An Introduction to

Radio Architectures and Signal Processing

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**Short Overview**

Radio transmitters and receivers are essentially interfaces between digital data (bits) and electromagnetic waves.

This can be further divided to front-end processing and baseband processing.

So in addition to antennas, the front-end includes:
- Amplification stages
- Frequency translation stages
- Filtering stages
- Sampling and A/D/A interface

In practice, the front-end usually includes both analog and digital signal processing stages (not only analog).

Different radio architectures then mean how the above functionalities are organized in the radio chain.

**Principal Transmitter Functionalities**

- **Digital BB/IF**
  - baseband waveform generation (coding, constellation mapping, pulse-shaping, etc.)
  - sample rate conversions
  - possibly frequency translation to an intermediate frequency (IF)

- **Analog BB/IF/RF**
  - further bandlimitation (if needed)
  - convert the signal to the final RF range, possibly through an IF stage
  - power amplification and final band-limitation
Principal Receiver Functionalities

Analog RF/IF/BB
- translates the RF signal to lower frequencies
- also implements some band-limitation and amplification
- typically part of selectivity

Digital IF/BB
- final selectivity and frequency translation to BB (if not done already in analog)
- sample rate conversions
- channel equalization, detection and decoding

Also extracting synchronization information (carrier & timing) is one central element on the receiver side.

Focus in this presentation

Here we focus on the basics and fundamentals of frequency translations, filtering and sampling in radios.

Also different radio architectures, like superheterodyne, direct-conversion and low-IF, are reviewed, and how the above functionalities are deployed in them.

Anothe central theme is related to different nonidealities of the most essential circuit modules (amplifiers, mixers, oscillators, samplers, ADC's), and how they affect the radio.

In general, we focus mostly on the receiver side, since the challenges are somewhat bigger compared to transmitters
- detecting the weak desired signal(s) in the presence of much stronger (easily tens of dB's) neighboring signals

Notice: Interactive demonstrations and additional (supporting) material available at
Essential System Characteristics

The essential system characteristics effecting on the front-end design are:

- Signal bandwidth
  - 25 kHz in DAMPS
  - 200 kHz in GSM
  - 1.25 MHz in IS-95
  - ~5 MHz in UMTS/WCDMA
  - 8 MHz in DVB-T
  - Up to 20 MHz in 3GPP-LTE and WiMAX
  - 25 MHz in 802.11a/g
  - In the order of 100 MHz in emerging 4G systems

- Signal-to-Interference-Plus-Noise-ratio needed to detect the signal properly.

- Dynamic range, depending on minimum desired signal level (receiver sensitivity) and the maximum level of other signals to be tolerated in nearby frequencies in the radio.

Relatively small distortion effects in a strong non-desired signal component may cause huge relative distortion for a weak desired signal!!

And due to technological limitations, the weak desired signal can be separated form the adjacent spectral components only at a rather late stage in the processing chain.

Main Targets in Radio Parts in Wireless Communication Systems

1. Terminals
   - small-size
   - low-cost
   - low power consumption
   - multiband & multimode capabilities

2. Base-stations
   - system performance may not be compromised.
   - size and cost also important
   - several parallel RX/TX systems typically in operation simultaneously (multiradio BS).

On both sides, multi-antenna techniques coming to or already in practical use
   - The need for multiple RF chains increases the importance of cost and size aspects.
   - RF impairment effects in multi-antenna systems need to be carefully examined.
Key Areas of Development

- Analog and digital VLSI technologies
- MEMS (micro electro-mechanical systems) technologies
- Packaging
- Analog RF-ASIC design (innovative circuit topologies, etc.)
- High-speed & low-power DSP architectures
- DSP algorithms
- Energy efficiency at all levels
- Analog/DSP-tradeoffs: role of DSP increasing
  - Relatively low progress of ADC technologies as the main bottleneck for increased use of DSP
  - Place for innovative RX/TX architectures
  - Discrete-time analog processing is also an interesting possibility (e.g. DRP by TI)
- Dirty RF: There are possibilities to relax the analog component requirements by compensating analog distortion effects through advanced DSP techniques

Possible Solutions Regarding Multimode Capabilities

Supporting multiple radio access technologies in a single device (like mobile phone or BS):

- Separate highly integrated radios for different systems
  or
- Configurable HW platform, possibly with separate analog front-ends for different frequency bands

It helps if most of the radio functionality is defined in DSP software or easily reconfigurable DSP hardware

=> **Software Radio**: a single hardware solution adaptable to different system standards by changing software and/or reconfiguring digital HW

also called

- Software configurable radio
- Software defined radio
- Flexible radio

Practical solutions are compromises of the above, at least so far.
Cognitive Radio Concepts

With a spectrum analyzer, it is easy to observe that the licensed bands of the radio spectrum are not in a very efficient use. For example, there are gaps in the spectrum which are seldom used, some of the services are not operating all the time, etc.

The terms
  - cognitive radio
  - flexible spectrum use
  - spectrum pooling

refer to ideas of taking most parts of the radio spectrum into more efficient use by relaxing the strict allocation of different services to different frequency bands and allowing opportunistic secondary use of the empty parts of the spectrum, using flexible communications waveforms that would be most suitable for a particular environment.

For implementing the cognitive radio ideas, effective software radio type of transceiver implementation is vital.

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What is needed in the receiver front-end?

- Amplification to compensate for transmission losses
- Selectivity to separate the desired signal from others
- Tunability to select the desired signal
- Conversion to digital domain

In the following, we examine different receiver architectures and non-idealities affecting in the different building blocks used in the receivers.

Main Components for Receiver Front-Ends

- **Amplifiers**
  - Low-noise amplifiers (LNAs) in the first stages.
  - Automatic gain control (AGC) needed to cope with different signal levels.

- **Filters**
  - Impossible to achieve sufficient selectivity by tunable RF filters (operating in the RF frequency band of the modulated signal) to separate the desired signal from others.
  - Sufficient selectivity can be achieved by fixed (RF or IF) filters based on special technologies (SAW, Surface Acoustic Wave, ceramic, crystal, mechanical) or analog filters operating on baseband or relatively low bandpass center frequencies
Main Components for Receiver Front-Ends

(...cont’d...)

or multirate digital filters up to some hundreds of MHz range.

- Special complex filters, phase splitters (related also to Hilbert transformers) can be used to suppress certain frequency range from the negative part of the frequency axis. Such filters find application in certain special receiver architectures.

Main Components for Receiver Front-Ends

• Mixers

- Complex (I/Q, quadrature) mixer: pure frequency translation by the local oscillator frequency:

- Real mixer produces the combination of frequency translations in both directions:

More details in the separate material! (earlier complex signals and systems material)
Main Components for Receiver Front-Ends

- **Oscillators**
  - Voltage (or current or digitally) controlled oscillators (VCO, ICO, DCO) are used to generate the local oscillator (LO) signals in a tunable manner.
  - In a communications transceiver (receiver+transmitter, RX+TX), the frequency synthesizer is one of the main blocks. It is used for generating all the needed LO signals in a controllable manner.

- **Analog-to-digital interface**
  - Various ADC technologies, like Flash, Successive Approximation and Sigma-Delta
  - Key performance metrics:
    - Number of bits / Dynamic range / SNR
    - Sampling jitter effects
  - Main bottleneck in advanced DSP-based architectures.
  - Room for innovative solutions

Classical Receiver Architecture: The Superheterodyne

**Example:** One common choice in GSM900 receivers has been 1st IF = 71 MHz, 2nd IF = 13 MHz

Majority of all the receivers have been based on the superheterodyne principle in the past.
Filtering Requirements in Superheterodynes

Selectivity is achieved at the IF stage(s) working at fixed center frequency using special filter technologies.

- The RF filter should provide sufficient attenuation for the image band at the distance of \(2xf_{IF}\) in frequency.
- The final IF stage should have sufficient selectivity to suppress the neighbouring channels sufficiently.
- In case of double (or triple) super heterodyne, the first (and second) IF stage should provide enough attenuation at twice the next IF frequency.
- Image reject (I/Q-) mixer is one possibility to reduce the RF filter requirements (but not sufficient as the only solution)

Some Drawbacks of the Superheterodyne Architecture

Some parts are difficult to integrate

- IF-filter
- RF-filter
- Oscillators

Power consumption high

- External components => parasitics
- Several submodules => low impedance (e.g., 50 \(\Omega\)) levels used for matching the modules

Complicated structure

⇒ There is great interest for simpler architectures which could be integrated more easily.

Spurious responses

- LO and IF signals and harmonics and mixtures leaking to different places may cause problems.
Direct-Conversion Receiver Architecture

“Zero IF” –principle, simply I/Q down-convert the target radio channel directly to baseband.

Advantages

- No image bands, or the image is the signal itself at negative frequencies => RF-filtering not so critical in this sense
- Not so much spurious responses
- Simple structure, no IF filters
- Selectivity can be easily partitioned between analog and digital filters
- If most selectivity in analog parts, low requirements for sampling and ADC blocks

Problem: Difficulties in implementation: dc offsets, second-order intermodulation, leakage between RX and TX in full duplex operation

Direct-conversion principle has become quite popular in recent years in mobile terminals, and is also one widely-used solution in base-stations!
DC-Offsets in Direct-Conversion Receivers

DC-offsets appear mainly due to finite isolation between the mixer RF and LO ports:

Constant DC-offset can be compensated by measuring it without signal and then subtracting it during reception.

In TDMA systems, different channels/bursts may have different signal levels and different AGC-values and hence different DC-offsets => compensation is difficult.

Also 1/f -type of noise appearing in active components may be a problem. This effect also appears at DC and low frequencies.

Other issues

Signal leakage from antenna to the surroundings takes place more easily than in superhet.

Sensitivity to 2nd-order intermodulation is another drawback that will be discussed later in more details.

Low-IF Receiver Architecture

The idea is to use I/Q down-conversion and a low IF frequency which is just high enough to cope with the DC-offset problem (e.g., 250 kHz in case of GSM).

The channel filtering and final demodulation can be done after A/D-conversion in the DSP domain. Channel selectivity can also partitioned between analog and digital front-ends.

As we shall see on the next pages, I/Q down-conversion in the analog domain may not, in practice, provide sufficient attenuation for the image band.

More attenuation can be obtained by using a phase splitter attenuating the image band on the negative part of the frequency axis.

Even more attenuation could possibly be achieved through baseband digital signal processing.

On the other hand, the low-IF concept is facilitated by the fact that system specifications (like GSM) don't allow the maximum signal level to appear in the nearest adjacent channels in case of a very low desired signal level.
Low-IF Receiver by Steyert et al.

In this architecture, the image frequencies are suppressed 25 … 30 dB by the phase splitter (Hilbert) implemented as a polyphase RC network, and another 25 … 30 dB by the I/Q downconversion approach.

In this case, full complex mixer (4 real mixers) are needed as illustrated in the block-diagram.

About Analytic Bandpass Signals

Real bandpass signal and the corresponding ideal analytic bandpass signal (obtained ideally through Hilbert transform):

Other side of the desired signal spectrum suppressed by a practical phase splitter with finite attenuation (e.g., for improving image attenuation in case of I/Q down-conversion):

To realize such a system, it is sufficient to design an allpass filter which approximates 90° phase shift in the passband and stopband with sufficient accuracy, depending on the stopband attenuation requirement.
About Gain and Phase Imbalance in I/Q Systems

Quadrature I/Q downconversion is trying to produce a pure frequency translation which would suppress the image band completely.

In practice, there is some mismatch (imbalance) of gain and/or phase in the components involved (oscillator, amplifiers, mixers).

Consequently, the image suppression is, in practice, far from complete.

Image Rejection as a Function of Gain and Phase Imbalance

Let's take a single complex exponential

\[ \cos \omega_c t + j \sin (\omega_c t) = e^{j \omega_c t} \]

as a starting point.

Assuming now that \( g \) is the gain imbalance ratio and \( \phi \) is the phase difference due to imbalance, we can write:

\[
\cos \omega_c t + j g \sin (\omega_c t + \phi) = e^{j \omega_c t} \frac{1 + ge^{j \phi}}{2} + e^{-j \omega_c t} \frac{1 - ge^{-j \phi}}{2}
\]

From the latter form, we can identify that due to imbalance, the ideal exponential is now split to two mirror exponentials

\[ \Rightarrow \text{this means cross-talk between the mirror or image frequencies in general} \]

The ratio of the image and desired signal powers is obtained as (image rejection ratio, IRR)

\[
R^2 = \frac{|1 - ge^{-j \phi}|^2}{|1 + ge^{j \phi}|^2} = \frac{1 + g^2 - 2g \cos \phi}{1 + g^2 + 2g \cos \phi}
\]
Image Rejection as a Function of Gain and Phase Imbalance

This can be applied to all cases of quadrature mixing where gain and/or phase imbalance appears (dimensioning).
**Frequency-Dependent Imbalance Model**

Especially in wide-band systems, the imbalance parameters may be frequency dependent and can be written as \( g(f) \) and \( \phi(f) \).

Then also the image suppression ratio is frequency-dependent:

\[
R(f)^2 = \frac{1 + g(f)^2 - 2g(f)\cos\phi(f)}{1 + g(f)^2 + 2g(f)\cos\phi(f)}
\]

**Measured example:**

![Graph showing frequency vs. attenuation](image)

**Effects of Gain and Phase Imbalance in Different Architectures**

In case of direct conversion receiver (or final demodulation of an I/Q-signal), gain mismatch and phase errors cause “self images”:

This is usually not a problem with low-order modulations, for which an image attenuation of, e.g., 20 dB doesn’t essentially effect the system performance. For high-order modulations, this issue is more critical.

In other cases of I/Q down-conversion, like the low-IF receiver, the image signal may be at a considerably stronger level (up to 100 dB!!) than the desired signal, and I/Q imbalance is very critical:
Selectivity Requirements in System Specifications

Radio system specifications (e.g., for cellular systems) don't allow a strong adjacent channel signal to be present when a weak desired signal is to be received. (Radio Resource Management functionalities of wireless systems take care of this!)

For example, the GSM specifications give:

- The maximum levels for the 3 adjacent channels on both sides (at 200, 400 and 600 kHz from the carrier) in case of a GSM interferer and desired signal 20 dB above the reference sensitivity level of -102 dBm.

- Maximum levels for more distant signals (>600 kHz from carrier), blocking signals, in case of a sinusoidal interferer and desired signal 3 dB above the reference sensitivity level.

These are usually stemming from linearity constraints but help also in the image problem, especially when the IF is reasonably low.
Summary: Techniques for Providing Image Rejection in Different Architectures

1. RF, first IF filters
   - Challenges to get sufficient performance in integrated solutions.
   - Not applicable in low-IF or direct conversion cases.

2. Quadrature down-conversion
   - Mostly for low-IF and direct conversion cases.
   - Utilized sometimes also in superheterodynes ("image reject mixer") to simplify the RF filter.
   - Rejection limited by gain and phase imbalance.
   - Produces complex signal and the consecutive signal processing blocks must be duplicated, i.e., implemented for both for I and Q branches.

3. Phase splitter (passive polyphase RC network)
   - Mostly for low-IF and direct conversion cases, but can be used also in superhets.
   - Produces complex signal and the consecutive signal processing blocks must be duplicated.

4. Advanced baseband digital signal processing
   - If the imbalance parameters can be estimated, the effects can be compensated fairly well. In recent years, various effective and practical solutions have been developed for this purpose.

Essential Non-Idealities and Performance Measures of the Analog Front-End

Distortions in a receiver front-end are mostly caused by

- Nonlinearities in the front-end that distort the desired signal and create new spectral components as harmonic or intermodulation products of the various spectral components of the received signal.

- Insufficient attenuation of spectral components that are translated in the spectrum due to imaging or aliasing caused by the (inevitable) sampling and A/D-conversion stages.

- Noise produced by the analog stages, sampling process, and A/D-conversion.

Depending on the system specifications, the quality of the front-end should be good enough in two senses:

1. The possible distortion in the desired signal (without other signals) itself should not cause any significant degradation of bit-error-rate (BER) in detection.

2. The distortions in the front-end should not cause new interfering signal components to appear in the signal band that would significantly affect the bit-error-rate (BER) in detection.
Essential Non-Idealities and Performance Measures of the Analog Front-End

Earlier low-order modulations, like QPSK, GMSK, or $\pi/4$ QPSK, were used in wireless communications, and the first criterion was not very critical. However, increasingly higher-order modulations have been taken into use

- 8PSK in Edge
- 16QAM, e.g., in WCDMA/HSDPA
- 64QAM, e.g., in DVB-T, 802.11a/g, WiMAX, 3GPP-LTE

$\Rightarrow$ Bigger and bigger demands for the quality of receiver front-end processing.

Sensitivity of the receiver is mainly determined by the noise produced by the receiver front-end components. It determines the minimum detectable signal in noise-limited situation. In general:

$$\text{Receiver sensitivity} = \text{thermal noise power (dBm)} + \text{noise figure (dB)} + \text{implementation loss (dB)} + \text{required SNR (dB)}$$

In room temperature:

$$\text{Thermal noise power (dBm)} = -174 \text{ dBm} + 10\log_{10} B$$

where $B$ is the equivalent noise bandwidth of the RX.

For example in GSM, minimum S/N is 9 dB, $B=200$ kHz, and required sensitivity is -102 dBm $\Rightarrow$ NF<10 dB.

With low received signal levels (as above), the receiver front-end components can be assumed to be linear. With higher signal levels, the nonlinearity of the amplifiers and other components produce harmful intermodulation products. In this way the nonlinearity limits the dynamic range of received signals from above.

Intermodulation is measured by 1 dB compression point or the so-called $IP^3$ figure.

In RF circuit design, there are always tradeoffs between noise figure, linearity, and power consumption.
Essential Non-Idealities and Performance Measures of the Analog Front-End

- **Frequency accuracy and stability** are determined by the local oscillators of the receiver. In systems like GSM, the receiver is locked to the network, and the frequency stability is very good. However, to guarantee that the receiver is able to synchronize to the network, certain frequency accuracy, stability, and settling time requirements are set for the components.

The short term instability of the oscillators appear as phase noise, and it is very critical for the performance of the system.

- **Leakage** effect means that strong signals, especially local oscillator signals are connected, e.g., through spurious capacitances to places where they are not supposed to be connected. This means that various harmonics, subharmonics, and mixtures of the local oscillator frequencies are usually added to the signal.

In receiver design, the frequencies of the strongest spurious frequencies can be calculated. By selecting the local oscillator frequencies properly, most of the spurious frequencies can be placed outside the desired frequency bands at RF and IF.

- In the case of analog I/Q signal processing, amplitude and phase responses of the I and Q branches are never exactly the same. The effects of the gain and phase imbalance depend greatly on the receiver architecture.

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### Noise Figure

The noise factor of an amplifier stage (or some other component) is determined by the ratio of S/N ratios at the input and output:

$$F = \frac{SNR_{in}}{SNR_{out}}$$

The noise figure is:

$$NF = 10 \log_{10} F.$$  

For a cascade of $n$ stages, the overall noise factor is

$$F_T = \frac{F_1 + \frac{F_2 - 1}{g_1} + \frac{F_3 - 1}{g_1 g_2} + \cdots + \frac{F_n - 1}{g_1 g_2 \cdots g_{n-1}}}{g_1 g_2 \cdots g_{n-1}}$$

where $F_i$ and $g_i$ are the noise factor and power gain of stage $i$.

Typically the noise figure of the first components (before significant amplification) is most critical in receivers, because the target signal level is weakest in those stages.
Intermodulation

Consider a test where there are two nearby frequencies $f_1$ and $f_2$ in the system frequency band (like in the neighbouring channels).

In general, nonlinearities produce intermodulation frequencies of the form

$$k_1 f_1 + k_2 f_2$$

where $k_1, k_2$ are integers.

Third-order intermodulation produces (among some others) frequencies $2f_1 - f_2$, $2f_2 - f_1$ which may easily fall in the desired signal band.

Second-order intermodulation produces frequencies $f_1 + f_2$, $f_2 - f_1$.

In general, with low-enough signal levels, the levels of second- and third-order intermodulation products are proportional to the 2nd and 3rd power of the fundamental signal level, respectively.

Third-order IMD of RF modules is always a major concern. Depending on the radio architecture, also second- and third-order products of analog IF modules may be problematic.

Intercept points - IP2, IP3

The device-specific IP2, IP3-values can be used to determine the strength of second- and third-order intermodulation products, in reference to the fundamental signal component.

The overall IP2 and IP3 values for a cascade of stages can be calculated in a fairly straightforward way from IP values of the components.

These calculations are mostly relevant in the first receiver stages, before the channel selection filtering, where there are strong adjacent channel/blocking components present.

Specifications for the IP values can be derived from the knowledge of the maximum signal levels in adjacent frequency channels and the minimum signal level and required SNR of the wanted signal.
**IP2, IP3 (cont’d)**

In general, if the fundamental (target) signal level (at a certain point in the receiver chain) is \(x\) [dBm] and we know the IP3-value calculated from the preceding stages, then third-order intermodulation products are \(2(\text{IP3} - x)\) [dB] below the fundamental signal level (assuming, of course, that \(x \ll \text{IP3}\)).

This gives the basis for relatively easy dimensioning in receiver system calculations.

These calculations are mostly relevant in the receiver stages before the channel selection filtering, where there are strong adjacent channel/blocking components. In the calculations, the fundamental signal level is determined by the worst-case blocker level.

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**Example of IP3 Calculations**

Assume that the IIP3 value of a receiver is 10 dBm.

Assume further that there are two strong blocking signals at nearby frequencies, causing 3rd-order intermodulation to the desired signal band.

Suppose the radio system specifications allow the blocking signals to be at 60 dB higher level than the desired signal.

Finally, we assume that 10 dB signal to interference ratio (SIR) is sufficient at detection interface.

Then we can calculate that if the blocking signal level is 35 dB below the IP3 point, the intermodulation product is 70 dB below the blocking signal level, and the 10 dB SIR target is exactly satisfied.

This means that the maximum tolerated blocking signal level at the input of the receiver is -25 dBm.

In practice, some implementation margin against other interference and noise is of course needed.
Effects of 2nd and 3rd-Order Intermodulation

2nd-order intermodulation products are generally clearly stronger than 3rd-order products.

=> If 2nd-order intermodulation products of some strong signals (blocking signals) appear on the signal band, then better linearity is required.

- In typical superheterodyne receivers, only 3rd-order intermodulation is a problem, because signals causing 2nd-order products on the signal band are attenuated by the RF filter.

- In wideband superhet with relatively low IF, also the 2nd-order intermodulation may become a problem.

- 2nd-order intermodulation is always a problem in direct-conversion and low-IF receivers.

Non-Idealities in Oscillators

In-band effects (i.e., may exist even if no adjacent channels and blockers are present):

- Constant phase error rotates the constellation; This can be corrected by baseband processing afterwards

- Phase noise: random fluctuations in the instantaneous phase/frequency of the oscillator cause random constellation rotations, reducing the noise margin and increasing BER:

The RF effects of phase noise tend to be far more critical than in-band effects:

- Mixing products of the phase noise spectrum and strong adjacent channel signals (reciprocal mixing) may produce spurious signals which overlap the desired signal. The following figure shows the noisy LO spectrum, RF-spectrum, and spectrum after mixing with practical LO with phase noise.
Example of Phase Noise Calculations

VCO phase noise in a GSM 900 terminal

- wanted signal: -102 dBm + 3 dB = -99 dBm
- blocking signal: -43 dBm @ 600 kHz
- Minimum S/N=9 dB
- target signal bandwidth: 200 kHz

Assume that the phase noise spectrum is flat within the noise bandwidth with x dBc/Hz (i.e., the noise power in 1 Hz bandwidth is x dB in reference to the power of the VCO at the LO frequency)

\[ \text{noise power at target signal bandwidth: } -43 \text{ dBm + x + } 10 \log_{10} 200000 < -99 - 9 \text{ dBm} \]

\[ x < -118 \text{ dBc/Hz (@600 kHz)} \]
About the Co-Operation of Transmitter and Receiver

**Duplexing Techniques**

Uplink and downlink signals can be separated using different carrier frequencies or different time slots:

- **In Time Division Duplex (TDD)**, TX and RX use the same frequency but different time slots.
- **In Frequency Division Duplex (FDD)**, RX and TX are working at the same time but use different carrier frequencies.

The GSM system is based on FDD, but it has also a TDD element in a sense through nonoverlapping sequencing of TDMA slots in uplink and downlink (in individual terminal).

- In the basic system, terminal TX and RX are not working at the same time.
- In multislot services (HSCSD, GPRS) with high data rate, the RX and TX may need to operate at the same time.

UMTS system, including the enhancements (GPRS, LTE), is based on the FDD principle and the transmitter and receiver are usually operating at the same time. (There is also a TDD variant of UMTS, but it is not currently in use.)

WLAN and WiMAX systems are based on TDD.

Simultaneous Operation of Transmitter and Receiver

There are two ways of sharing the antenna between the transmitter and receiver:

- **Switching**: Connecting the antenna to RX or TX as needed.
- **Duplexer**: Using a band-splitting filter to connect the RX band to the receiver and transmitter to the TX band.

As a passive device, the duplexer attenuates the received signal and thus increases the noise figure of the receiver considerably. Better noise figure can be achieved with switching, even though passive filtering is usually needed also in this case both in RX and TX.

If the TX and RX are not working at the same time, then

- Either switching or duplexer can be used.
- The same frequency synthesizer can be used for RX and TX (if the settling is fast enough).
- Less spurious frequencies, because TX is silent when RX is working.

If the TX and RX are working at the same time, then

- Duplexer has to be used.
- The frequency synthesizer has to produce all the frequencies needed by TX and RX at the same time.
- Controlling the spurs gets more difficult; isolation between TX and RX is a big concern.
Critical Issues on Transmitter Side

Sufficiently bandlimited radio waveform at target center-frequency (tunability) and at desired output power.

Main performance criteria

- Spectral purity
- Power efficiency
- Signal distortion

Transmitter architecture

- Two-stage conversion (with IF) is a safe solution, but difficulties in integration.
- I/Q imbalance is a problem in "low-IF" type of architectures.
- Direct conversion has problems, like leakage from power amplifier output to LO may have bad effects.
- Direct Digital Synthesis (DDS) is a promising approach, but the DA-converter performance is still a problem in mobile systems.
- DSP based enhancement of the signal quality can be used also on the transmitter side; digital pre-distortion.

Power Amplifier (PA)

- For constant envelope modulations (FSK, GFSK, MSK, GMSK) nonlinear power amplifiers can be used and good power efficiency is achieved.
  - With exactly constant envelope signals, nonlinearities produce only harmonics, but no other new spectral components.
- In other cases (linear modulations, and especially CDMA and OFDM), linear power amplifier is needed.
  - Nonlinearity causes spectral regrowth and signal distortion.
- Back-off measures the headroom between the average transmitted signal power and the maximum (saturated) output power of the PA.
  - Increasing back-off improves linearity, but reduces power-efficiency.
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### Sampling Theorem

The sampling theorem says that a (real or complex) lowpass signal limited to the frequency band \([-W, W]\) can be represented completely by discrete-time samples if the sampling rate \((1/T)\) is at least \(2W\).

In case of a complex signal, each sample is, of course, a complex number.

In general, discrete-time signals have periodic spectra, where the continuous-time spectrum is repeated around frequencies \(\pm 1/T, \pm 2/T, \pm 3/T, \ldots\)

In case of complex signals, it is not required that the original signal is located symmetrically around 0.

Any part of the periodic signal can be considered as the useful part. This allows many possibilities for multirate processing of bandpass signals.

In general, the key criterion is that no destructive aliasing effect occur on top of the desired part of the spectrum.
Real vs. Complex Discrete-Time Signals

**Real signal:**

Here $2W$ real samples per second are sufficient to represent the signal.

**Complex signal:**

Here $W$ complex samples per second are sufficient.

- The resulting rates of real-valued samples are the same.
- However, the quantization effects may be quite different. (Recall from the standard treatment of SSB that Hilbert-transformed signals may be difficult.)

Real Bandpass Sampling

Down-conversion can also be implemented by sampling a bandpass signal. Any part of the periodic spectrum can be selected for further processing.

Concerning the sampling frequency, it is sufficient that no aliasing appears on top of the desired band.

In general, the feasible sampling frequencies are determined from $W$, $B$ (useful signal bandwidth), and $f_s$.

Minimum sampling frequency is $B+W$, which is adequate in the case where the center frequency of the desired signal is $kf_s \pm f_s/4$:
Quadrature Sampling

In this case we are sampling the complex analytic signal obtained by a phase-splitter:

\[ f_{\text{c}} - W/2 \quad f_{\text{c}} \quad f_{\text{s}} \quad k f_{\text{s}} \quad (k+1)f_{\text{s}} \quad f \]

(in above assumed again that \( f_{\text{c}} = k f_{\text{s}} + W/2 \))

Of course, in practice the image bands in between the shown periodic replicas are not completely attenuated but only suppressed to a level that is determined by the amplitude and phase imbalances in the phase splitter & sampler & ADC blocks.

The gain and phase imbalance analysis of quadrature down-conversion applies also to this case.

Second-Order Sampling

Quadrature sampling can be approximated by the following structure:

At the carrier frequency, the sampling time offset corresponds exactly to the 90° phase shift. Farther away from the center frequency this is only approximative, but for relatively narrowband signals, it works.

The nonideality can be evaluated using the phase imbalance analysis.
Analysis of Second-Order Sampling

The second-order sampling concept works perfectly at the carrier frequency (ignoring the other sources of I/Q imbalance) but only approximately at other frequencies. At frequency $f_c + f_\Delta$, a time-shift of $1/(4f_c)$ corresponds to a phase shift of

$$\frac{1}{4f_c} \cdot 2\pi = \left[1 + \frac{f_\Delta}{f_c}\right] \frac{\pi}{2} \text{ rads}$$

We are actually dealing with phase imbalance and the image rejection formula for quadrature mixing can be utilized. The resulting image rejection is:

$$R = \frac{1 - \cos \left( \frac{f_\Delta}{f_c} \cdot \frac{\pi}{2} \right)}{1 + \cos \left( \frac{f_\Delta}{f_c} \cdot \frac{\pi}{2} \right)}$$

**Example:** $f_c = 1$ GHz

<table>
<thead>
<tr>
<th>$f_\Delta$</th>
<th>$f_\Delta/f_c$</th>
<th>Phase imbalance</th>
<th>Image rejection</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.1 MHz</td>
<td>0.0001</td>
<td>±0.009°</td>
<td>82.1 dB</td>
</tr>
<tr>
<td>1 MHz</td>
<td>0.001</td>
<td>±0.09°</td>
<td>62.1 dB</td>
</tr>
<tr>
<td>10 MHz</td>
<td>0.01</td>
<td>±0.9°</td>
<td>42.1 dB</td>
</tr>
<tr>
<td>100 MHz</td>
<td>0.1</td>
<td>±9°</td>
<td>22.1 dB</td>
</tr>
</tbody>
</table>

Problems with Wideband Sampling

**Analog to Digital Converter (ADC)**

Sampling a wideband signal, containing several channels is a tempting approach for designing a flexible radio receiver. However, there are some great challenges to do this.

The strongest signal in the ADC input signal band should be in the linear range of the ADC. When the desired signal is weak, a large ADC dynamic range is needed, the resolution of the converter has to be many bits, e.g., 14 ... 17 bits.

**Sampling**

The sampling to get a discrete time signal is done usually with a track-and-hold circuit (T/H).

In practical sampling clocks and sampling circuits, there are unavoidable random variations in the sampling instants, **sampling aperture jitter**. In bandpass sampling, the requirements for aperture jitter become very hard.
Quantization Noise in ADCs

In general, the maximum S/N-ratio for an A/D-converter is

\[
\text{SNR} = 6.02n + 4.76 - CF_{dB} + 10 \log_{10} \left[ \frac{f_s}{2B} \right] \quad \text{(dB)}
\]

where

- \( n \) is the number of bits
- \( CF_{dB} \) is the Crest Factor in dB
- \( B \) is the useful signal bandwidth
- \( f_s \) is the sampling rate.

Crest factor for a voltage signal is defined as the ratio of the peak absolute value and the RMS-value. The maximum SNR is achieved when the signal utilizes the A/D-converter’s full voltage range, which is assumed to be symmetric around DC. For a sinusoidal signal, the crest factor is 3 dB.

The last term takes into account the processing gain due to oversampling in relation to the useful signal band. When the quantization noise outside that useful signal band is filtered away, the overall quantization noise power is reduced by the factor \( f_s/2B \).

The number of additional bits needed to quantize a wideband signal can be estimated by:

\[
10 \log_{10} \left[ \frac{CF_B^2 \cdot P_B}{CF_d^2 \cdot P_d} \right] / 6 \quad \text{bits}
\]

where

- \( P_B \) is the worst case power in the full band
- \( P_d \) is the minimum useful signal power
- \( CF_B \) is the crest factor in the worst case test situation
- \( CF_d \) is the crest factor of the desired signal.

Usually, in communications receivers, the worst case power is determined from the adjacent channel or blocking specifications. For many types of communications signals (e.g., CDMA, OFDM), the crest factor is much higher than that of the sinusoid, which might be a valid assumption in the blocking test situation.

Spurious-Free Dynamic Range

Practical ADC’s have also discrete spectral frequency components, spurious signals (or spurs), in addition to the flat quantization noise.

In many applications, the spurious-free dynamic range, SFDR, is the primary measure of the dynamic range of the converter.
Track&Hold Circuit Nonidealities

Advanced bandpass sampling approaches could mean that we are sampling a tens-of-MHz to GHz-range signal with a relatively low sampling rate.

Noise Aliasing

Wideband noise at the sampling circuitry will be aliased to the signal band. In case of bandpass sampling, aliasing increases with increasing subsampling \((f_s/f_c)\) factor. Basically, the noise figure depends on the subsampling factor.

Therefore, it is important to have a good noise figure for the track&hold circuit and/or to have sufficient amplification in the analog front-end.

Aperture Jitter

Aperture jitter is the variation in time of the exact sampling instant, that causes phase modulation and results in an additional noise component in the sampled signal.

Aperture jitter is caused both by the sampling clock and the sampling circuit.
SNR Due to Sampling Jitter

The noise produced by aperture jitter is usually modelled as white noise, which results in a signal-to-noise ratio of

\[
SNR_{aj} = 20 \log_{10} \left( \frac{1}{2\pi f_{\text{max}} T_a} \right)
\]

where \(f_{\text{max}}\) is the maximum frequency in the sampler input and \(T_a\) is the RMS value of the aperture jitter.

This model is derived for a sinusoidal input signal, but applied also more generally. In critical test cases of the wideband sampling receiver application, the blocking signal is often defined as a sinusoidal signal, and the model is expected to work reasonably well.

The processing gain due to oversampling effects in the same way as in case of quantization noise.

Example of Sampling Jitter Effects

- 14 bits
- 1 ps RMS jitter

About A/D-Conversion for SW Radio

It is obvious that the requirements for the T/H-circuit and A/D-converter are the main bottlenecks for implementing receiver selectivity with DSP.

One promising A/D-converter technology in this context is the sigma-delta (ΣΔ) principle.

- This principle involves low-resolution, high-speed conversion in a noise-shaping configuration, together with decimating noise filtering.
- In case of lowpass and bandpass sampling with suitable fixed center frequency, this principle can be combined nicely with the selectivity filtering part of the receiver.

Noise filtering in basic ADC:

Noise filtering in sigma-delta converter:
Summarizing Selectivity Tradeoffs

1. **Selectivity at IF (superhet)**
   - high-cost IF filters
   - less demands for analog circuits after IF
   - simple A/D converter

2. **Selectivity by analog baseband processing**
   (direct conversion case)
   - no costly IF filters
   - more RF gain needed
     => RF has to very linear to avoid intermodulation effects
   - simple A/D converter

3. **Selectivity by baseband digital filtering**
   (direct conversion and low-IF cases)
   - no costly IF filters
   - more flexible than case 2: selectivity can be easily adapted to different systems
     - more RF gain with high linearity needed
     - high dynamic range A/D-converters (14…17 bits)

4. **Selectivity by digital filtering after wideband IF sampling**
   - simplified IF filter
   - high flexibility
     - suitable for multi-channel receivers with common analog parts
     - high dynamic range A/D-converters (14…17 bits)
   - very strict demands for low jitter sampling clock

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Contents

1. Background and introduction
2. Basic receiver architectures and their properties
   - Main components; frequency translations and filtering
   - Receiver architectures: superhet, direct conversion, low IF
   - Mirror-frequency considerations and effects of I/Q imbalance
   - Non-idealities and performance measures of the analog front-end: sensitivity, noise figure, intermodulation, phase noise
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Classification of DSP-Based Receiver Architectures

Location of ADC (A/D-converter)

1. Baseband or low IF
2. IF
3. RF

Analog front-end bandwidth - ADC bandwidth

1. Single channel
2. Few channels
3. Frequency slice
4. Service band (e.g., GSM)
5. Frequency band (like 2 GHz range)

The bandwidths of the analog front-end and ADC may or may not go hand in hand. Some examples below.

Narrowband front-end
Per-channel down-conversion (0, or IF)

- Selectivity in analog part, good IF/BB filters needed.
- "Normal" frequency synthesizer needed.
- No big demands for ADC dynamic range or jitter.
- Sampling rate requirements are such that it is enough to attenuate aliasing to the desired band.
- Narrowband (baseband or bandpass) ADC (like $\Sigma\Delta$) can be utilized.
- Can in general build on real mixing (IF) or I/Q mixing (0 or IF), depending on the characteristics of RF filtering.
Wideband front-end, per-channel down-conversion so that desired channel is around a fixed center frequency (0 or IF)

- Analog front-end simplified in the sense that highly selective IF filters are not needed.
- "Normal" frequency synthesizer needed.
- Selectivity in digital part, with fixed center frequency.
- Higher demands for ADC dynamic range and jitter.
- Sampling rate requirements are such that it is enough to attenuate aliasing to the desired band.
- Narrowband ADC (like $\Sigma\Delta$) can still be utilized (in the sense that high accuracy is only needed at fixed band).
- Can in general build on real mixing (IF) or I/Q mixing (0 or IF), depending on the characteristics of RF filtering.

Wideband front-end, wideband (like slice or service band) down-conversion so that desired channel is located in a wider frequency range

- Analog front-end simplified in the sense that highly selective IF filters are not needed.
- Only single or few LO frequencies needed for each service band -> simplified synthesizer; fast frequency hopping becomes feasible through digital tuning.
- Selectivity in digital part, with tunable center frequency.
- High demands for ADC dynamic range and jitter.
- Sampling rate requirements are such that no aliasing into the whole band is allowed (or needs to be tuned).
- Wideband ADC (or $\Sigma\Delta$ with tunable center frequency!?) needed.
- Can in general build on real mixing (IF) or I/Q mixing (0 or IF), depending on the characteristics of RF filtering.
About the Choice Between Lowpass and Bandpass Sampling

Due to the I/Q gain and phase imbalance problems in practical analog circuitry, the wideband downconversion - wideband sampling approach is very difficult to implement at 0 (or low) IF. But utilizing a combination of different techniques for mitigating these effects, the mentioned approach is becoming feasible, but mostly on the base-station side.

On the other hand, wideband IF sampling is very challenging due to the aperture jitter and other implementation problems concerning the sampling circuitry (usually track&hold).

The ADC requirements (apart from the sampling process) concern mainly the spurious-free dynamic range, and are not so heavily depending on the choice between lowpass or IF sampling.

However, the useful ADC bandwidth has a great impact, e.g., on power consumption. So the cost/complexity metrics for per-channel A/D-conversion (usually ΣΔ) and multichannel A/D-conversion (usually something else than ΣΔ) are quite different.

About Direct Sampling Architecture

In high-performance systems, it is necessary to have some selectivity and gain before sampling. The reasons are

- signal aliasing
- noise aliasing

Sampling is inherently more noisy operation than mixing!

Sampling directly from the antenna signal is usually not adequate.

The Ultimate SW Radio Architecture

The needed technologies are not mature for challenging radio system specifications in the frequency bands used in mobile systems!

However, direct sampling is already an interesting architecture in various applications

- For example, satellite-based positioning (GPS/ Galileo) where the dynamic range requirements are greatly reduced comparing with wireless communications.
Direct Sampling & Analog Discrete-Time Processing

Texas Instruments (TI) has introduced so-called digital radio processor (DRP) concept that is based on direct sampling, together with analog discrete-time processing to implement main part of the channel selectivity, down-conversion, and sampling rate reduction.

- For example, CIC/running-sum filters can be implemented with switched-capacitor techniques with analog processing.
- Then the ADC is operating at relatively low rate and has reduced dynamic range requirements compared to digital direct-sampling approach.

Such designs have so far been considered for systems with low-order modulation and relatively narrow bandwidth or reduced dynamic range, like GPS, Bluetooth, and GSM/GPRS.

In such architectures, also the sampling process may be designed to provide frequency selectivity. Then the idea of the sampling process is not anymore just taking instantaneous sample values, but to

- Integrate the signal over a finite-length interval
- Weighting the input signal by a proper window during the integration interval. Rectangular window results in sinc-response, other kind of windows can be designed for optimized performance.

Some Dependencies and Conclusions

Considering sampling and A/D-conversion

- highest signal frequency determines the T/H bandwidth and jitter requirements
- signal bandwidth (after analog RF/IF/baseband filtering) determines the minimum sampling rate

Increasing the degree of bandpass subsampling \( f_c / f_s \) leads to

- lower sampling rate
- more selectivity needed before sampling
- more noise aliasing => more gain needed before sampling
- lower processing gain -> more bits from ADC & tighter jitter requirements

Implementing the receiver selectivity in DSP-part leads to

- simplified analog part
- hard requirements for the T/H and ADC dynamic range

Wideband sampling

- has been mostly considered at IF due to I/Q-imbalance problems; direct conversion/low-IF becoming feasible, depending on system specs

Using IF sampling

- sets hard requirements for the T/H circuitry and jitter of the sampling clock.
Connection to Advanced Broadband Wireless System Developments

The latest and future wireless communication systems use increasing bandwidths for data transmission. For example, 3GPP-LTE and WiMAX have the maximum bandwidth of 20 MHz, and the next generation ("IMT-advance") is targeted to bandwidths of up to 100 MHz.

Especially, LTE is using frequency-division multiplexing, and the spectrum entering the receiver resembles that of a multichannel receiver for more narrowband systems. Some characteristics and comments:

- The wide bandwidth makes it possible to utilize fast frequency hopping and other forms of frequency diversity, to enhance the transmitted data rate.
- In LTE (and other similar systems) the power levels of the frequency slots of different users are well-controlled (e.g., 20 dB maximum variation in the power levels). This is in contrast to, e.g., multichannel GSM receiver, were the dynamic range is much bigger. This makes it feasible to implement the needed wideband receivers for such systems.
- Direct conversion architecture is preferred. Actually, for most of the frequency channels, the low-IF model is valid.
- IQ-imbalance is significant, but not very critical because of the well-controlled power levels. DSP-based IQ-imbalance compensation is interesting in case of high-order modulations.
- In these systems, and in OFDM systems in general, the frequencies at or close to DC in baseband processing are commonly not utilized in order to make direct conversion receiver design easier.

RF Challenges in Cognitive Radio

As seen above, the RF implementation challenges can be greatly relieved in radio environments, where the signal levels are well-controlled by the radio resource management functionalities.

A completely different case is the opportunistic spectrum use in the cognitive radio context. Here the spectral gaps (so-called white spaces) between licensed users (primary users) are intended to be used locally by secondary users.

The secondary system has no control of the power levels in adjacent frequency slots, and it should also operate at relatively low power levels, in order not to introduce interference to the primary users.

The needed flexibility regarding the signal waveforms, bandwidths, and centre frequencies can be only achieved through software radio techniques. In order to efficiently utilize the spectral gaps, a cognitive radio receiver should be able to operate in the presence of strong primary signals in adjacent frequency slots.
Beamforming and Dynamic Range

Among the various multiantenna techniques, beamforming on the receiver side helps to reduce the dynamic range of the received signal. This is achieved by directing zeros of the radiation pattern towards strong interfering transmitters.

Naturally, to achieve this benefit in the analog-to-digital interface, the combination of the antenna signals should be done at the analog RF stage.

Multimode Terminal Receivers

In flexible multi-mode receivers, the target is to use common blocks for different systems as much as possible.

A long-term target is to make the transceiver configurable for any system. However, presently a combination of a few predetermined systems is more realistic, e.g., GSM/3GPP/WLAN.

A realistic approach has the following elements:

- Separate RF stages for different systems.
- Common IF/baseband analog parts; bandwidth according to the most wideband system.
- Common ADC at IF or baseband; fixed sampling rate.
- Especially in the terminal side: careful choice of IF frequency & sampling rate to make the down-conversion simple. Typically, $f_{IF} = (2k+1) f_s / 4$.
- Digital channel selection filtering optimized for the different systems.

![Multimode Terminal Receivers Diagram]
DSP for Flexible Receivers

In advanced SW radio concepts, the selectivity filtering and down-conversion are moved from analog continuous-time part to the discrete-time/DSP part.

⇒ Here efficient multirate filtering techniques become very important.

It also helps to move as much as possible functionality from the analog or digital front-end to baseband processing.

⇒ All-digital synchronization concept becomes very interesting in this context.
  o Free-running local oscillators for demodulation and frequency conversion.
  o Free-running sampling clock.
  o Errors are compensated in digital part.

Some errors due to RF-front-end can be corrected by DSP.

⇒ Dirty RF paradigm
  o I/Q imbalance compensation
  o Compensation of intermodulation effects
  o etc.

Digital Channel Selection & Down-Conversion

Digital Down-Conversion

1. Desired channel centered at fixed IF

⇒ Fixed down-conversion
  Special choices of $f_{IF}$ and $f_s$ make things easy.
  Especially when $f_{IF} = (2k+1) f_s / 4$, the signal aliases to $f_s / 4$ and down-conversion is very easy.

2. Wideband sampling case

⇒ Tunable down-conversion and NCO (numerically controlled oscillator) needed.

Also frequency translation due to sampling rate conversion operations can be utilized.

There are possibilities to move the I/Q mixer and NCO closer to baseband, which helps to reduce the power consumption.

Channel Selection Filtering

- After down-conversion, efficient lowpass decimator structure is needed.

- Multiplierles CIC-filters are commonly used in the first decimation stages, FIR-filters and the last stages. $N$th-band IIR filters also an efficient solution.

Adjusting Symbol Rates

- Different systems use different symbol/chip rates.
- Common sampling clock frequency is preferred.
  ⇒ Decimation by a fractional factor is needed.
- This can be done at baseband or earlier in the decimation chain.
Further reading

Receiver architectures: [1] - [9]
Low-IF architecture: [9]
Transmitter issues: [14]
Direct digital synthesis: [15]
Software radio architecture: [12], [13], [16], [17]
Advanced sampling & analog discrete-time processing: [18] – [22]
Cognitive radio challenges: [23]

Literature