Software-Defined Radio

Architectural Design of Digital Receivers

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Structure of the course

- Part 1: Architectural design of digital receivers
  - Introduction
  - Basics on digital receivers
  - Digital mixers and numerical oscillators
  - Digital modulation and pulse shaping
  - Sampling and AD conversion

- Part 2: Parameter estimation and synchronization in digital receivers

Basic idea of SDR

- Motivation and background
  - Radio with functionality defined by software
  - Radio for different standards: GSM, DECT, CDMA, UMTS, …
  - Programmable on-board processing units for satellites
  - Cognitive radios to avoid interference
- Barriers and objections
  - Lifetime of mobiles is 1-2 years
  - Mobiles must be very cheap and not fancy
  - Programmable hardware (FPGA, DSP, µP) not powerful enough
  - Customized chips (ASICs) much cheaper for mass market

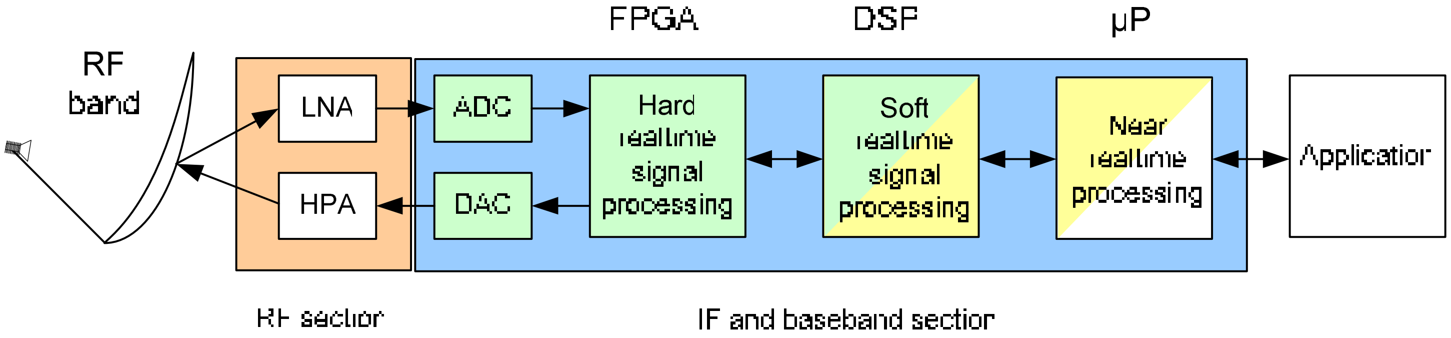
Benefits of SDR solutions

- Flexibility
  - Speed factor (time to market)
  - Easy testing of new algorithms
  - Design and implementation of HW for several product life cycles
- Costs
  - ASICs much too expensive for niche markets
  - Provide more than one product with a single hardware
  - Remote upgrade facility
- Availability
  - Powerful and cheap FPGAs, DSPs, and general purpose (ARM) processors provided by different manufacturers

Definition of the ideal SDR

- ADCs and DACs placed next to the antenna
- ADCs and DACs directly connected to FPGAs or DSPs
- General purpose machines for further processing
- Data delivered to applications

Block diagram of the ideal SDR



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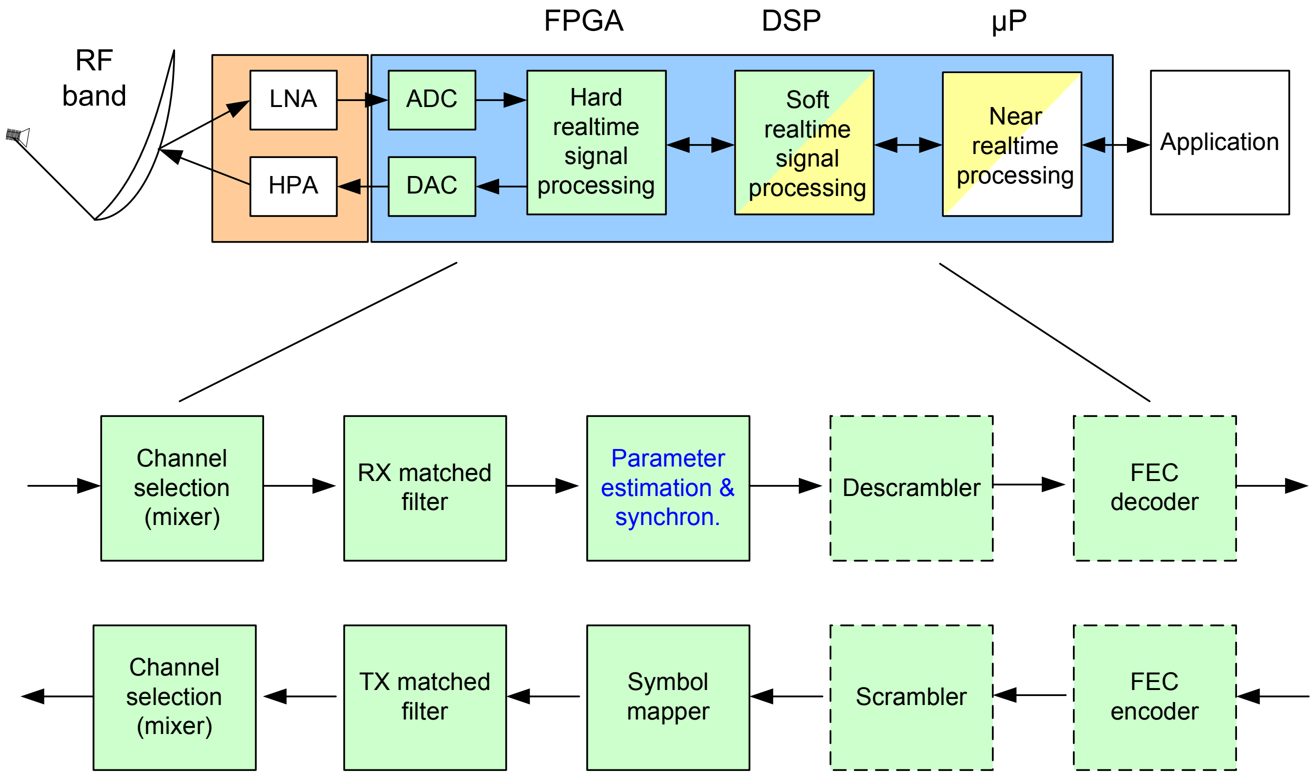
RX and TX function blocks



FPGA
DSP
µP

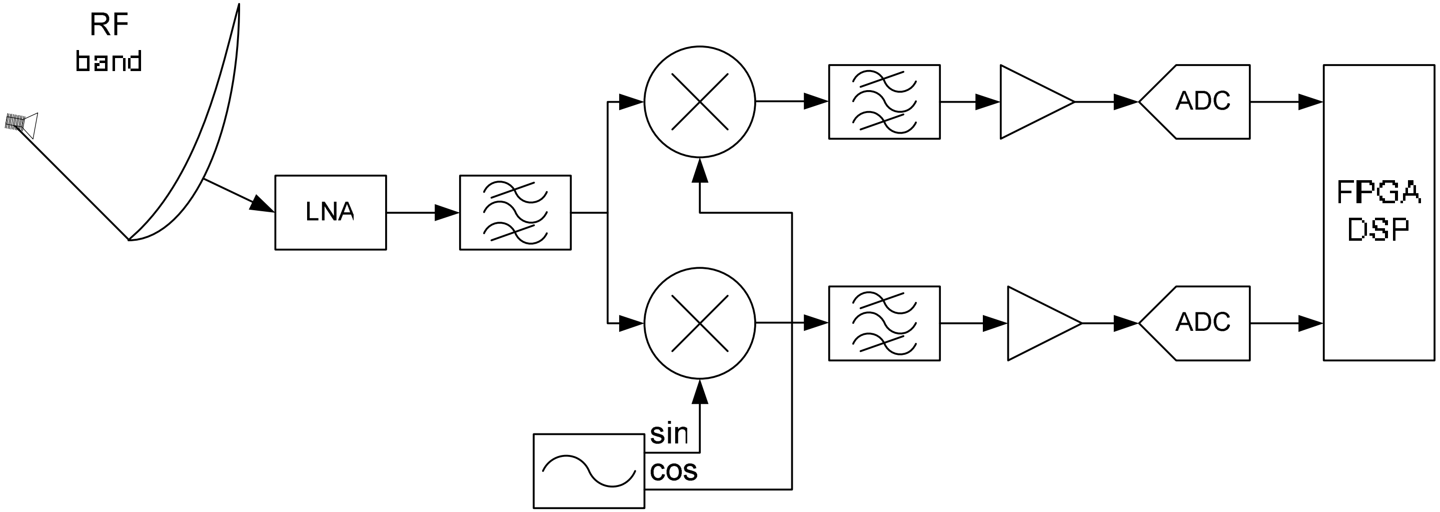
RF
band

LNA
HPA
ADC
DAC
Hard realtime signal processing
Soft realtime signal processing
Near realtime processing
Application

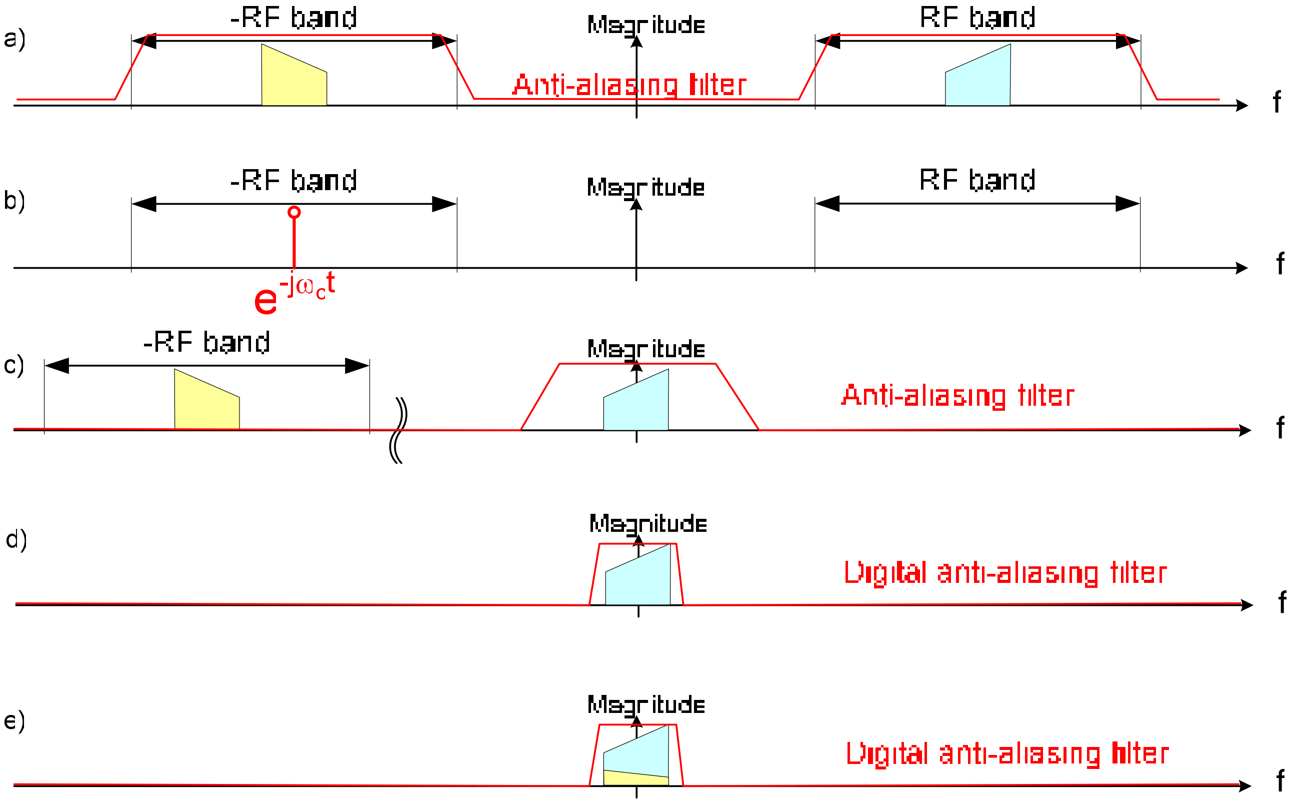
Channel selection (mixer)
RX matched filter
Parameter estimation & synchron.
Descrambler
FEC decoder

Channel selection (mixer)
TX matched filter
Symbol mapper
Scrambler
FEC encoder

Direct conversion architecture



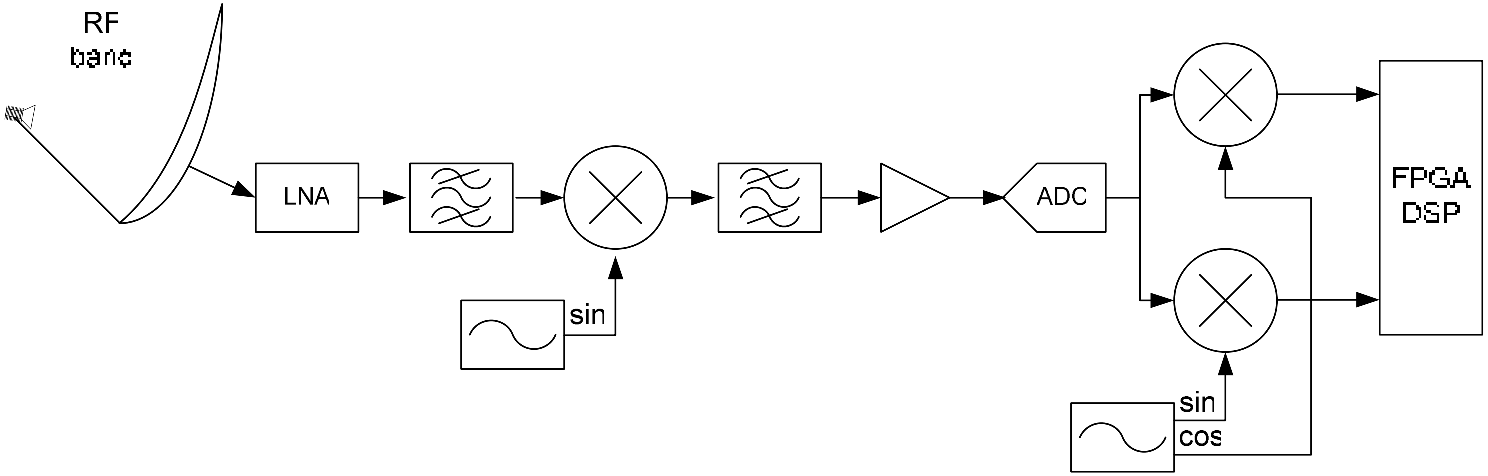
Spectrum


a)
-RF band
Magnitude
RF band
Anti-aliasing filter
f
b)
-RF band
Magnitude
RF band
$e^{-j\omega_c t}$
f
c)
-RF band
Magnitude
Anti-aliasing filter
f
d)
Magnitude
Digital anti-aliasing filter
f
e)
Magnitude
Digital anti-aliasing filter
f

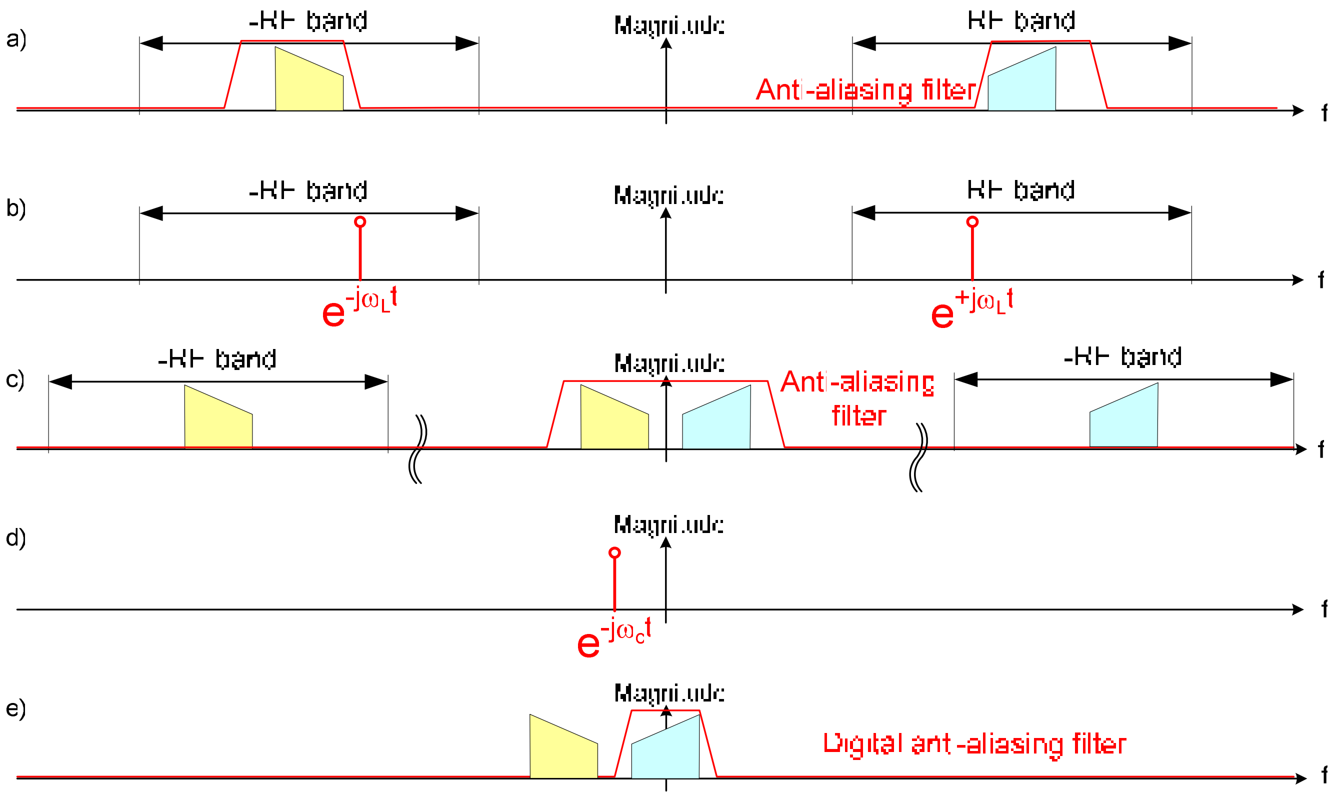
Notes on direct conversion

- Benefits
  - Low complexity
  - Suitable for ASIC solutions
  - Simple filtering requirements
- Drawbacks
  - Totally balanced ADC required
  - Oscillator parameters must be constant over the RF band
  - I/Q balance (sideband and XT suppression)
  - DC offset problem

Multi-stage conversion architecture



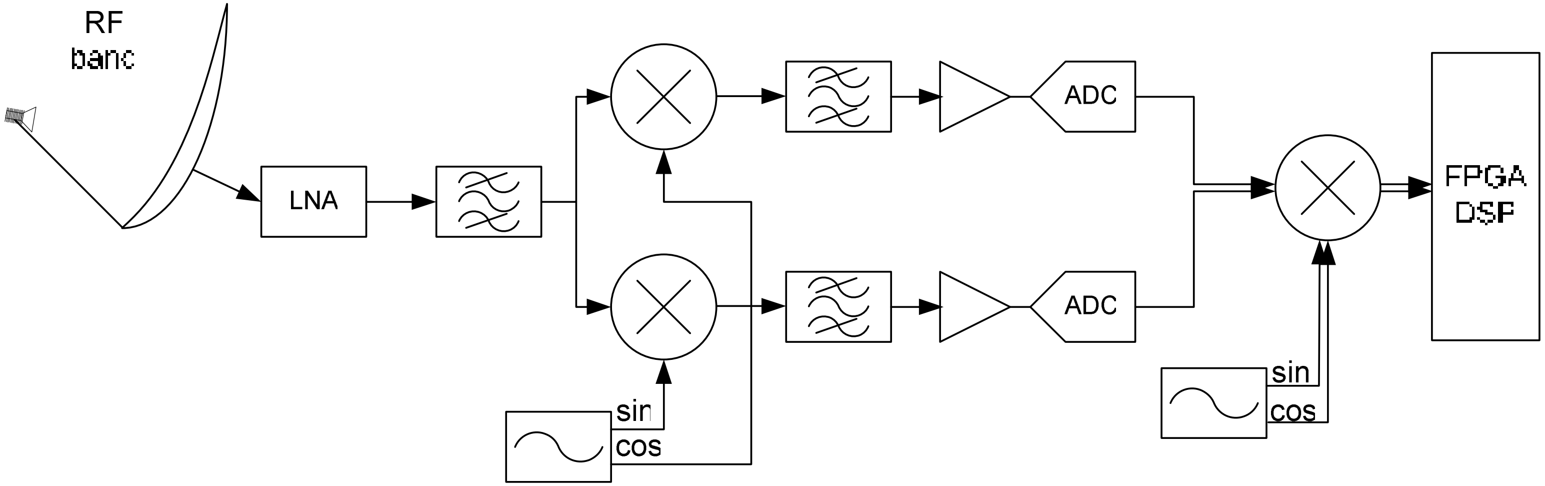
Spectrum


a)
-RF band
Magnitude
RF band
Anti-aliasing filter
f
b)
-RF band
Magnitude
RF band
$e^{-j\omega_L t}$
$e^{+j\omega_L t}$
f
c)
-RF band
Magnitude
Anti-aliasing
filter
-RF band
f
d)
Magnitude
$e^{-j\omega_c t}$
f
e)
Magnitude
Digital anti-aliasing filter
f

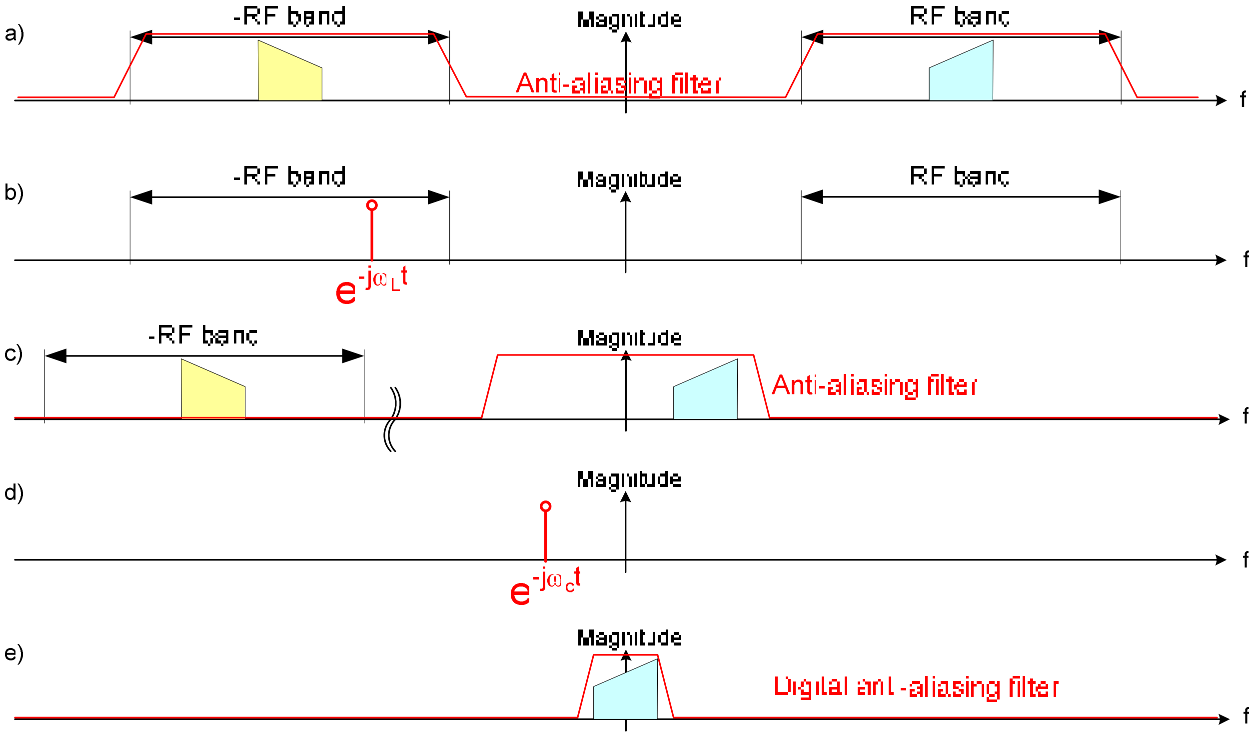
Notes on multi-stage conversion

- Benefits
  - No DC offset
  - Only one ADC required
  - I/Q separation in the digital domain
- Drawbacks
  - Higher complexity
  - Several local oscillators needed (higher phase noise)
  - Higher ADC sampling rates (powerful FPGAs needed)

Low-IF architecture



Spectrum


a)
-RF band
Magnitude
RF band
Anti-aliasing filter
f
b)
-RF band
Magnitude
RF band
$e^{-j\omega_L t}$
f
c)
-RF band
Magnitude
Anti-aliasing filter
f
d)
Magnitude
$e^{-j\omega_c t}$
f
e)
Magnitude
Digital anti-aliasing filter
f

Notes on low-IF conversion

- Benefits
  - No DC offset
  - Reduced filter complexity in the analog domain
- Drawbacks
  - Higher effort (complex mixers)
  - Higher I/Q balance required (sideband, XT)

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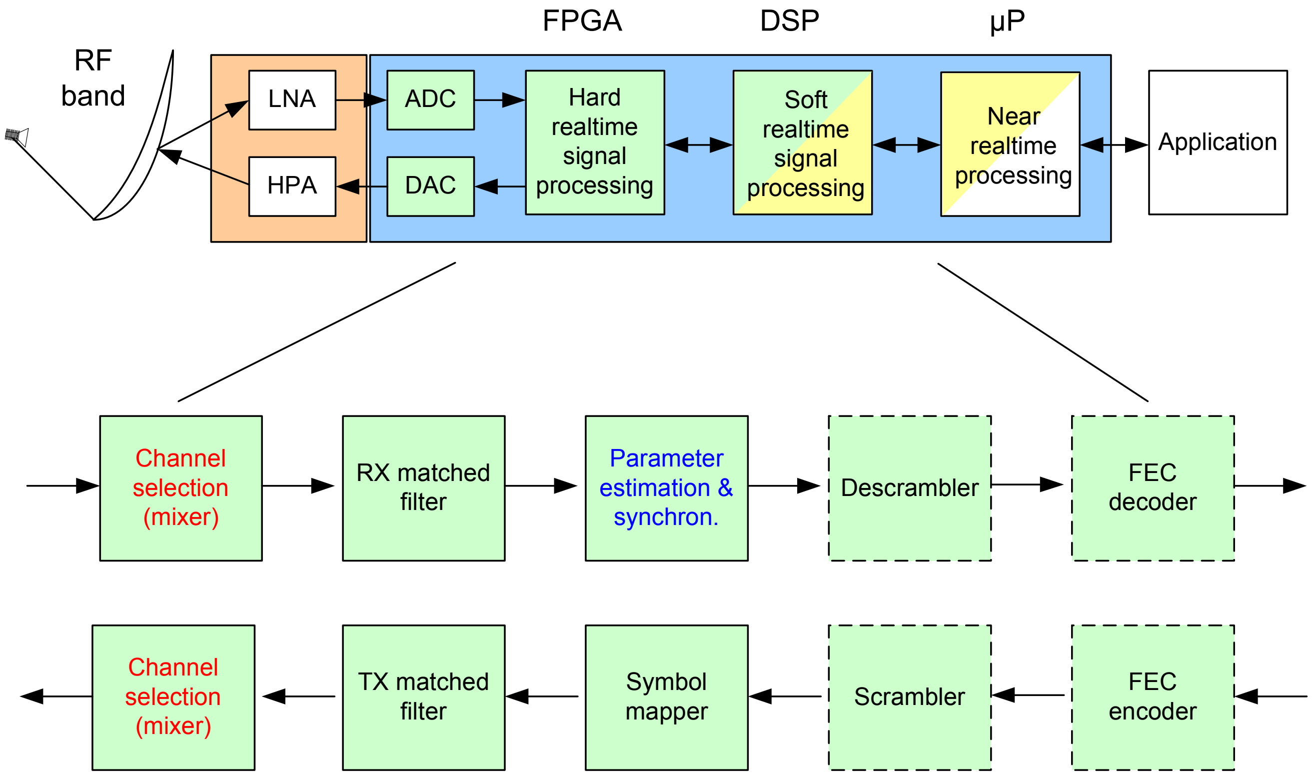
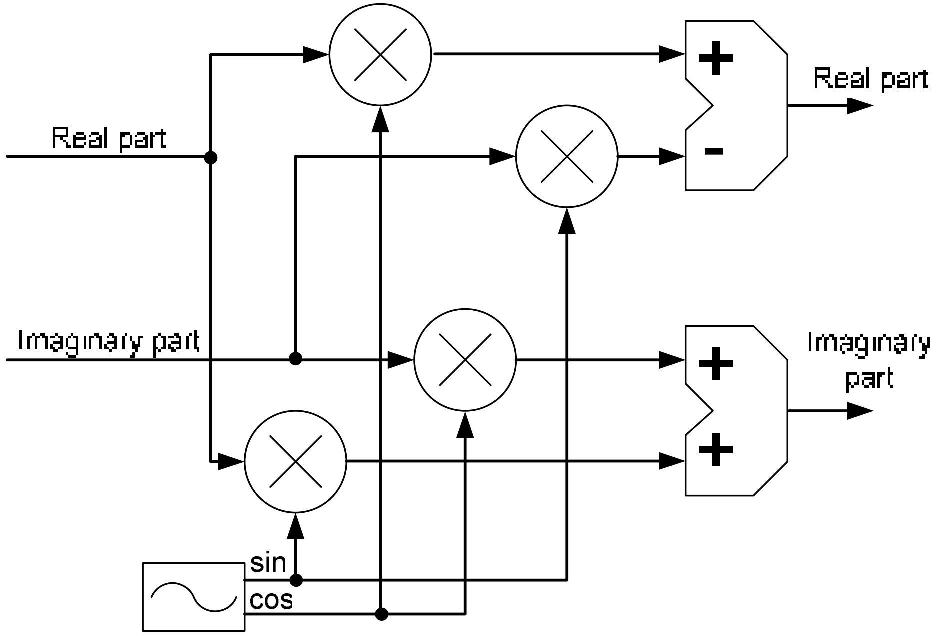


Mixer stages

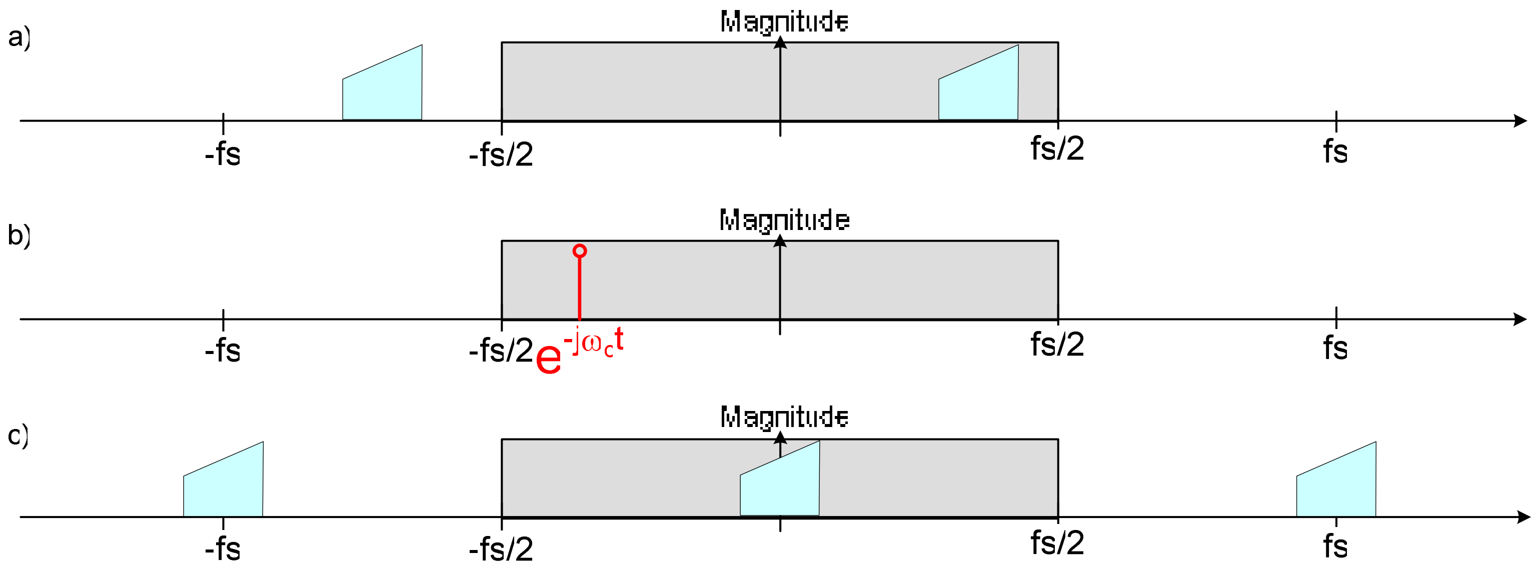
- Purpose
  - Shifting of frequency bands
  - Theoretical background: convolution theorem
- FPGA solutions
  - State-of-the art in SDR (IF-band processing)
  - Multipliers and memories available in silicon
- Components
  - Multipliers and adders
  - Numerical controlled oscillator (NCO)

Complex mixer

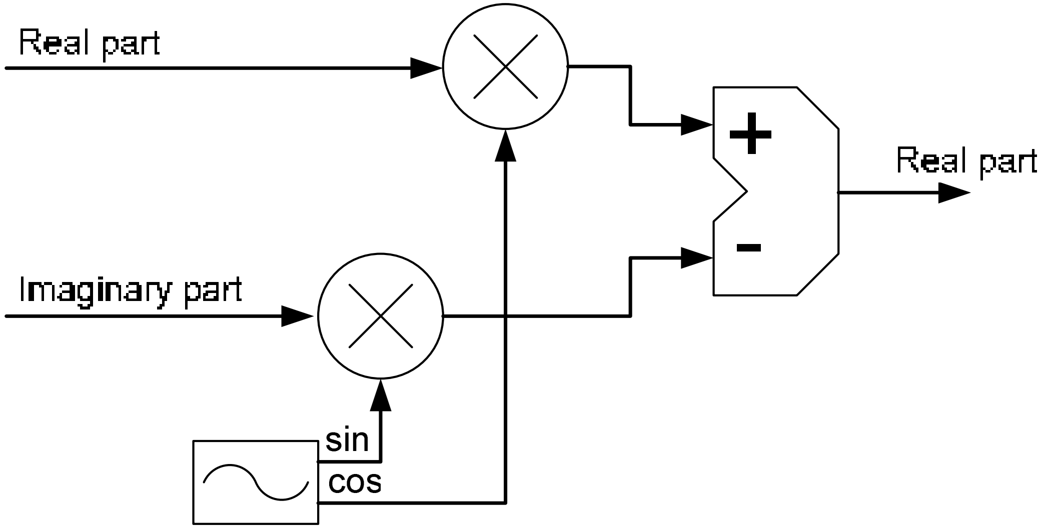


$$
\begin{aligned}
(\mathrm{Re}[z] + j\,\mathrm{Im}[z]) \cdot e^{j\omega_c t} &= (\mathrm{Re}[z] + j\,\mathrm{Im}[z]) \cdot (\cos\omega_c t + j\sin\omega_c t) \\
&= (\mathrm{Re}[z] \cdot \cos\omega_c t - \mathrm{Im}[z] \cdot \sin\omega_c t) + j(\mathrm{Re}[z] \cdot sin\omega_c t + \mathrm{Im}[z] \cdot \cos\omega_c t)
\end{aligned}
$$

Spectrum

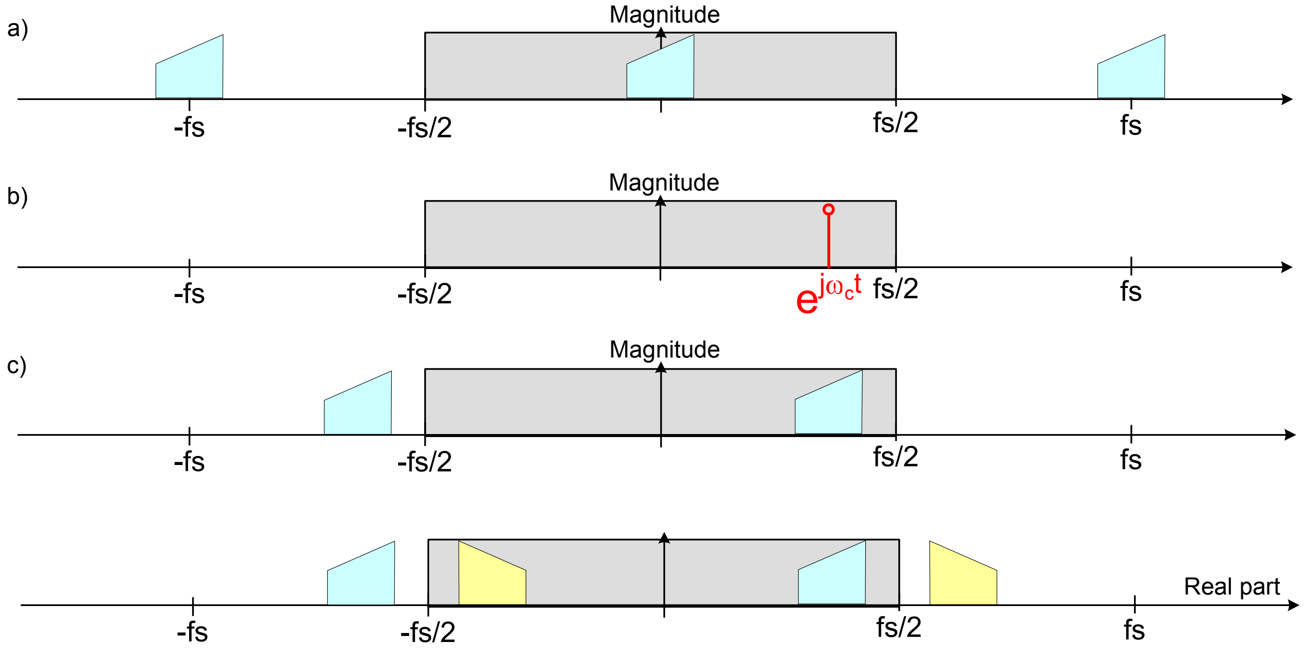

a)
Magnitude
-fs
-fs/2
fs/2
fs
b)
Magnitude
-fs
-fs/2
$e^{-j\omega_c t}$
fs/2
fs
c)
Magnitude
-fs
-fs/2
fs/2
fs

Real up-mixer

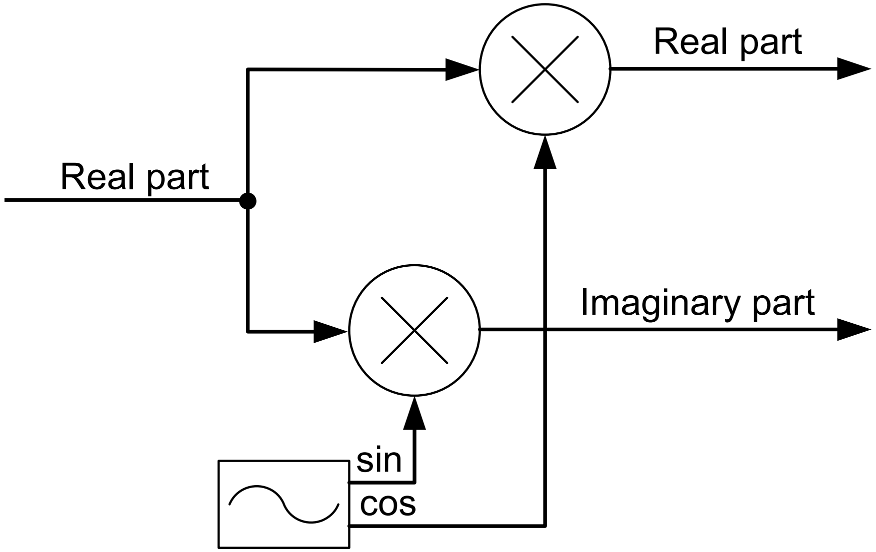


$$\mathrm{Re}\{(\mathrm{Re}[z] + j\,\mathrm{Im}[z]) \cdot e^{j\omega_c t}\} = \mathrm{Re}[z] \cdot \cos\omega_c t - \mathrm{Im}[z] \cdot \sin\omega_c t$$

Spectrum

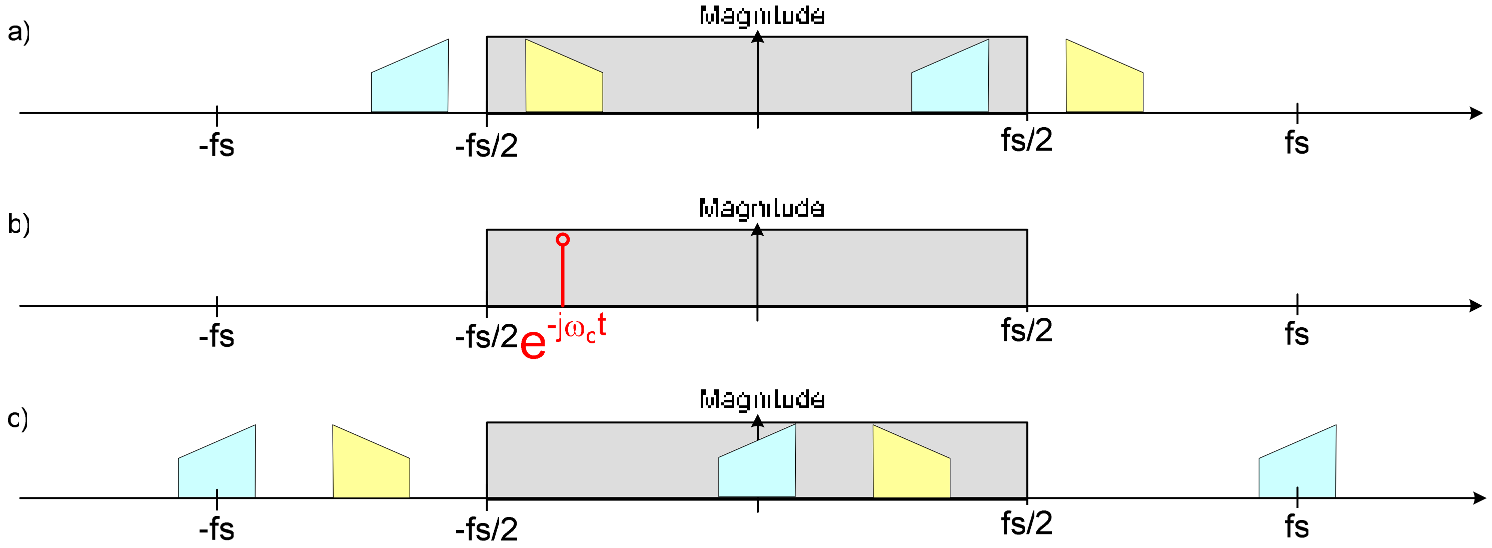

a)
Magnitude
-fs
-fs/2
fs/2
fs
b)
Magnitude
-fs
-fs/2
$e^{j\omega_c t}$
fs/2
fs
c)
Magnitude
-fs
-fs/2
fs/2
fs
Real part
-fs
-fs/2
fs/2
fs

Real down-mixer



$$\mathrm{Re}[z]\cdot e^{-j\omega_c t} = \mathrm{Re}[z]\cdot\cos\omega_c t - j\,\mathrm{Re}[z]\cdot sin\omega_c t$$

Spectrum


a)
Magnitude
-fs
-fs/2
fs/2
fs
b)
Magnitude
-fs
-fs/2
$e^{-j\omega_c t}$
fs/2
fs
c)
Magnitude
-fs
-fs/2
fs/2
fs

Numerical controlled oscillator

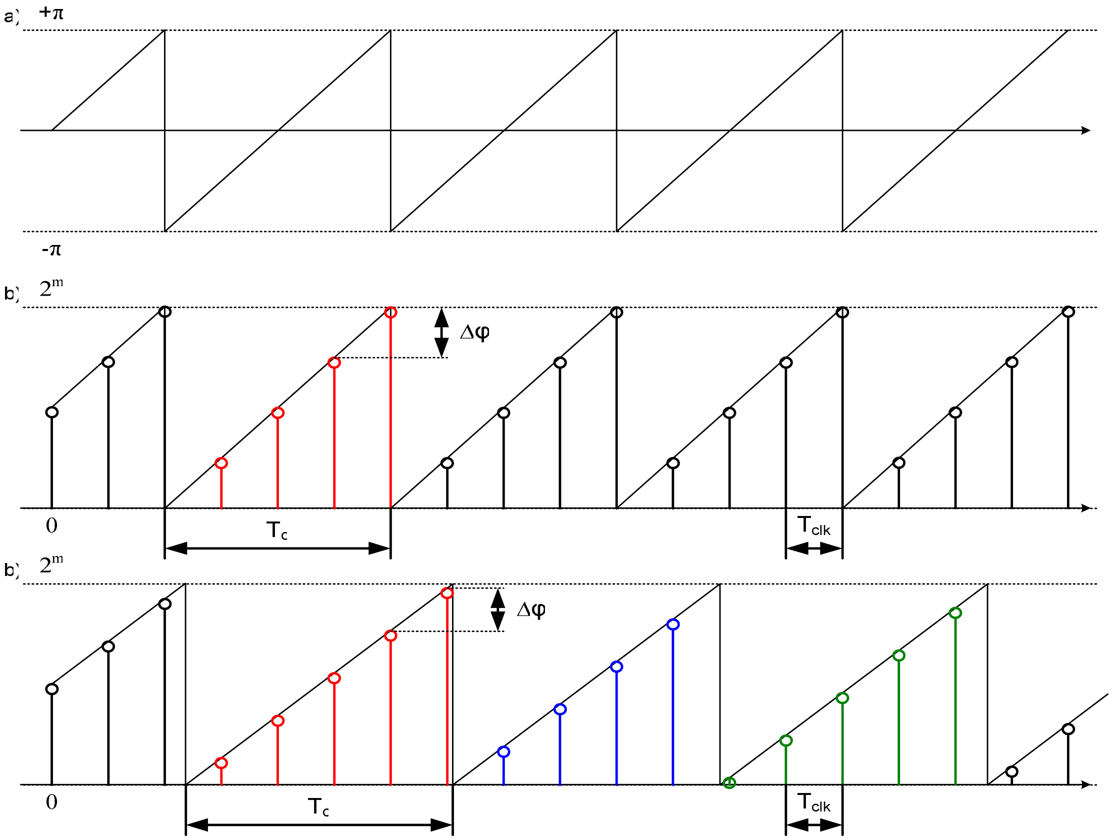
- Advantages over analog solutions
  - Accurate frequency, phase, and amplitude control
  - Extremely fine frequency resolution
  - No settling time after changes
  - No temperature drift
  - No aging

NCO basics

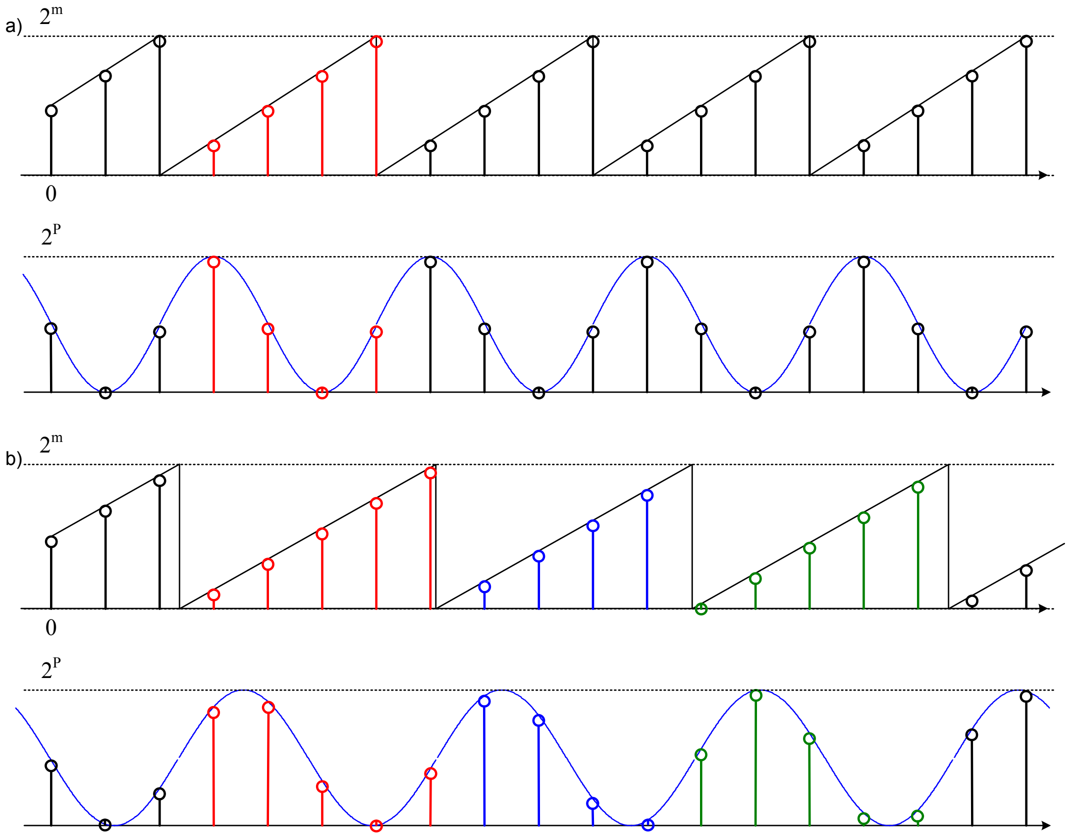
- Main components
  - Phase accumulator
  - Phase-to-amplitude converter
  - DA converter with anti-aliasing filter

$$s(t) = \sin(2\pi \cdot f_{\mathrm{c}} \cdot t)$$

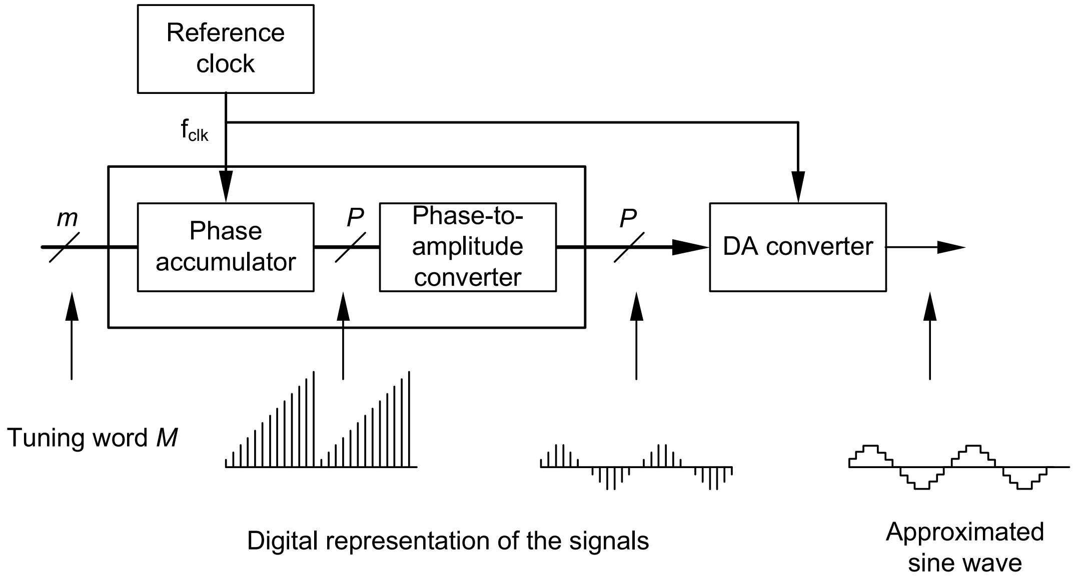
Phase accumulator



Phase-to-amplitude conversion



Digital synthesizer



$$s(t) = \sin(2\pi \cdot f_{\mathrm{NCO}} \cdot t): \quad f_{\mathrm{NCO}_{\min}} = \frac{f_{\mathrm{clk}}}{2^m}, \quad f_{\mathrm{NCO}} = \frac{f_{\mathrm{clk}}}{2^m} M$$

NCO imperfections

- Digital domain
  - Truncation of the phase accumulator word ($m \to p$)
  - Integer resolution of the sine values in the LUT
- Analog domain
  - DA conversion errors
    - Differential nonlinearity
    - Integral nonlinearity
    - Missing codes
  - Anti-aliasing filter
  - Phase noise of the clock

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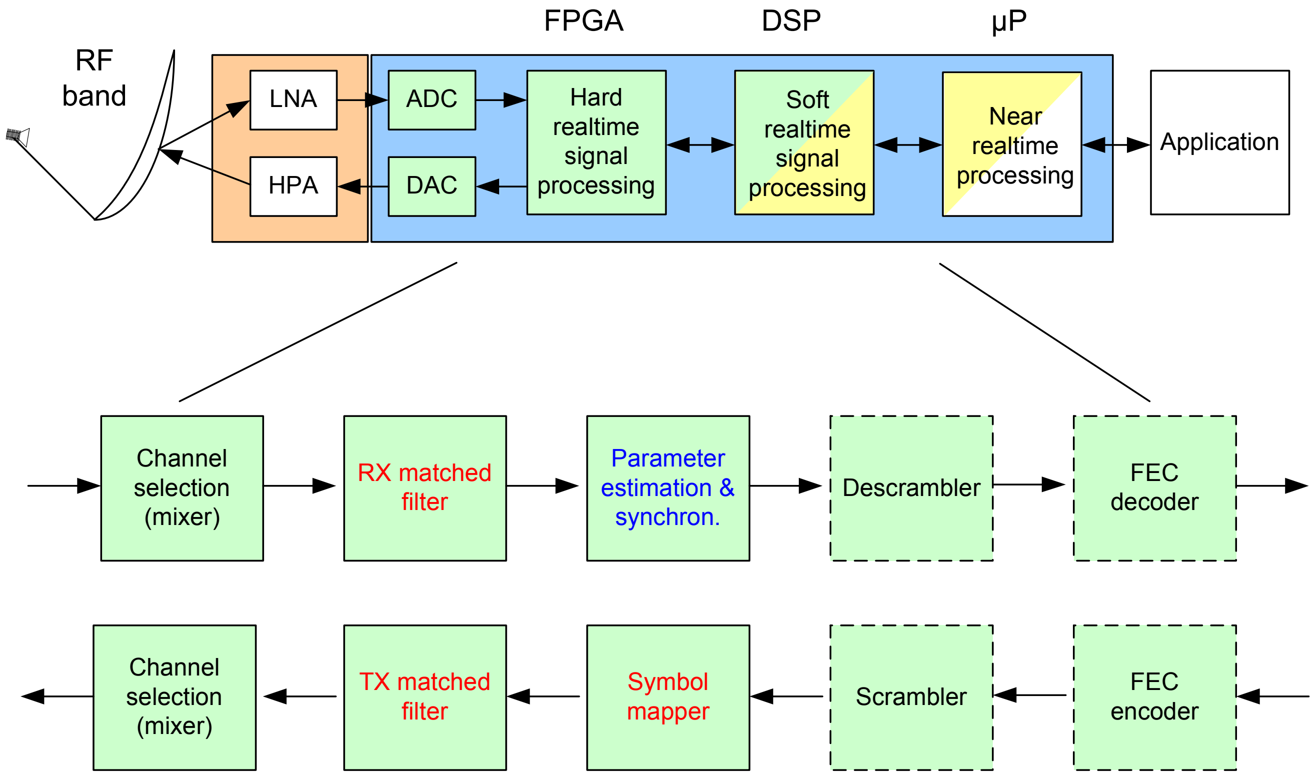
RX and TX function blocks



FPGA DSP μP

RF band

LNA

HPA

ADC

DAC

Hard realtime signal processing

Soft realtime signal processing

Near realtime processing

Application

Channel selection (mixer) → RX matched filter → Parameter estimation & synchron. → Descrambler → FEC decoder

Channel selection (mixer) ← TX matched filter ← Symbol mapper ← Scrambler ← FEC encoder

Digital modulation

- Modulation parameters
  - Amplitude-shift keying (ASK)
  - Frequency-shift keying (FSK)
  - Phase-shift keying (PSK)
  - Mixed forms (QAM, APSK)

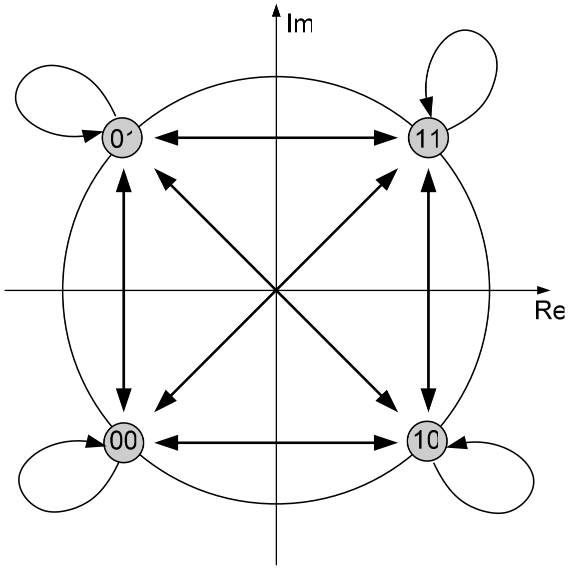
$$z(t) = A(t) \cdot \cos[2\pi f_c t + \varphi(t)] = A(t) \cdot \mathrm{Re}[e^{j(2\pi f_c t + \varphi(t))}]$$

$$s(t) = \underbrace{e^{j2\pi f_c t}}_{\text{Carrier signal}} \cdot \underbrace{A(t) \cdot e^{j\varphi(t)}}_{\text{Baseband signal}}$$

Baseband signals

$$s(t) = A(t) \cdot e^{j\varphi(t)} = A(t) \cdot \cos\varphi(t) + j \cdot A(t) \cdot \sin\varphi(t)$$

$$\text{QPSK}: \; s(t) = \begin{cases} +1+j & A(t)=1 & \varphi(t)=+\pi/4 \\ +1-j & A(t)=1 & \varphi(t)=-\pi/4 \\ -1+j & A(t)=1 & \varphi(t)=+3\pi/4 \\ -1-j & A(t)=1 & \varphi(t)=-3\pi/4 \end{cases}$$



QPSK scatter plot (Gray mapping)

Digital demodulation

- Timing errors
- Phase errors
- Frequency errors

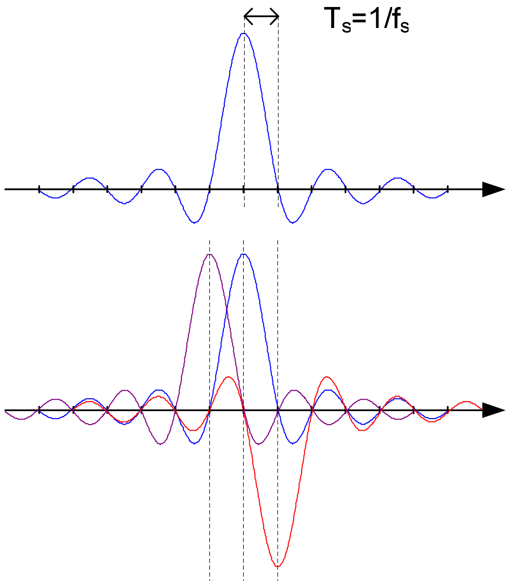
$$\mathrm{TX}: s(t) = e^{j2\pi f_c t} \cdot A \cdot e^{j\varphi(t)}$$

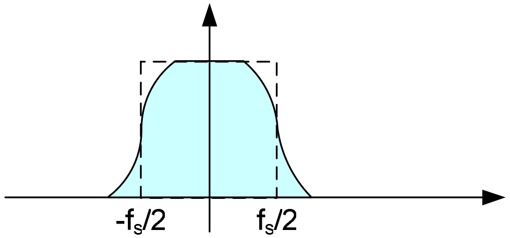
$$\mathrm{RX}: r(t) = e^{-j2\pi(f_c-\Delta f_c)t} \cdot s(t-t_0) + w(t)$$

$$= e^{j(2\pi\Delta f_c t+\varphi_0)} \cdot A \cdot e^{j\varphi(t-t_0)} + w(t)$$

Pulse shaping

- Nyquist theorems
  - Ideal brickwall filter does not introduce ISI
  - Shaping with odd symmetry does not introduce ISI





RCos function

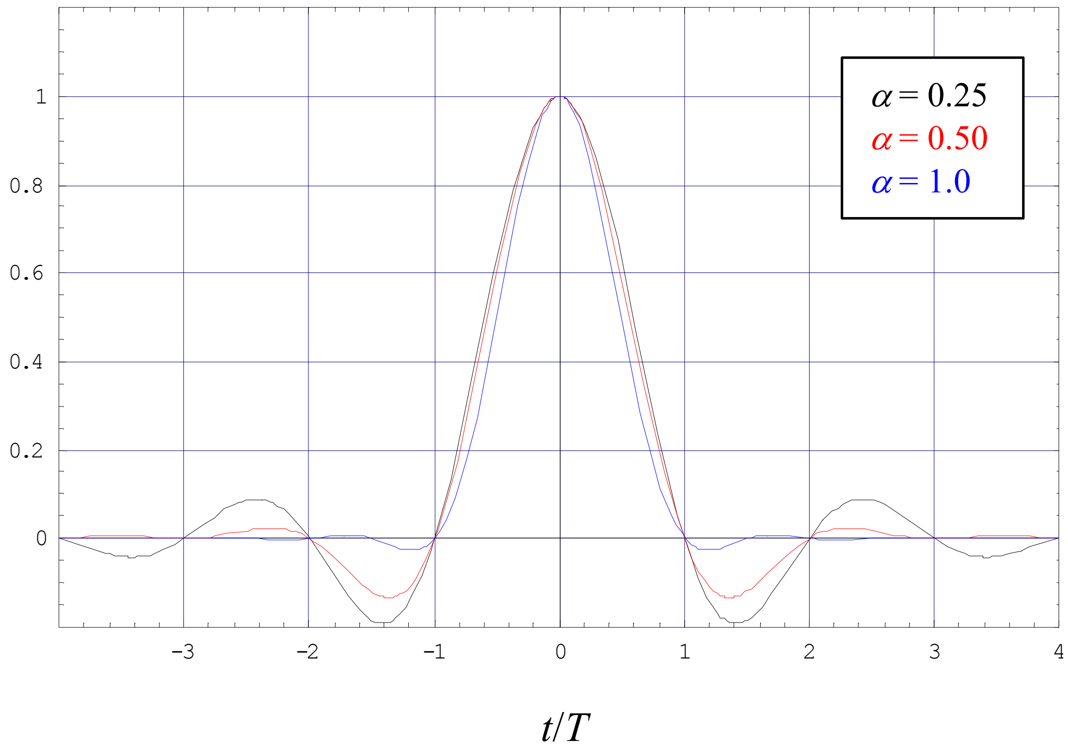
- Spectral shape

$$G(f) = \begin{cases} T_s, \; |fT_s| \le \tfrac{1}{2}(1-\alpha) \\ T_s \cos^2\left[\tfrac{\pi}{2\alpha}\left(|fT_s| - \tfrac{1-\alpha}{2}\right)\right], \; \tfrac{1}{2}(1-\alpha) < |fT_s| \le \tfrac{1}{2}(1+\alpha) \\ 0, \; |fT_s| > \tfrac{1}{2}(1+\alpha) \end{cases}$$

- Impulse response

$$g(t) = \int_{-\infty}^{\infty} G(f)\, e^{j2\pi ft}\, df = \operatorname{sinc}(t/T) \frac{\cos(\alpha\pi t/T)}{1-(2\alpha t/T)^2}$$

RCos impulse response



Matched filter

- Maximized signal-to-noise ratio (SNR) at the receiver
- Spectrum of the received signal is equal to the transfer function of the whole filter chain
- Impulse response of the RX matched filter is the complex conjugate of the mirrored impulse response of the channel: $h^*(-t)$
- Root-raised cosine (RRCos) filters are standard in satellite communications
  - Implementation as FIR structure
  - Parameter: roll-off factor α

RRCos filters

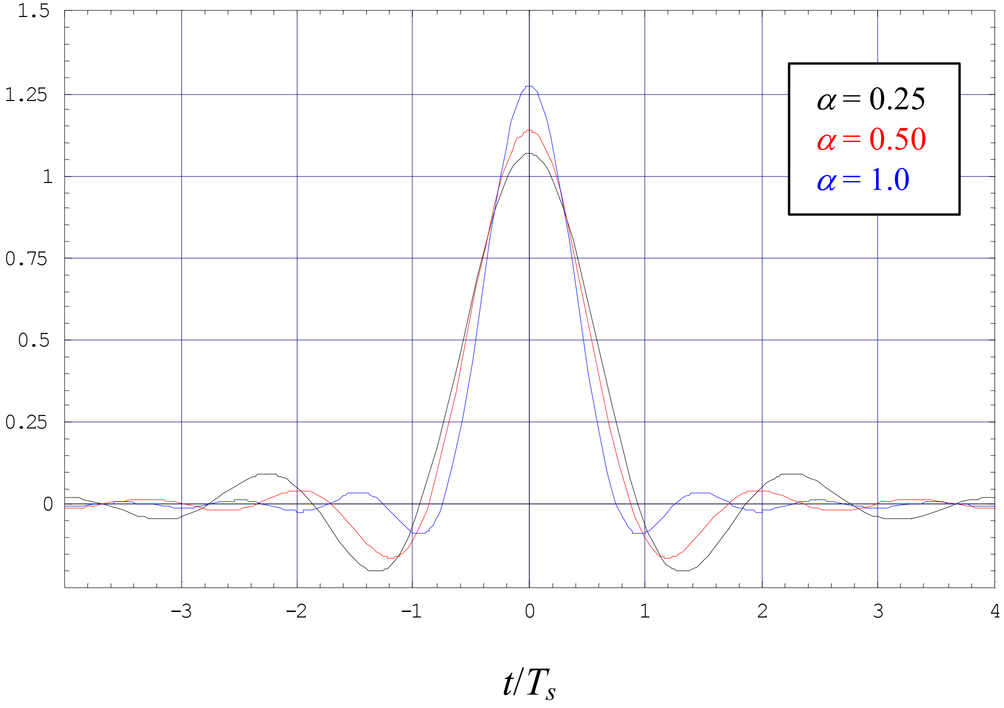
- Spectral shape

$$
\begin{aligned}
G(f) &= |H(f)|^2 \\
&= \begin{cases} T_s, \ |fT_s| \le \frac{1}{2}(1-\alpha) \\ T_s \cos^2\left[\frac{\pi}{2\alpha}(|fT_s| - \frac{1-\alpha}{2})\right], \ \frac{1}{2}(1-\alpha) < |fT_s| \le \frac{1}{2}(1+\alpha) \\ 0, \ |fT_s| > \frac{1}{2}(1+\alpha) \end{cases}
\end{aligned}
$$

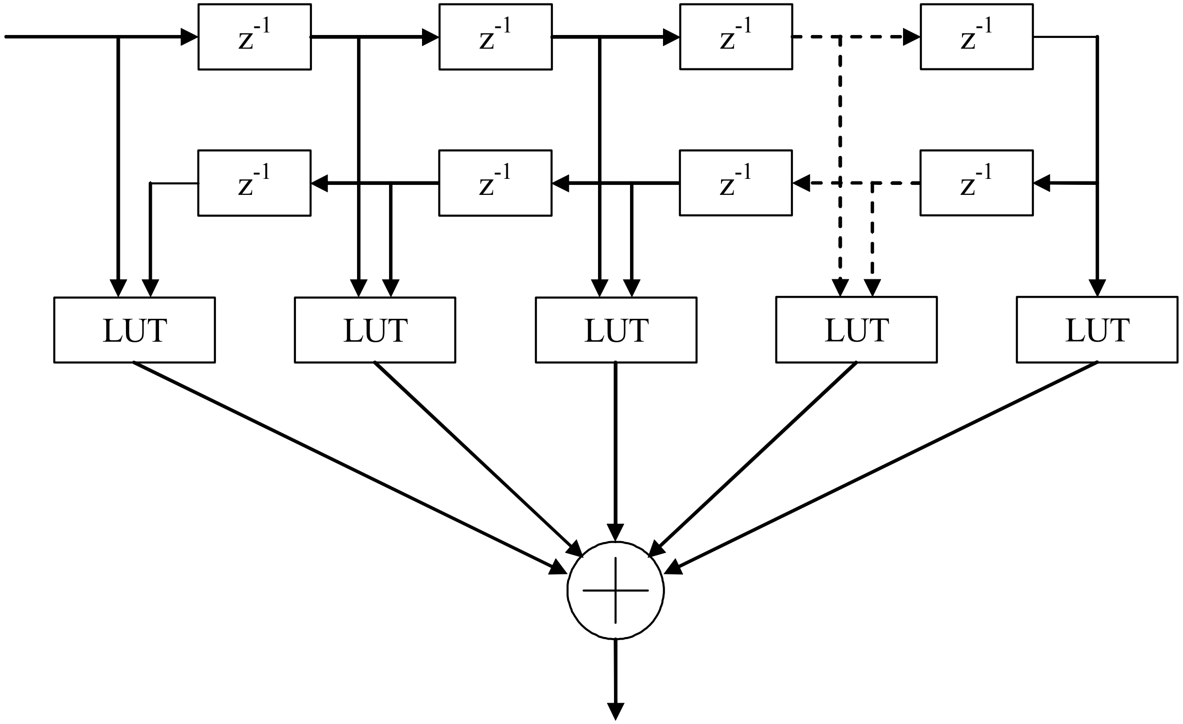
- Impulse response

$$
\begin{aligned}
h(t) &= h^*(-t) \\
&= \int_{-\infty}^{\infty} H(f) e^{j2\pi ft} df \\
&= \frac{1}{\sqrt{T_s}\pi} \frac{4\alpha \cos[\pi(1+\alpha)t/T_s] + \pi(1-\alpha)\mathrm{sinc}[(1-\alpha)t/T_s]}{1-(4\alpha t/T_s)^2}
\end{aligned}
$$

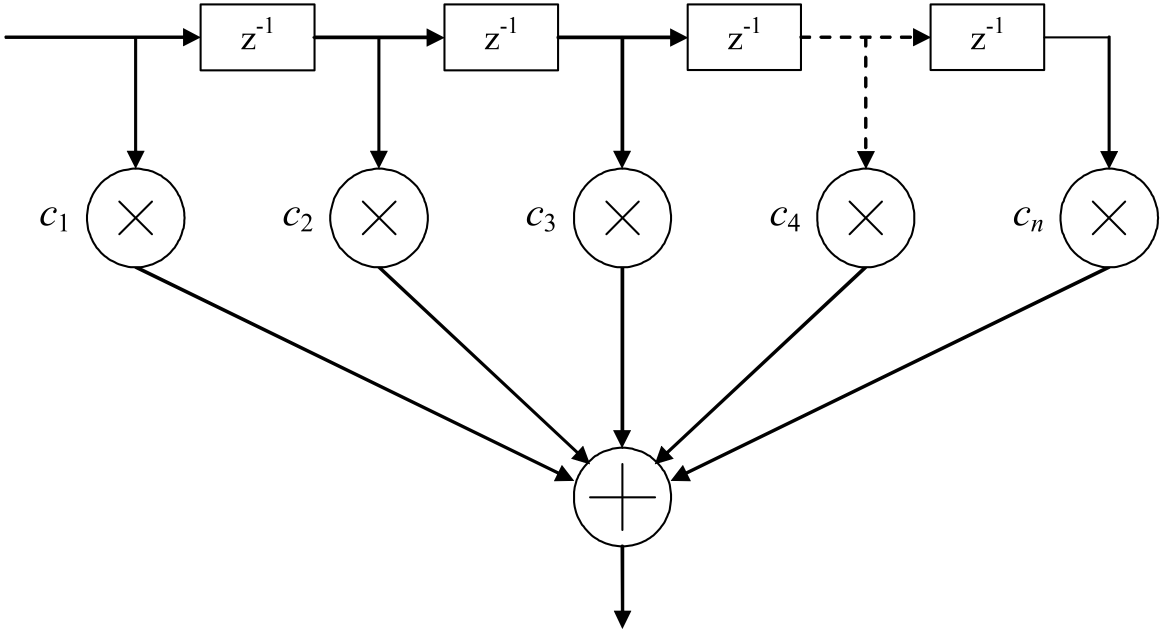
RRCos impulse response



TX implementation of RRCos



RX implementation of RRCos



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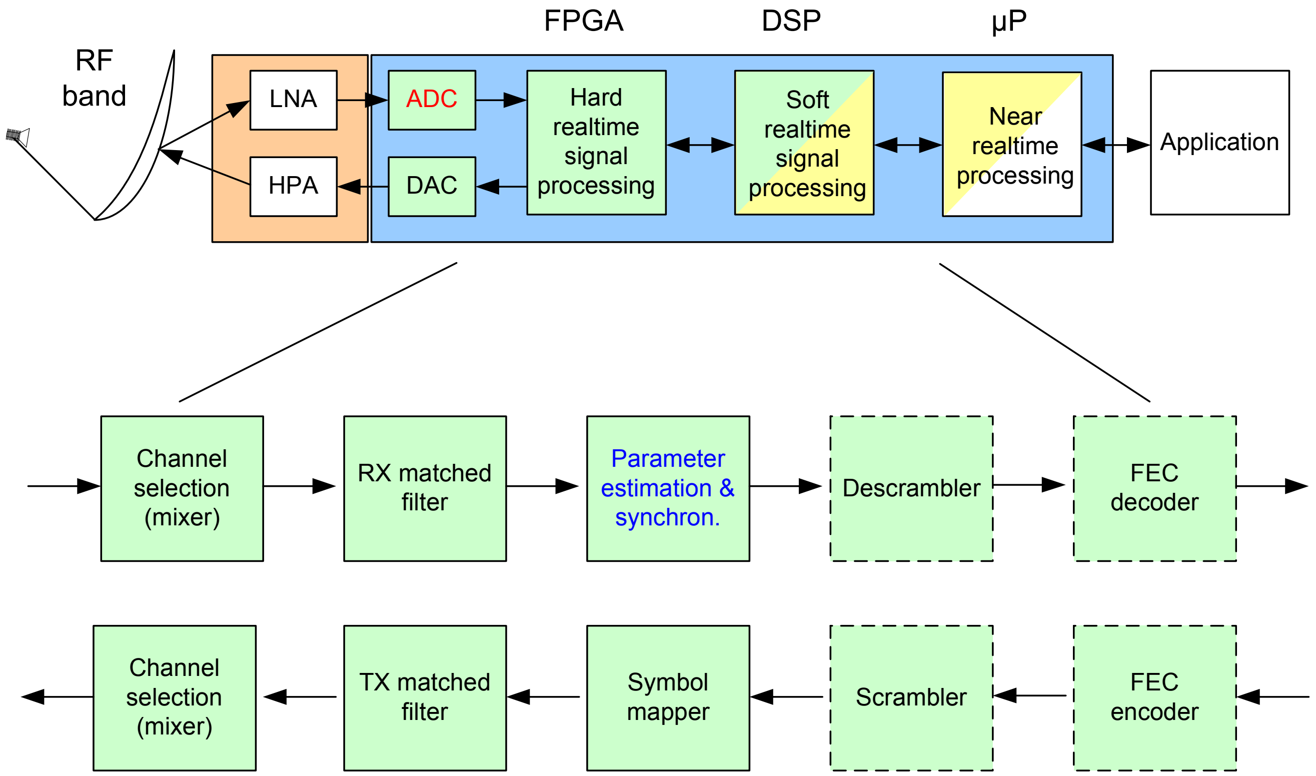


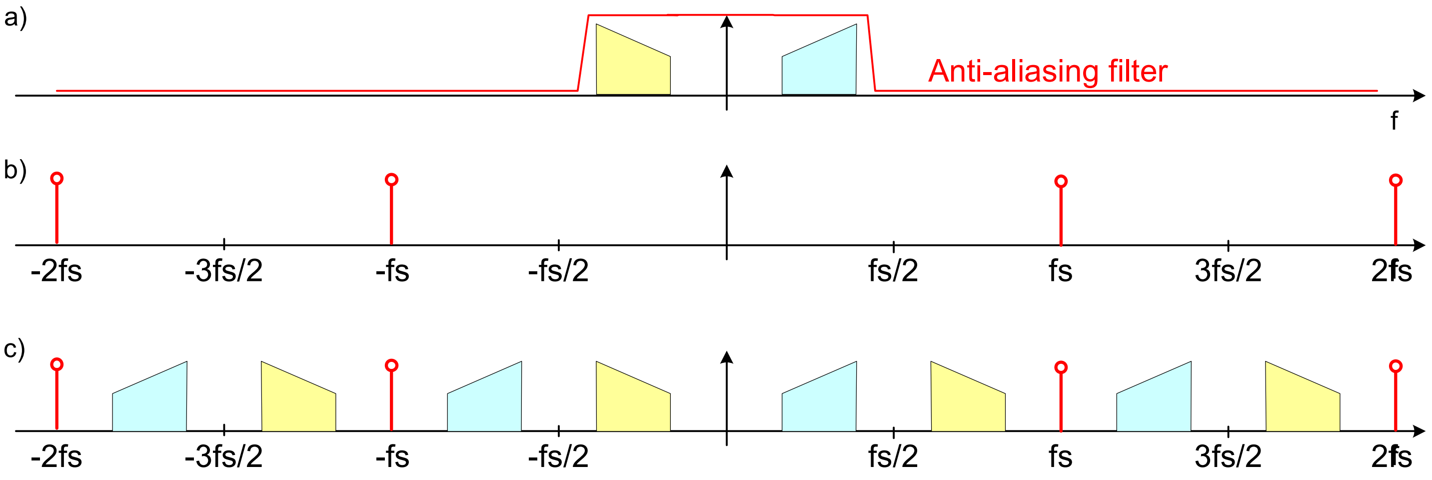
AD conversion

- Sampling
  - Time-continuous signal converted into a time-discrete signal
  - Nyquist criterion must be satisfied so as to avoid interference
- Quantization
  - Amplitude-continuous signal converted into a series of digital numbers
  - Quantization noise due to finite resolution

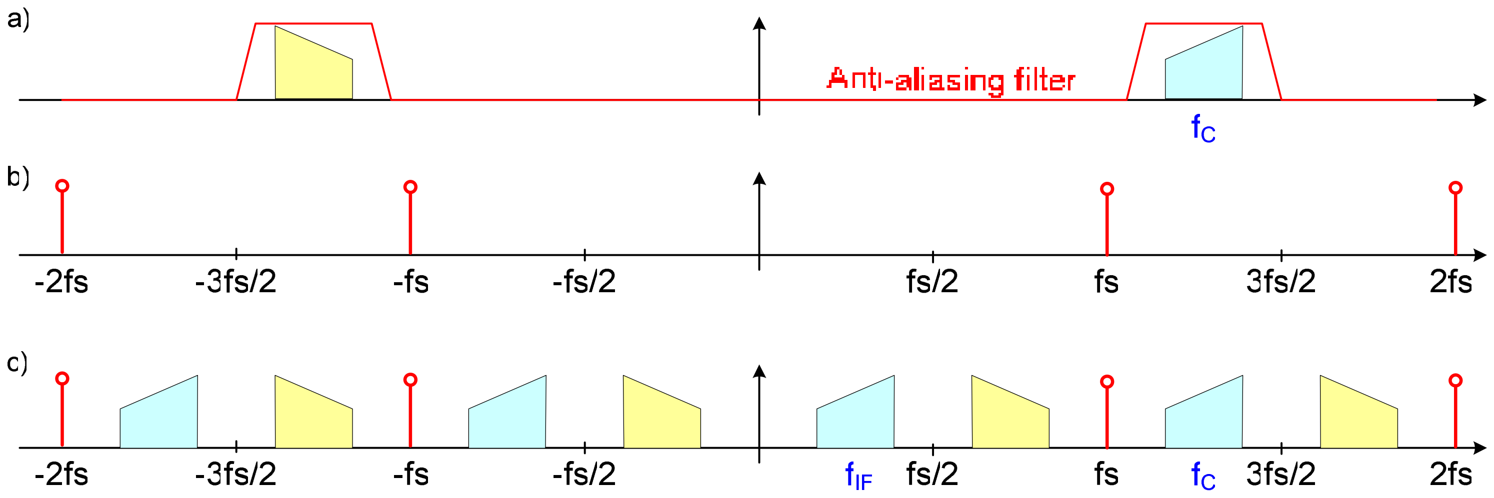
Sampling principles

- Super-Nyquist sampling
  - Sampling frequency must be at least twice the highest frequency component of the analog signal
- Sub-Nyquist sampling
  - Sampling frequency of the signal must be at least twice the bandwidth of the analog signal: $f_{IF} = f_C - n \cdot f_s$

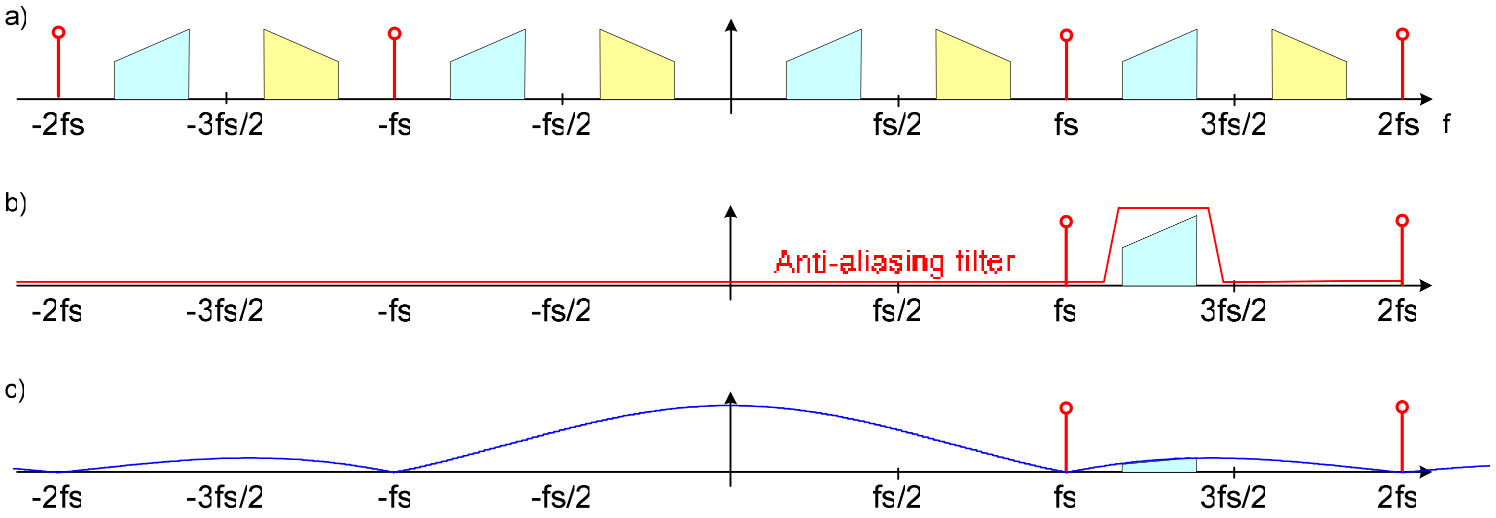
Super-Nyquist sampling



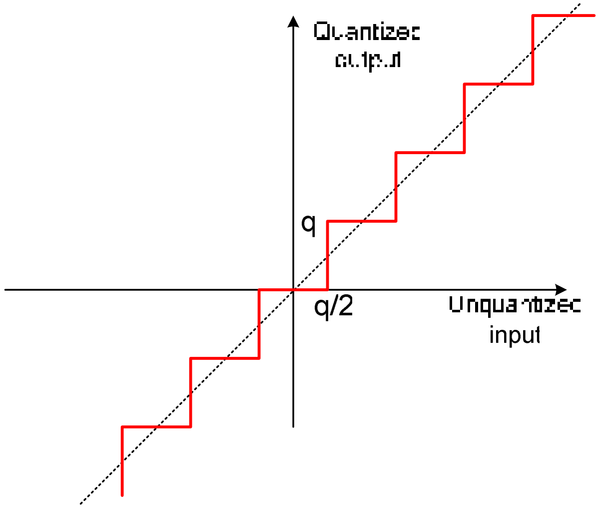
Sub-Nyquist sampling

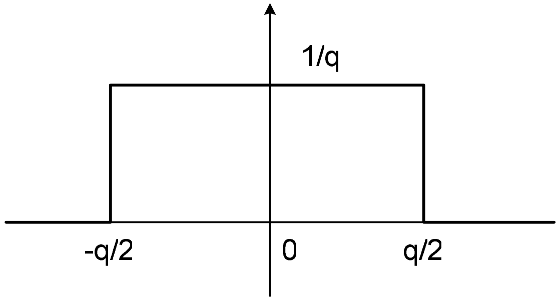


Sampling used for up-conversion



Effect of quantization



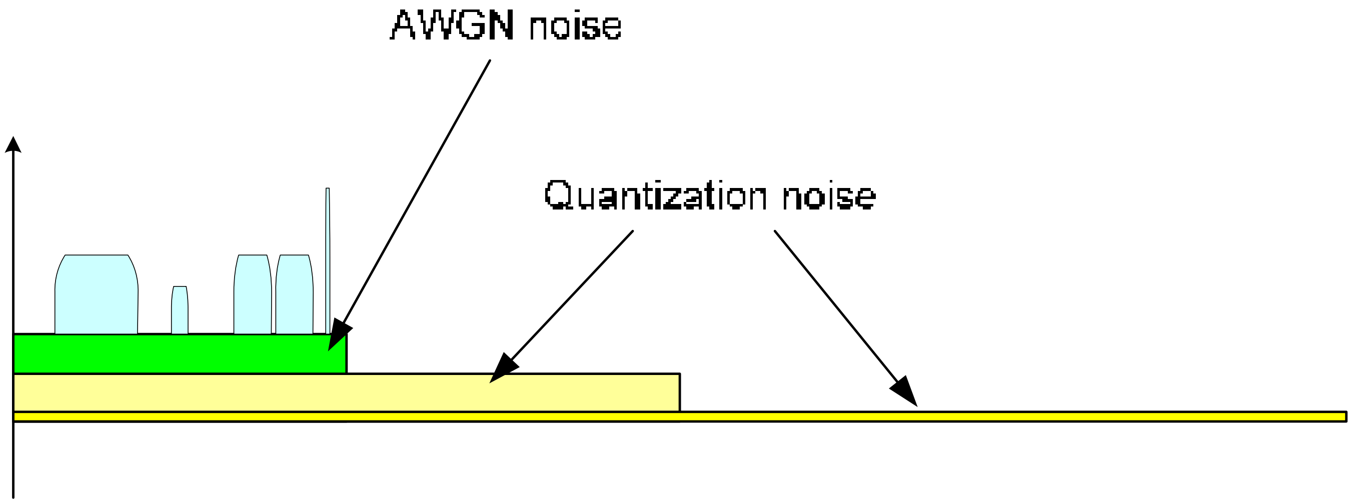


$$P_Q = \sigma_E^2 = \int_{-\infty}^{+\infty} e^2 f_E(e)\, de = \frac{1}{q}\int_{-q/2}^{+q/2} e^2\, de = \frac{q^2}{12}$$

Quantization noise

- The quantization noise is white if:
    - Subsequent quantization errors are uncorrelated
    - The quantization sequence is not correlated to the input signal
    - Quantization error is uniformly distributed over ±q/2
- This is achieved with signals:
    - Having large amplitudes compared to the quantization interval
    - Having a noisy-like amplitude distribution
- Conditions usually satisfied by multi-carrier and modulated signals
- Conditions violated by clean carrier signals (pure sinusoidal shape)

Spectral distribution



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