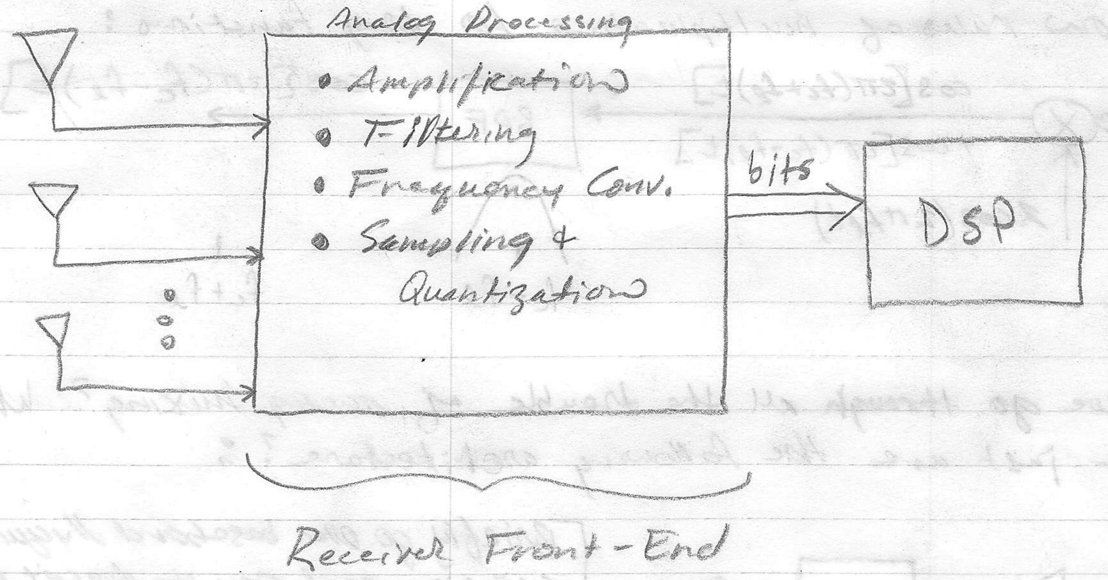
Section 11.1 in M+E

Lecture 8: Signal Conditioning Overview

In modern radios, we move analog signals into the digital domain as soon as possible. Gone are the days of chaining op-amps together to do matched filtering and data detection. Digital signal processing ~~(immune to temperature effects & variations in discrete components (capacitors, resistors))~~ has many advantages over analog signal processing:

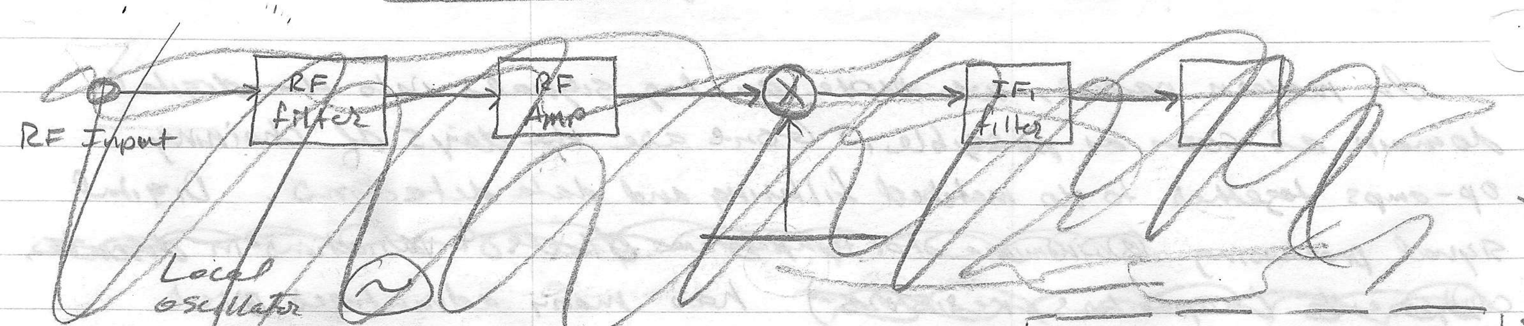
1. Flexibility: recompile vs. re-do layout and re-wire
2. Accuracy: DSP is immune to temp. and component variations
3. Storage: Signals can be stored for later analysis or synthesis
4. Cheaper

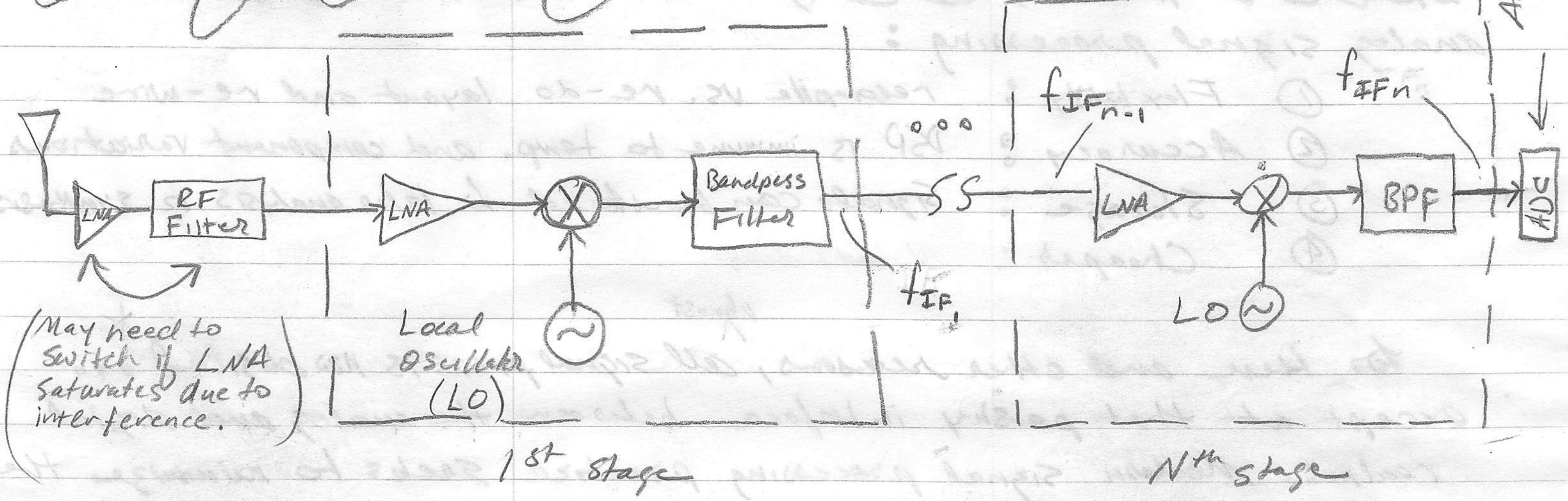
For these and other reasons, almost all signal proc. is now digital*... except at that pesky interface between the analog and digital realms. Modern signal processing practice seeks to minimize the analog signal processing, but it can't be eliminated entirely. After all, the received signals are analog.



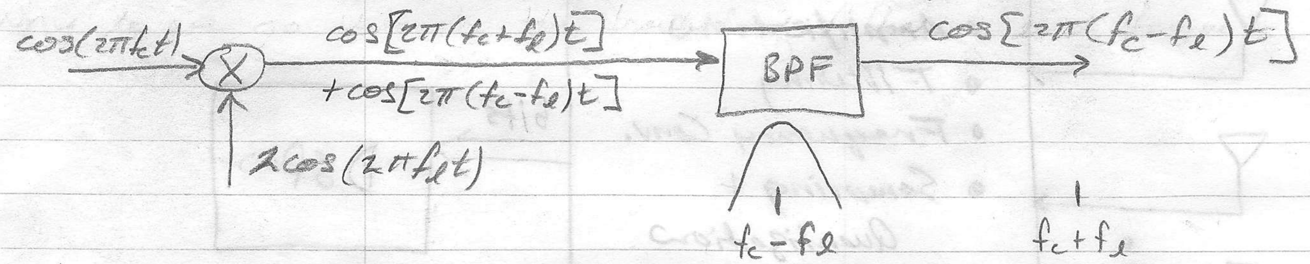
* Note: there is a resurgence of analog for extremely power sensitive apps.

Traditional Superheterodyne RF Front-End:

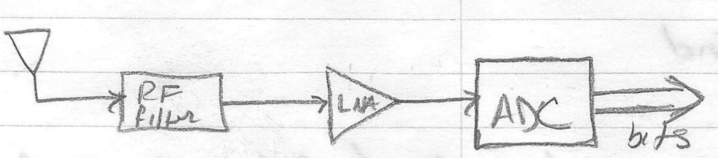




This is a widely used approach to RX (and TX) design. It is based on rules of multiplication for trig functions:



Q: Why do we go through all the trouble of analog mixing? Why don't we just use the following architecture?:



[Briefly go over baseband Nyquist criterion and say it doesn't apply – it's the bandpass sampling thm. that applies, and f_{samp} can be low. More later...]

A: This approach may be the way of the future, as ADCs get faster. Meanwhile, it's still somewhat large, expensive, & power hungry because

{amplification, transmission, filtering, sampling} all easier (cheaper and more power efficient) at intermediate frequencies

Consider for example the requirements of the bandpass RF filter. The quality (Q) factor is given by:

$$Q = \frac{f_c}{B_{3dB}}$$

f_c — center frequency

B_{3dB} — 3-dB bandwidth

Suppose we only wished to select the GPS L_1 C/A band, then

$$Q \cong \frac{1575.42}{2} = 788$$

This is extremely high — hard to find Q values larger than 500 and even then the filters are expensive.

Two options: ① widen bandwidth and sample faster — (ARL DFE)

② mix down to some IF before filtering and sampling.

Also consider amplification requirements. In the analog processing, we will need to amplify the incoming signal by 50 → 100 dB (Wow!) Easiest to spread this out over several IF stages to avoid leakage & self-jamming.

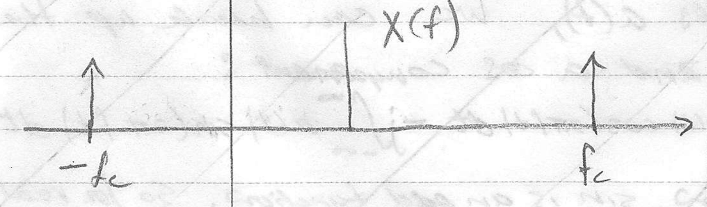
Finally, consider ADC cost: lower sampling rates are cheaper.

[Show slides of ARL DFE; maybe this is the future?]

<u>Frequency Conversion</u> in the <u>Fourier Domain</u>

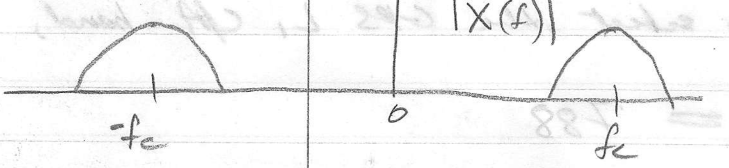
$$X(f) = \mathcal{F}[\cos(2\pi f_c t)] = \frac{1}{2}\mathcal{F}[e^{j2\pi f_c t}] + \frac{1}{2}\mathcal{F}[e^{-j2\pi f_c t}]$$

$$= \frac{1}{2}\left[\delta(f-f_c) + \delta(f+f_c)\right]$$



Now with modulation: $X(f) = \mathcal{F}\left[a(t)\cos(2\pi f_c t)\right] = \frac{1}{2}\underbrace{\mathcal{F}[a(t)]}_{A(f)} * \left[\delta(f-f_c)+\delta(f+f_c)\right]$

$$X(f) = \frac{1}{2}\left[A(f-f_c) + A(f+f_c)\right]$$

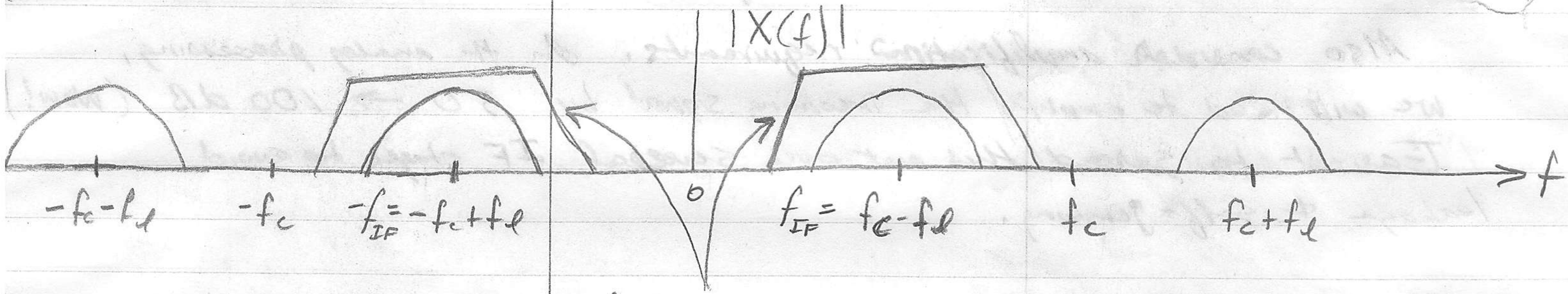


Now consider frequency conversion:

$$\mathcal{F}\left[4a(t)\cos(2\pi f_c t)\cos(2\pi f_\ell t)\right] = A(f) * \left[\delta(f-f_c)+\delta(f+f_c)\right] * \left[\delta(f-f_\ell)+\delta(f+f_\ell)\right]$$

(assume $f_\ell < f_c$)

$$X(f) = A(f-f_\ell-f_c) + A(f+f_\ell-f_c) + A(f-f_\ell+f_c) + A(f+f_\ell+f_c)$$



We use a bandpass filter to isolate the signal component that we want:
$f_{IF} = f_c - f_\ell$ ("intermediate frequency")

In the diagram above, we have assumed that $f_\ell < f_c$. This corresponds to "low-side mixing." [They'll explore high-side mixing in HW.]

HW

Skip (HW):

Q: What changes if $f_\ell > f_c$ ("high-side mixing")?
[See discussion on p. 4]

Irrelevant

A: For real signals $a(t)$, we can break up the Fourier transform into a sin and a cos component:

$$A(f) = \mathcal{F}[a(t)] = \int_{-\infty}^{\infty} a(t)\cos(2\pi f t)\,dt - j\int_{-\infty}^{\infty} a(t)\sin(2\pi f t)\,dt$$

cos is an even function, sin is an odd function. So for real $a(t)$, the real part of $A(f)$ is an even function, and the imag. part is an odd function. $A(f) = A_e(f) + jA_o(f)$

~~Note: Re-do this with a bandpass representation: $x_c(t)\cos\omega_c - x_s(t)\sin(\omega_c)$~~
Irrelevant

It does flip in that case.

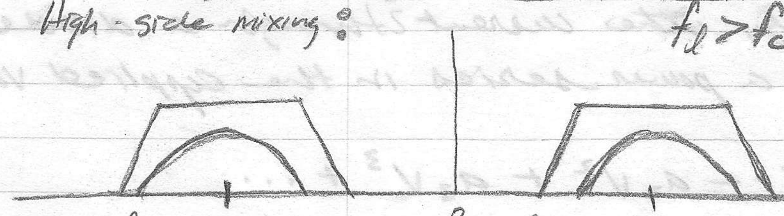
$A_e(f) = A_e(-f)$, $-A_o(f) = A_o(-f)$

Thus $A(-f) = A_e(f) - j A_o(f) = A^*(f)$ (Hermitian Fcn.)

So if a frequency-converted $A(f)$ crosses completely to the other side of the zero line, its phase gets reversed:

HW

High-side mixing: $f_\ell > f_c$



$a(t)\cos[2\pi(-f_c + f_\ell)t]$

[See discussion p. 4]

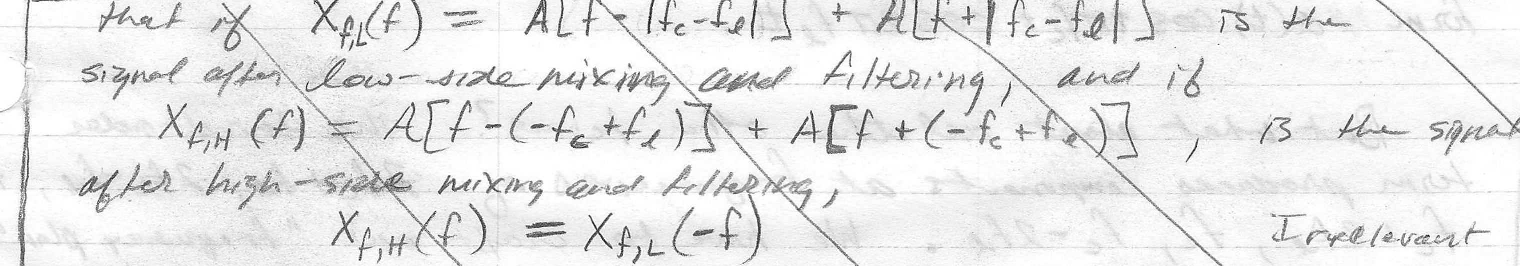
Note that with high-side mixing, a small increment in f_c actually decreases the intermediate frequency $f_{IF} = -f_c + f_\ell$. Also note that if $X_{f,L}(f) = A[f - |f_c - f_\ell|] + A[f + |f_c - f_\ell|]$ is the signal after low-side mixing and filtering, and if
$X_{f,H}(f) = A[f - (-f_c + f_\ell)] + A[f + (-f_c + f_\ell)]$, is the signal after high-side mixing and filtering,

$$X_{f,H}(f) = X_{f,L}(-f)$$

Irrelevant

I think the only relevant point here is that the carrier phase gets reversed. Thus the phase $\Phi_{IF}(t) = 2\pi f_{IF} t$ has a component that decreases with increasing carrier phase. If $\phi_c(t) = 2\pi f_c t$, then $\Phi_{IF}(t) = -\phi_c(t) + 2\pi f_\ell t$. That's a carrier phase reversal.

Image Frequencies



Watch how signal @ $f_\ell - f_{IF}$ gets mapped into vicinity of f_{IF}:

$$\cos[2\pi(f_\ell - f_{IF})t]\cos 2\pi f_\ell t = \tfrac{1}{2}\cos 2\pi f_{IF} t + \tfrac{1}{2}\cos[2\pi(2f_\ell - f_{IF})t]$$

The image reject filter must be narrow enough to exclude signals at the image frequency $f_{image} = f_\ell - f_{IF}$. Thus $B_{IR} < 2f_{IF}$. This is usually easily achievable in practice with a 2-3 stage downconversion).

Q: In practice, how do we make the multipliers required for mixing?

A: We use some circuit element with a nonlinear voltage-current response. For example, the current i through a diode can be expressed as a power series in the applied voltage V:

$$i = a_0 + a_1 V + a_2 V^2 + a_3 V^3 + \ldots$$

Suppose we make $V = a(t)\cos(2\pi f_c t) + \cos(2\pi f_\ell t)$

The second-order term in V then gives rise to a product of the form $a(t)\cos 2\pi f_c t \cos 2\pi f_\ell t$.

But what about all the other terms? The third-order term produces components at frequencies of $2f_c + f_\ell$, $2f_c - f_\ell$, f_ℓ, $f_c + 2f_\ell$, f_c, $f_c - 2f_\ell$. We have to craft our "frequency plan" to unwanted combinations from falling within the passband. With multiple mixing stages, this can be a real juggling act.

[Take a look at the GP2015 front-end. Trace through its frequency plan.]

Note that a direct-to-baseband downconversion (also known as "zero IF") is becoming ever more popular.

Leave out or do in homework. The zero IF approach in theory does not require image rejection because it selects only the desired band. But there are challenges with DC bias.

Our Ettus boxes use zero IF.

HW: Consequence of High-Side Mixing

In high-side mixing, $f_\ell > f_c$. As a consequence, the result of the mixing operation is

$$2a(t)\cos(2\pi f_c t)\cos(2\pi f_\ell t) = a(t)\cos[2\pi(f_c - f_\ell)t] + \underbrace{HFT}_{\text{Filtered Out}}$$

But we don't treat or model the cos as having a negative frequency, so we invoke $\cos(a) = \cos(-a)$ to render the frequency positive:

$$a(t)\cos[2\pi(-f_c + f_\ell)t]$$

f_c and f_ℓ will have nominal values and offsets:

$$f_c = f_{c_{nom}} + f_{cD} \quad , \quad f_\ell = f_{\ell nom} + f_{\ell D}$$

f_{cD}: Doppler due to relative motion + TX clock

$f_{\ell D}$: Doppler due to RX clock

We model the received and downmixed signal as

$$a(t)\cos[2\pi f_{IF} t + \theta(t)]$$

where
$$f_{IF} = \begin{cases} f_{c nom} - f_{\ell nom} & (\text{HS mixing}) \\ f_{\ell nom} - f_{c nom} & (\text{LS mixing}) \end{cases}$$

Thus, we have that

$$\theta(t) = \begin{cases} \theta_{HS}(t) = \int_0^t f_{\ell D}(\tau) - f_{cD}(\tau)\, d\tau + \theta(0) & (\text{HS mixing}) \\ \theta_{LS}(t) = \int_0^t f_{cD}(\tau) - f_{\ell D}(\tau)\, d\tau + \theta(0) & (\text{LS mixing}) \end{cases}$$

Note that $\dot{\theta}_{HS}(t) = -\dot{\theta}_{LS}(t)$