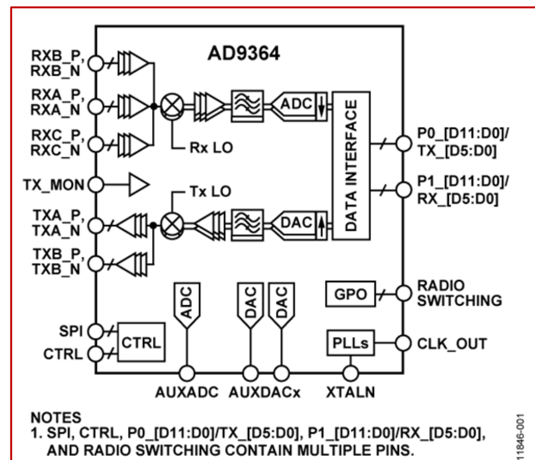
https://www.analog.com/media/en/technical-documentation/data-sheets/AD9364.pdf

The AD9364 is a 1 x 1 channel high performance, highly integrated RF Agile Transceiver™. Its programmability and wideband capability make it ideal for a broad range of transceiver applications. The device combines an RF front end with a flexible mixed-signal baseband section and integrated frequency synthesizers, simplifying design-in by providing a configurable digital interface to a processor. The AD9364 operates in the 70 MHz to 6.0 GHz range, covering most licensed and unlicensed bands. Channel bandwidths from less than 200 kHz to 56 MHz are supported.

The AD9364 is packaged in a 10 mm × 10 mm, 144-ball chip scale package ball grid array (CSP_BGA).

Applications

- Point to point communication systems
- Femtocell/picocell/microcell base stations
- General-purpose radio systems



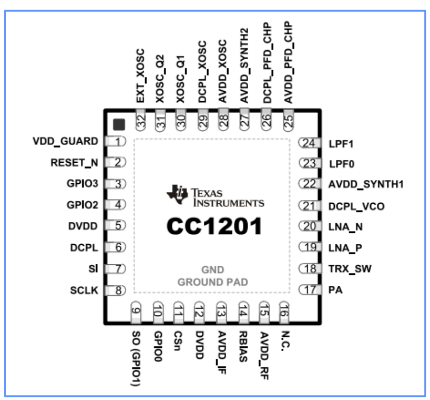
RF Transceivers – On chip

TI CC1201 Texas Instrument

http://www.ti.com/lit/ds/symlink/cc1201.pdf

The CC1201 device is a fully integrated single-chip radio transceiver designed for high performance at very low-power and low-voltage operation in cost-effective wireless systems. All filters are integrated, thus removing the need for costly external SAW and IF filters. The device is mainly intended for the ISM (Industrial, Scientific, and Medical) and SRD (Short Range Device) frequency bands at 164–190 MHz, 410–475 MHz, and 820–950 MHz.

The CC1201 device provides extensive hardware support for packet handling, data buffering, burst transmissions, clear channel assessment, link quality indication, and Wake-On-Radio. The main operating parameters of the CC1201 device can be controlled through an SPI interface. In a typical system, the CC1201 device will be used with a microcontroller and only few external passive components.



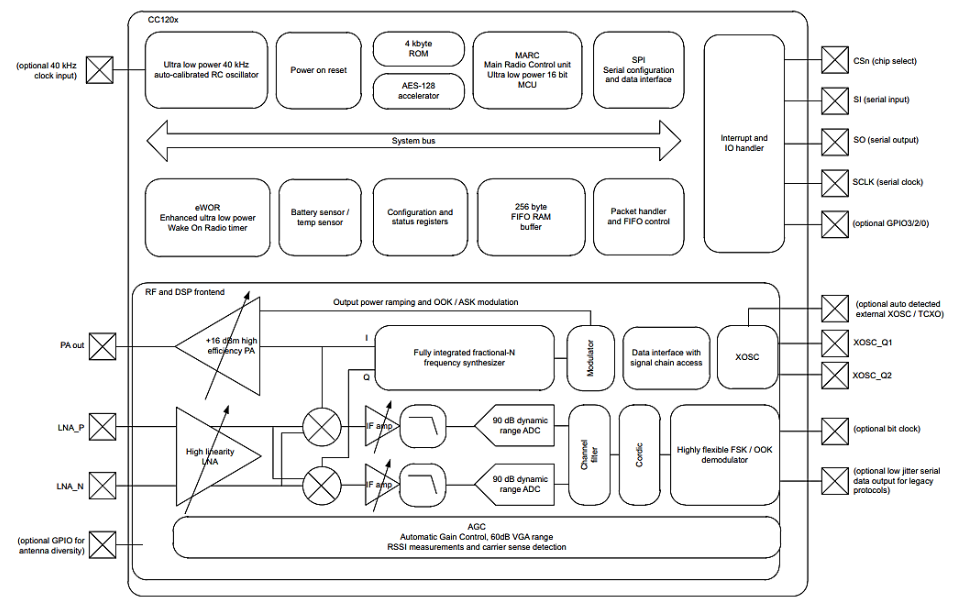
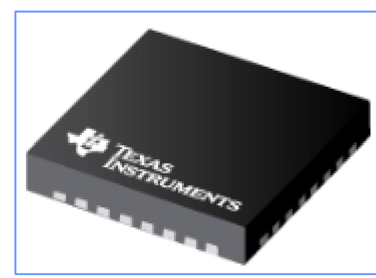


Figure 1-1. System Block Diagram

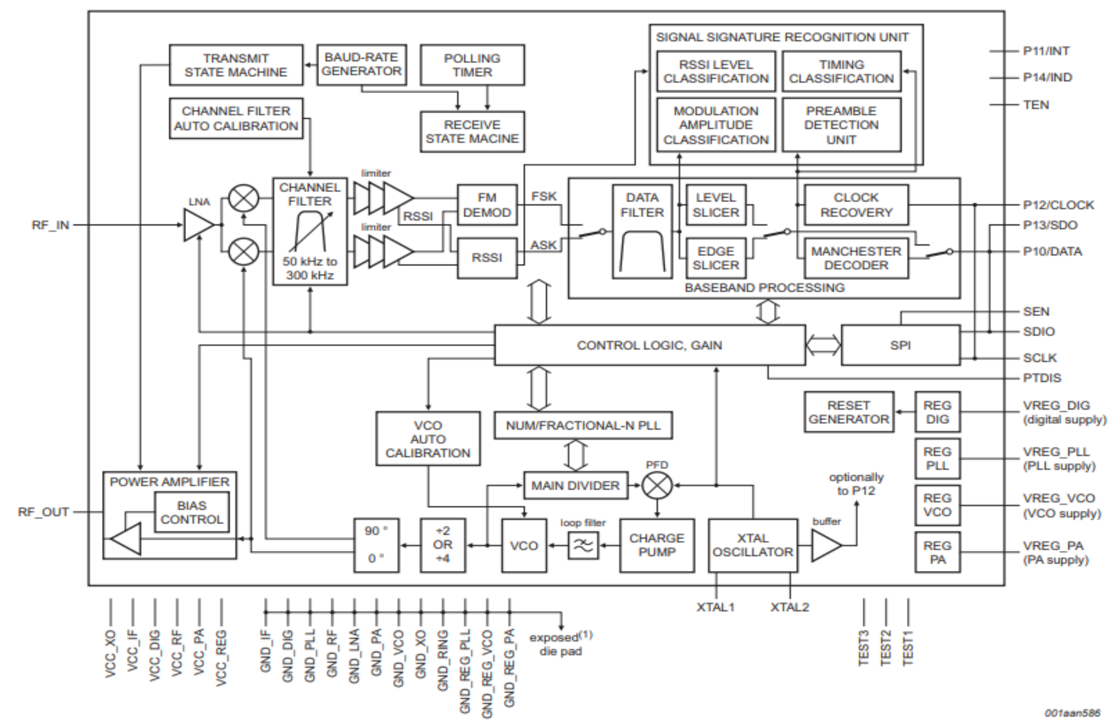


RF Transceivers – On chip

OL2381 NXP

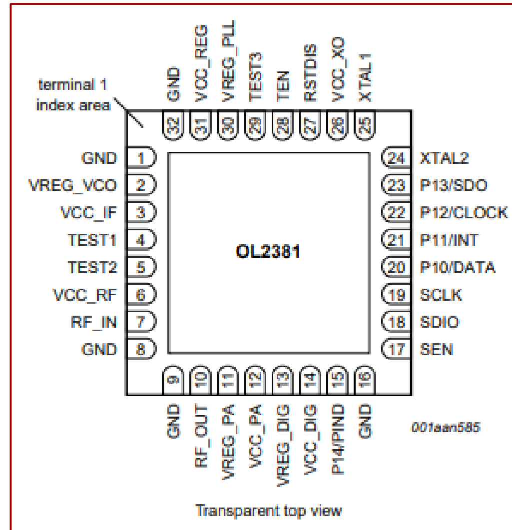
https://www.nxp.com/docs/en/data-sheet/OL2381.pdf

Check in the datasheet the TX and RX block diagrams (figures 3 and 4)



(1) All internal domain grounds including external GND pins 1, 8, 9, 16 and 32 are connected to the exposed die pad.

Fig 1. Block diagram



A highly integrated single-chip transceiver solution, the OL2381 is ideally suited to telemetry applications operating in the ISM/SRD bands. The small form factor, low power consumption and wide supply voltage range make this device suitable for use in battery powered, handheld devices and their counter parts.

RF Transceivers – On chip

It is also interesting to look at this brochures from Analog Devices about SDR (Software Defined Radio) solutions.
https://www.analog.com/media/en/news-marketing-collateral/solutions-bulletins-brochures/Software-Defined-Radio-Solutions-From-ADI.pdf

Page 2 – You can see a slightly different version of the chip on slide 33 but with 2 pairs of TX and RX channels

Page 4 – Pay attention to the list of different chips that can be used to implement each of the different block in a transceiver (listed ADC and DAC, PLLs, RF amplifiers, LNAs, clock generators).

The key message is you may have all your transceiver into a single chip (flexible and compact, but performances/requirements might be limited) or you can implement the transceiver with different chips (more complicated PCBs, more components, but can meet more specific requirements)

High Speed Data Converters
- AD9680—dual, 14-bit, 1.0 GSPS ADC with JESD204B serial output interface
- AD9144—quad, 16-bit, 2.0 GSPS TxDAC+® transmit DAC with JESD204B serial output interface
- AD9625—dual, 14-bit, 2.5 GSPS ADC with JESD204B serial output interface

PLLs
- ADF4351, ADF4355-2—wideband synthesizer with integrated VCOs

RF Amplifiers
- ADL5601, ADL5602—50 MHz to 4 GHz broadband 20 dB linear amplifiers
- ADL5320—400 MHz to 2700 MHz, ¼ W RF driver amplifier
- ADL5604—700 MHz to 2700 MHz, 1 W RF driver amplifier
- ADL5610, ADL5611—30 MHz to 6 GHz RF amplifiers for high performance applications
- ADL5544, ADL5545—30 MHz to 6 GHz RF amplifiers for low power applications

Low Noise Amplifiers
- ADL5523, ADL5521—400 MHz to 4000 MHz low noise amplifiers

ADC Driver Amplifiers
- AD8366—dual variable gain amplifier
- ADA4961—low distortion, dc to 2.5 GHz ADC driver amplifier

Power Detector
- ADL5501—50 MHz to 6 GHz TruPwr™ detector

Clocking
- AD9523—low jitter clock generator with 14 outputs

Integrated ISM Band Transceivers
- ADF7023—transceiver operating in the license free ISM bands

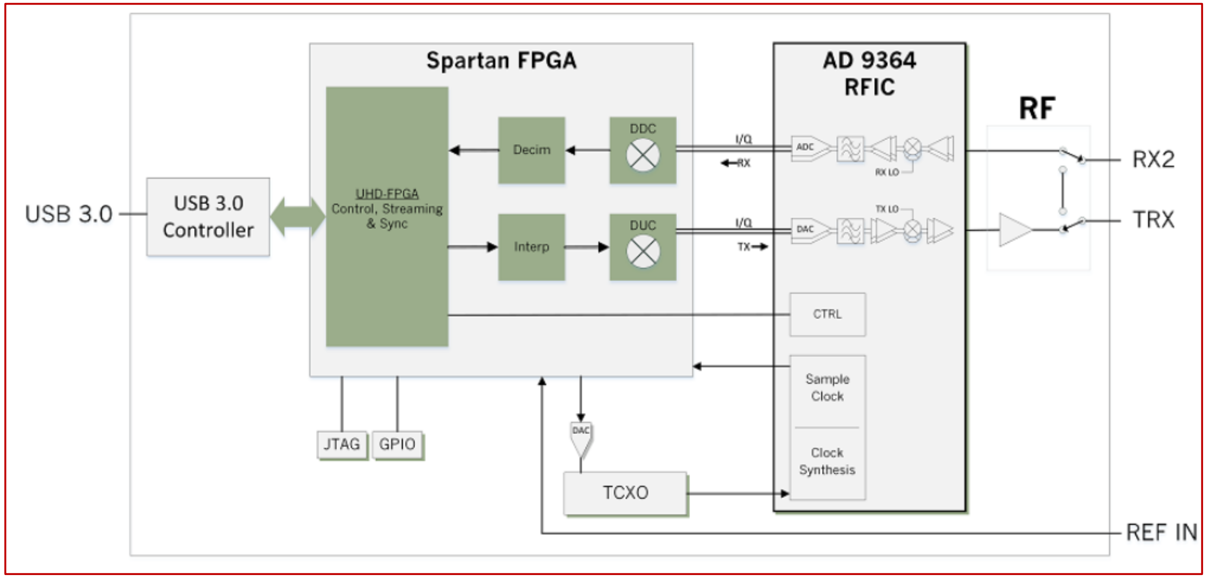
Power Management
- ADP5040—1.2 A buck regulator and dual 200 mA LDOs

RF transceivers - Systems

Look at this device, which is part of the SDR systems, essentially a reconfigurable device capable of transmitting/receiving RF signals and comms-> https://www.ettus.com/product/details/USRP-B200mini-i

First of all it is very small, 5*8*8cm for 24g weight!

Second, look at the block diagram inside



DDC -> digital down-converter

DUC -> digital up-converter

ADC -> analog-to-digital conv

DAC -> digital-to-analog conv

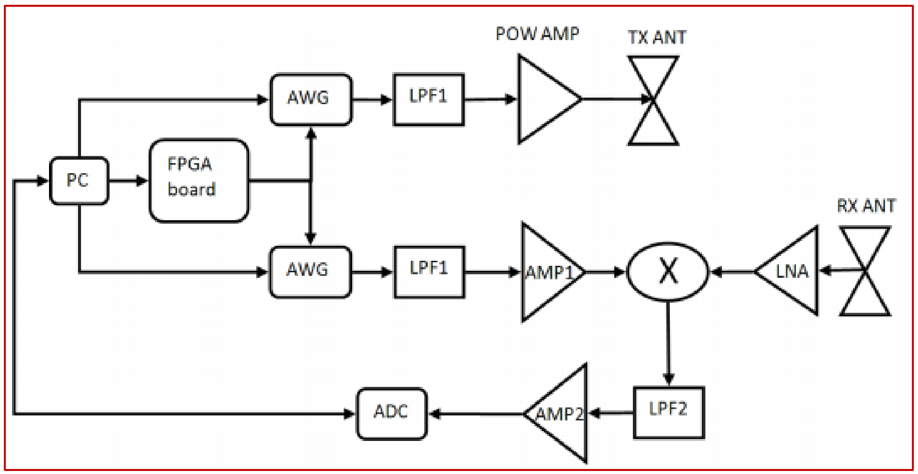
There is essentially one FPGA chip for the digital signal processing, and an analogue chip AD9364 for the transceiver part, plus a complementary power amplifier for the transmitter and a crystal oscillator (TCXO) generating the clock.

Can you recognise all the blocks within the AD9364 on the top part???

What about a radar system?

It's another story, but the structure of the transceiver can be very similar to a communication system. Take for example this one from a PhD thesis http://etheses.dur.ac.uk/9432/1/Through-The-Wall_Detection_Using_Ultra_Wide_Band_FMICW_Signals.pdf?DDD10+

In Fig. 5.1 and related chapter you can see the description of the various components, some of which you know (highlighted in bold).



AWG	Arbitrary waveform generator
LPF1-2	Low-pass-filter (cut-off at 3.5 GHz for 1, and at 10 MHz for 2)
PowAmp	Power amplifier ZHL-4240
LNA	Low Noise Amplifier ZX60-3011
X	Mixer ZEM-4300MH
AMP1	Medium-power amplifier ZKL-2R7
AMP2	Baseband amplifier ZHL-6A
ADC	Analog to digital converter
FPGA	Field Programmable Gate Array board

Note that the components highlighted above were not single chips, but connectorised components from MiniCircuits (we have looked at them when presenting individual components, mixers, amplifiers, and so on in week 2).

Summary

In the second group of slides we have discussed the following topics:

- **Transmitter** architectures: quadrature transmitters, 2 main topologies: direct conversion and superheterodyne.
- More on the effects of **phase noise**, in particular the reciprocal mixing effect at the receiver with a simple design exercise
- Examples of realistic implementations of RF transceivers, either on chip or using connectorised components

In the next set of slides we will see some design exercises, i.e. how to combine together different “blocks” to implement a transceiver (or part of it) to meet certain given requirements