TSEK38: Radio Frequency Transceiver Design
Lecture 2: Fundamentals of System Design

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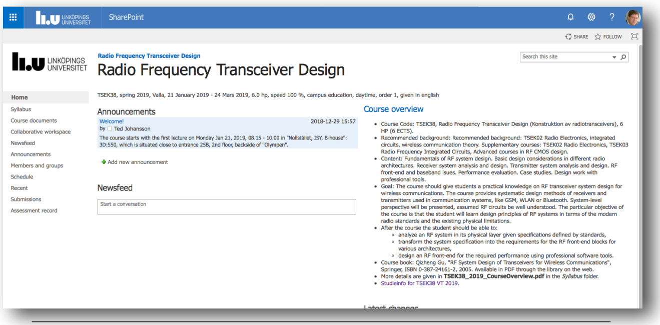
Repetion of Lecture 1

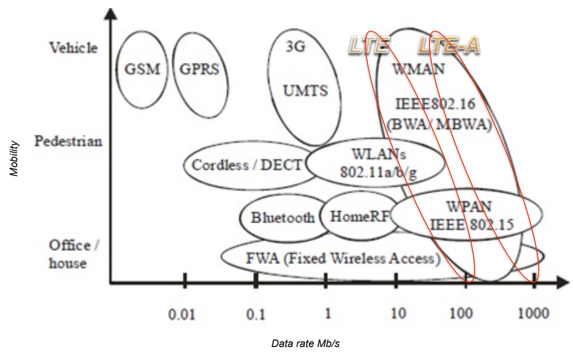
- Wireless communication systems today
- Some wireless standards
- Communication radio examples
- Architectures: the big picture

Lisam Course Room

- https://liuonline.sharepoint.com/sites/TSEK38/TSEK38_2019VT_FU



There is a standard for almost any need!



Examples of Standards

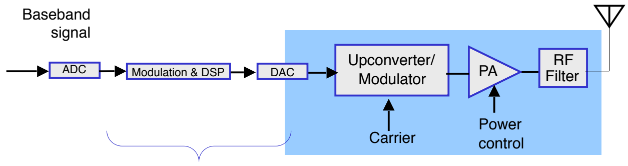
Standard	Access Scheme/Dupl	Frequency band (MHz)	Channel Spacing	Frequency Accuracy	Modulation Technique	Rate (kb/s)	Peak Power (uplink)
GSM	TDMA/FDMA/ TDD	890-915 (UL) 935-960 (DL)	200 kHz	90 Hz	GMSK	270.8	0.8, 2, 5, 8 W
DCS-1800	TDMA/FDMA/ TDD	1710-1785 (UL) 1805-1850 (DL)	200 kHz	90 Hz	GMSK	270.8	0.8, 2, 5, 8 W
DECT	TDMA/FDMA/ TDD	1880-1900	1728 kHz	50 Hz	GMSK	1152	250 mW
IS-95 cdmaOne	CDMA/ FDMA/ FDD	824-849 (RL) 869-894 (FL)	1250 kHz	N/A	OQPSK	1228	N/A
Bluetooth	FHSS/TDD	2400-2483	1 MHz	20 ppm	GFSK	1000	1,4,100 mW
802.11b (DSSS)	CDMA/TDD	2400-2483	20 MHz	25 ppm	QPSK/CCK	1, 2, 11 Mb/s	1 W
WCDMA (UMTS)	W-CDMA/TD-CDMA/F/TDD	1920-1980 (UL) 2110-2170 (DL)	5 MHz	0.1 ppm	QPSK, 16/64QAM	3840 (max)	0.125, 0.25, 0.5, 2W
LTE	OFDMA (DL) SC-FDMA (UL) FDD/TDD	700-2700	scalable to 20MHz	4.6 ppm /32 ppm 0.1 ppm (BS)	QPSK, 16/64QAM	DL/UL 300/75 Mb/s	Scalable with BW to 250 mW

5G

	4G (Today, Before Further Developments)	5G
Latency	10 ms	Less than 1 ms
Peak data rates	1 Gbps	20 Gbps
Number of mobile connections	8 billion (2016)	11 billion (2021)
Channel bandwidth	20MHz 200kHz (for Cat-NB1 IoT)	100MHz below 6GHz 400MHz above 6GHz
Frequency band	600MHz to 5.925 GHz	600MHz–mmWave (for example, 28GHz, 39GHz, and onward to 80 GHz
Uplink waveform	Single-carrier frequency division multiple access (SC-FDMA)	Option for cyclic prefix orthogonal frequency-division multiplexing (CP-OFDM)
User Equipment (UE) transmitted power	+23 decibel-milliwatts (dBm) except 2.5GHz time-division duplexing (TDD) Band 41 where +26dBm, HPUE is allowed IoT has a lower power-class option at +20dBm	+26dBm for less than 6GHz 5G bands at and above 2.5GHz

Source: 5G RF for dummies (Corvo)

Digital Transmitter

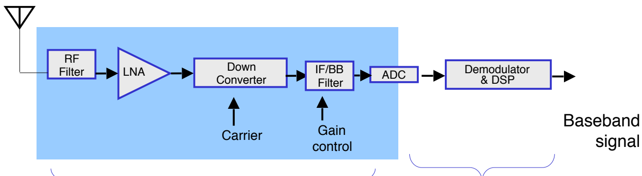


Digital baseband section (compression, coding, modulation, shaping)

RF section (up-conversion, filtering, power gain and control)

Tradeoff between power efficiency and spectral efficiency

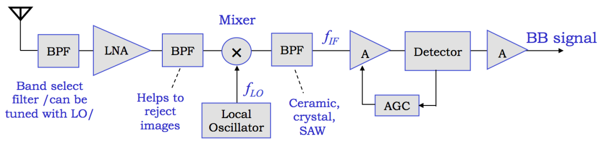
Digital Receiver

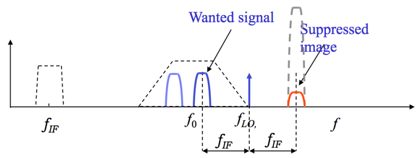


RF frontend: image rejection, low noise, gain control, down conversion, channel selection

Digital baseband section: equalization, demodulation, decoding, decompression

Superheterodyne receiver

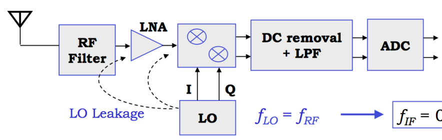




The trade-off between selectivity and sensitivity can be mitigated with better filters or double conversion superheterodyne

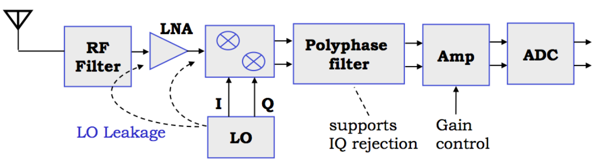
Homodyne receiver (Zero-IF)

- Direct conversion.
- Fewer components, image filtering avoided – no IR and IF filters.
- Large DC offset can corrupt weak signal or saturate LNA (LO mixes itself), notch filters or adaptive DC offset cancellation – eg. by DSP baseband control.
- Flicker noise (1/f) can be difficult to distinguish from signal.
- Channel selection with LPF, easy to integrate, noise-linearity-power tradeoffs are critical, even-order distortions low-freq. beat: differential circuits useful.



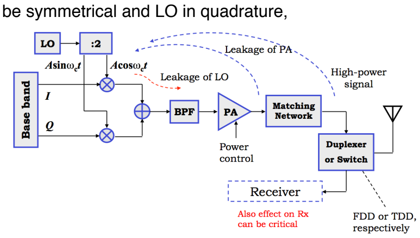
Low-IF receiver

- Tradeoff between heterodyne and homodyne.
- DC offset and 1/f do not corrupt the signal, like in the homodyne, still DC offset must be removed - saturation threat.
- Image problem reintroduced - close image!
- Still even-order distortions can result in low-frequency beat: differential circuits useful but not sufficient



Direct conversion transmitter

- Up-conversion is performed in one step, $f_{LO} = f_c$
- Modulation, e.g. QPSK can be done in the same process
- BPF suppresses harmonics
- LO must be shielded to reduce corruption
- I and Q paths must be symmetrical and LO in quadrature, otherwise crosstalk



Outline of lecture 2

- System Design Fundamentals
  Gu: Chapter 2 (with re-used slides from TSEK02)

- 2.1 Linear systems (shortened)
- 2.2 Nonlinear issues (extended)
- 2.3 Noise
- 2.4 Digital Base-Band (sampling, modulation, errors, …): skipped.

Linear and Nonlinear Systems (2.1.1)

- A system is said to be linear if it follows the superposition rule:

If

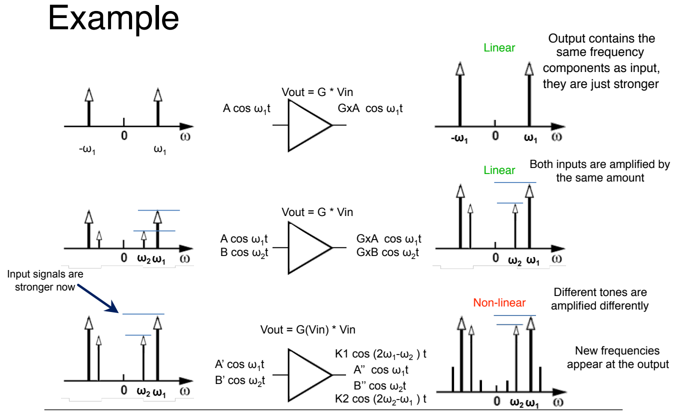
$X_1(t) \rightarrow Y_1(t)$

$X_2(t) \rightarrow Y_2(t)$

then

$aX_1(t) + bX_2(t) \rightarrow aY_1(t) + bY_2(t)$

Example

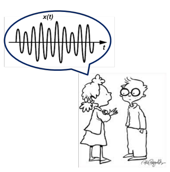


Time Invariance

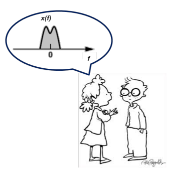
- Time-variant = response depends on the time of origin.
- A system is time-invariant if a time shift in its input results in the same time shift in its output.
  If $y(t) = f\,[x(t)]$
  then $y(t-\tau) = f\,[x(t-\tau)]$
- LTI: Linear time-invariant, $y(t-\tau) = L * [x(t-\tau)]$.
  A time shift in the input causes the same time shift in the output.
- Examples: filters, isolators, duplexers, linear amplifiers.

Time domain vs. frequency domain representation (2.1.2)

Time domain

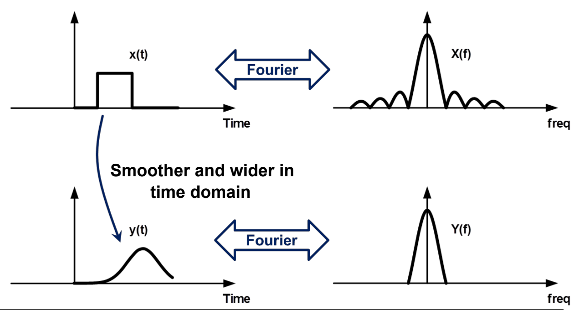


Frequency domain



- The two representations are related through Fourier transformation and both contain the same information

Example – effect of filtering



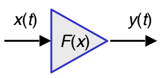
Bandwidth and frequency dependence

- Many different bandwidth definitions:
  - most common is ”3 dB”: range over which the gain drops by less than 3 dB.
- Many systems have frequency dependent transfer functions.
- Amplitude response is usually what we look at.
- But also phase response is usually a nonlinear function of the frequency, caused by varying delay in the circuit, and causing linearity degradation as well as the amplitude.

Fundamentals of System Design (Ch 2)

- 2.1 Band-pass and low-pass RF system representation
- **2.2 Nonlinear issues (extended)**
- 2.3 Noise

Non-linear issues (p. 29-)



- Static model

$$y(t) = F(x(t))$$

$$y(t) = \alpha_0 + \underbrace{\alpha_1 x(t)}_{\text{Linear part}} + \alpha_2 x^2(t) + \alpha_3 x^3(t)$$

nonlinear

3rd order polynomial approximation adequate for weak nonlinearities in RF blocks.

- Dynamic model

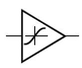
$$\dot{y}(t) = F(y(t), x(t))$$

(eg. RF amplifier)

Also polynomial approximation preferred for *F*() if weakly nonlinear. Volterra series especially for PA analysis to include memory effects.

Effects of Nonlinearity

- Consider a nonlinear memoryless system

x(t)  y(t)= $\alpha_0 + \alpha_1 V_{in} + \alpha_2 V^2_{in} + \alpha_3 V^3_{in} + \ldots$

- Let us apply a single-tone ($A\cos\omega t$) to the input and calculate the output:

$$x(t) = A\cos\omega t$$

$$y(t) = \alpha_1 A\cos\omega t + \alpha_2 A^2 \cos^2 \omega t + \alpha_3 A^3 \cos^3 \omega t$$

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

DC | **Fundamental** | **Second Harmonic** | **Third Harmonic**

Effects of Nonlinearity

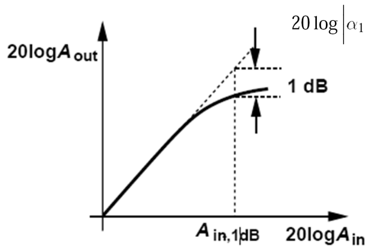
- Observations:
  - Fundamental amplitude reduced by gain compression, $\alpha_1 \cdot \alpha_3 < 0$.
  - Harmonics: not of major concern in a receiver (out of band), but critical in transmitters.
  - Even-order harmonics result from α_j with even j, and same for odd. Can be avoided by differential architecture
  - *n*th harmonic grows in proportion to A^n
  - Amplitude of *nth* harmonic grows in proportion to A^n_{in}

$$y(t) = \alpha_1 A\cos\omega t + \alpha_2 A^2 \cos^2 \omega t + \alpha_3 A^3 \cos^3 \omega t$$

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

Gain (1dB) Compression (p. 32)

- Eventually at large enough signal levels, output power does not follow the input power,



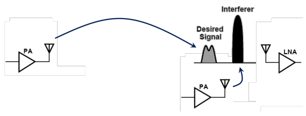
$$20\log\left|\alpha_1 + \frac{3}{4}\alpha_3 A^2_{in,1dB}\right| = 20\log|\alpha_1| - 1\text{ dB}.$$

$$A_{in,1dB} = \sqrt{0.145\left|\frac{\alpha_1}{\alpha_3}\right|}.$$

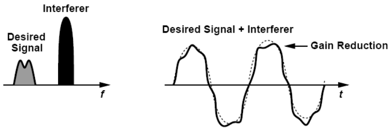
The P-1dB point correlates well to loss of linear behavior, getting out-of-spec in standards (EVM, ACPR, etc.) so for linear applications, operation beyond this point is useless.

Desensitization (p. 32)

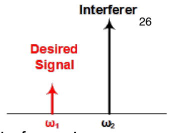
- At the input of the receiver, a strong interference may exist close to the desired signal.



- The small signal is superimposed on the large signal (time domain). If the large signal compresses the amplifiers, it will also affect the small signal.



Desensitization



- Assume $x(t) = A_1\cos\omega_1 t + A_2\cos\omega_2 t$ where A_1 is the desired component at ω_1, A_2 the interferer at ω_2.
- When in compression $y(t) = \left(\alpha_1 + \frac{3}{4}\alpha_3 A_1^2 + \frac{3}{2}\alpha_3 A_2^2\right) A_1 \cos\omega_1 t + \cdots.$
- For $A_1 << A_2$: $y(t) = \left(\alpha_1 + \frac{3}{2}\alpha_3 A_2^2\right) A_1 \cos\omega_1 t + \cdots$

If $\alpha_1\alpha_3 < 0$, the receiver may not sufficiently amplify the small signal A_1 due to the strong interferer A_2. The gain may even drop to zero.

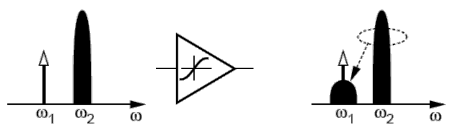
- Also called "blocker". Creates problems when trying to keep the number of filters low.

Cross-modulation (p. 32)

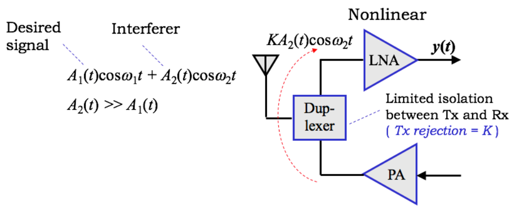
- Any amplitude variation (AM) of the strong interferer A_2 will also appear on the amplitude of the signal A_1 at the desired frequency and distort the signal.
- Interferer: $A_2(1 + m\cos\omega_m t)\cos\omega_2 t$ results in:

extra AM modulation occurs

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



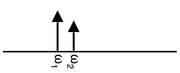
Cross-modulation example

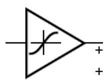


Typical for FDD transceivers with non-constant amplitude modulation, e.g. QAM

Intermodulation distortion (p. 33)

$A_1\cos\omega_1 t + A_2\cos\omega_2 t$

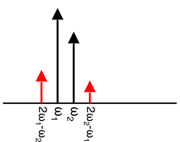




$\alpha_1(A_1\cos\omega_1 t + A_2\cos\omega_2 t)$
$+ \alpha_2(A_1\cos\omega_1 t + A_2\cos\omega_2 t)^2$
$+ \alpha_3(A_1\cos\omega_1 t + A_2\cos\omega_2 t)^3$

$= \frac{1}{2}\alpha_2(A_1^2 + A_2^2)$
$+ [\alpha_1 + \frac{3}{4}\alpha_3 A_1^2 + \frac{3}{2}\alpha_3 A_2^2] A_1\cos\omega_1 t$
$+ [\alpha_1 + \frac{3}{4}\alpha_3 A_2^2 + \frac{3}{2}\alpha_3 A_1^2] A_2\cos\omega_2 t$
$+ [\alpha_2 A_1 A_2]\cos(\omega_1 \pm \omega_2)t$
$+ [\frac{1}{2}\alpha_2 A_1^2]\cos 2\omega_1 t$
$+ [\frac{1}{2}\alpha_2 A_2^2]\cos 2\omega_2 t$
$+ [\frac{3}{4}\alpha_3 A_1^2 A_2]\cos(2\omega_1 + \omega_2)t$
$+ [\frac{3}{4}\alpha_3 A_2^2 A_1]\cos(2\omega_2 + \omega_1)t$
$+ [\frac{3}{4}\alpha_3 A_1^2 A_2]\cos(2\omega_1 - \omega_2)t$
$+ [\frac{3}{4}\alpha_3 A_2^2 A_1]\cos(2\omega_2 - \omega_1)t$
$+ [\frac{1}{4}\alpha_3 A_1^3]\cos 3\omega_1 t$
$+ [\frac{1}{4}\alpha_3 A_2^3]\cos 3\omega_2 t$

They are called Third Order Intermodulation (IM3) products



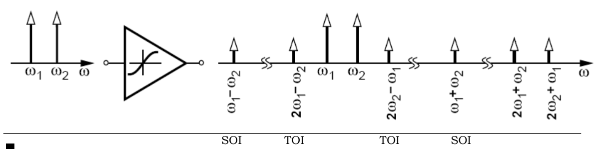
Intermodulation

Fundamental components:

$$\omega = \omega_1,\ \omega_2:\ \left(\alpha_1 A_1 + \frac{3}{4}\alpha_3 A_1^3 + \frac{3}{2}\alpha_3 A_1 A_2^2\right)\cos\omega_1 t + \left(\alpha_1 A_2 + \frac{3}{4}\alpha_3 A_2^3 + \frac{3}{2}\alpha_3 A_2 A_1^2\right)\cos\omega_2 t$$

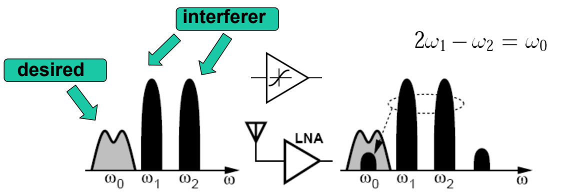
Intermodulation products:

$$\begin{aligned}\omega &= 2\omega_1 \pm \omega_2:\ \frac{3\alpha_3 A_1^2 A_2}{4}\cos(2\omega_1 + \omega_2)t + \frac{3\alpha_3 A_1^2 A_2}{4}\cos(2\omega_1 - \omega_2)t \\ \omega &= 2\omega_2 \pm \omega_1:\ \frac{3\alpha_3 A_1 A_2^2}{4}\cos(2\omega_2 + \omega_1)t + \frac{3\alpha_3 A_1 A_2^2}{4}\cos(2\omega_2 - \omega_1)t\end{aligned}$$



Intermodulation

- IM3 products do not interfere with main tones, so why should we be worried? But they interfere with adjacent channels!
- Intermodulation products are troublesome both in transmitter and in receiver.

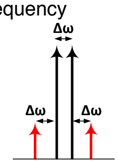


Intermodulation - Characterization

- We generate a test signal for Intermodulation characterization
  - Test Signal : $A_1 \cos \omega_1 t + A_2 \cos \omega_2 t = A \cos \omega_1 t + A \cos (\omega_1 + \Delta\omega)t$
  - Assumptions:
    - $A_1 = A_2 = A$
    - $\Delta\omega = \omega_2 - \omega_1$

this is called the two-tone test

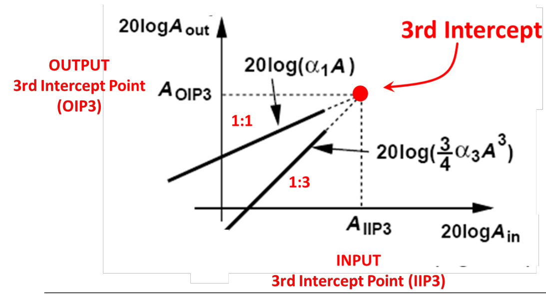
- We write the output signal again (after filtering high frequency components)
  Out = $[{}^3/_4 \alpha_3 A^3] \cos (\omega_1 - \Delta\omega)t$
  $+[\alpha_1 A + {}^9/_4 \alpha_3 A^3] (\cos\omega_1 t + \cos (\omega_1+\Delta\omega)t)$
  $+[{}^3/_4 \alpha_3 A^3] \cos (\omega_1 + 2\Delta\omega)t$



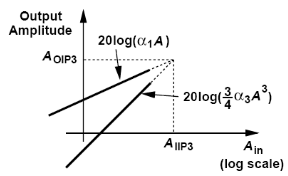
1) Frequency separations are the same
2) Before compression, main tones grow with A and IM3 products grow with A^3

Intermodulation – Intercept Point

OIP3[dBm]=IIP3[dBm] + G[dB]
OIP3[dBm]=P_1[dBm] + 0.5ΔP[dB]



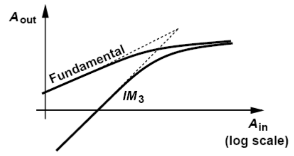
Intermodulation – Intercept Point

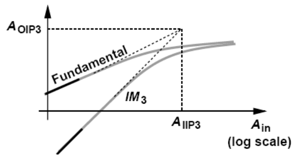


$$|\alpha_1 A_{IIP3}| = \left|\frac{3}{4}\alpha_3 A^3_{IIP3}\right|$$

$$A_{IIP3} = \sqrt{\frac{4}{3}\left|\frac{\alpha_1}{\alpha_3}\right|}.$$

$$\frac{A_{IIP3}}{A_{1dB}} = \sqrt{\frac{4}{0.435}} \approx 9.6 \text{ dB.}$$





Intermodulation: IP_3 and P_{1dB}

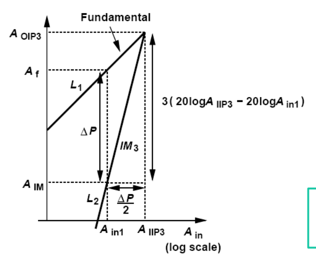
- The actual meaning of this expression is that under ideal conditions, the IP_3 is 9.6 dB higher then the P_{1dB} point (saturation).

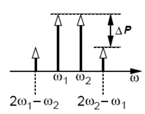
$$\frac{A_{IIP3}}{A_{1dB}} = \sqrt{\frac{4}{0.435}} \approx 9.6 \text{ dB.}$$

- So if you know one of these number, you can estimate the other.

Intermodulation – Intercept Point

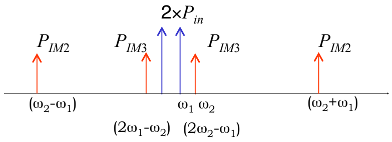
For a given input level (well below P1dB), the IIP3 can be calculated by halving the difference between the output fundamental and IM levels and adding the result to the input level, where all values are expressed as logarithmic quantities.





$$IIP_3|_{dBm} = \frac{\Delta P|_{dB}}{2} + P_{in}|_{dBm}$$

Third and second order intercept points

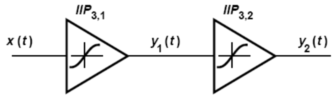


$$IIP_3 = P_{in} - IM_3/2 = P_{in} - (P_{IM3} - P_{in})/2 = (3P_{in} - P_{IM3})/2$$

$$IIP_2 = P_{in} - IM_2 = P_{in} - (P_{IM2} - P_{in}) = 2P_{in} - P_{IM2}$$

Useful formulas to calculate IP2 and IP3.
See derivation in Appendix 3a, p. 211

Cascaded Nonlinear Stages



$$y_1(t) = \alpha_1 x(t) + \alpha_2 x^2(t) + \alpha_3 x^3(t)$$
$$y_2(t) = \beta_1 y_1(t) + \beta_2 y_1^2(t) + \beta_3 y_1^3(t)$$

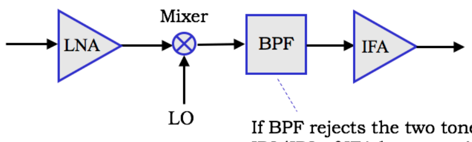
$$y_2(t) = \beta_1[\alpha_1 x(t) + \alpha_2 x^2(t) + \alpha_3 x^3(t)] + \beta_2[\alpha_1 x(t) + \alpha_2 x^2(t) + \alpha_3 x^3(t)]^2 + \beta_3[\alpha_1 x(t) + \alpha_2 x^2(t) + \alpha_3 x^3(t)]^3.$$

Cascaded Nonlinear Stages

- If each stage in a cascade has a gain greater than unity, the nonlinearity of the latter stages becomes increasingly more critical because the IP3 of each stage is equivalently scaled down by the total gain preceding that stage.

$$\frac{1}{IIP3_{total}} = \frac{1}{IIP3_A} + \frac{G_A}{IIP3_B} + \frac{G_A G_B}{IIP3_C}$$

Intermodulation example



If BPF rejects the two tones then IP3/IP2 of IFA less meaningful or can be neglected

(k = gain)

$$\frac{1}{A_{IP3}^2} \approx \frac{1}{A_{IP3,LNA}^2} + \frac{k_{LNA}^2}{A_{IP3,Mix}^2} + \frac{k_{LNA}^2 k_{Mix}^2}{A_{IP3,BPF}^2} + \frac{k_{LNA}^2 k_{Mix}^2 k_{BPF}^2}{A_{IP3,IFA}^2}$$

$$IIP_3 = 10\log\left(\frac{A_{IP3}^2}{2\times 50\Omega}\Big/ 1mW\right) dBm$$

Fundamentals of System Design (Ch 2)

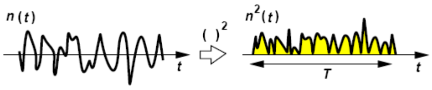
- 2.1 Band-pass and low-pass RF system representation
- 2.2 Nonlinear issues (extended)
- **2.3 Noise**

Noise (p. 37)

- Thermal noise most important. Random currents due to Brown motion of electronics.
- Flicker noise (1/f).
- Impulse noise (spikes).
- Quantization noise (ADC, DAC)(not fully random).
- Shot noise.

Noise

- Average noise power:

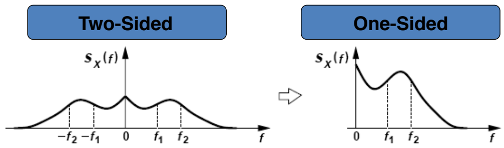


$$P_n = \lim_{T\to\infty} \frac{1}{T} \int_0^T n^2(t)dt \qquad (2.3.6)$$

- T must be long enough to accommodate several cycles of the lowest frequency.

Power Spectral Density (PSD)

- Power Spectral Density (PSD) is the $S_x(f)$.
- Total area under $S_x(f)$ represents the average power carried by $\int_0^\infty S_x(f)df = \lim_{T\to\infty} \frac{1}{T} \int_0^T x^2(t)dt$ (2.3.17)
- Two-sided spectrum is scaled down vertically by factor of 2.



Thermal Noise

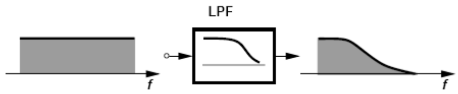
- Charge carriers, which are thermally affected generate a random varying current. It produces a random voltage which is called “thermal noise”.
- Thermal noise power is proportional to *T [K]*. The one-sided PSD of a resistor is given by:

$S_v(f) = 4kTR$ (k=1.38E-23 J/K) [V²/Hz]

- It is independent of frequency, because it is considered as “white” noise (noise power is the same over any given absolute bandwidth).

2.3.2 Effect of Transfer Function on Noise

- Effect of a low-pass filter on white noise:



- Generally, with a transfer function H(s):

$$S_y(f) = S_x(f)|H(f)|^2 \qquad \text{similar to (2.3.36)}$$

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

Noise Figure/Factor

- Noise figure (factor) shows the noise performance of a system.

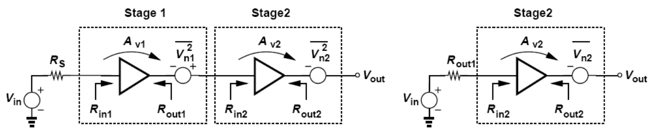
$$NF = \frac{SNR_{in}}{SNR_{out}} \qquad \text{Noise Factor}$$

- It can be expressed in dB:

$$NF\,|_{dB} = 10\log\frac{SNR_{in}}{SNR_{out}} \qquad \text{Noise Figure}$$

- NF depends on not only the noise of the circuit under consideration but the SNR provided by the preceding stage.
- If ideally a system adds no noise, F=1.
- If the input signal contains no noise, NF=∞.

Noise Figure of Cascaded Stages



$$NF_{tot} = 1 + (NF_1 - 1) + \frac{NF_2 - 1}{A_{P1}} + \cdots + \frac{NF_m - 1}{A_{P1}\cdots A_{P(m-1)}}$$

Gain is power gain, which depends on the impedance of each stage.

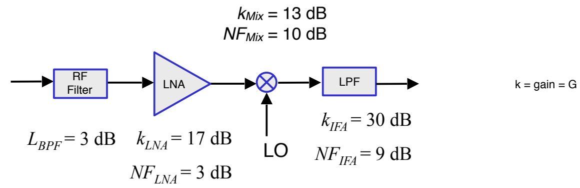
Noise Figure of Cascaded Stages

- Simplified:

$$F_{sys} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \ldots \frac{F_n - 1}{G_1 G_2 \ldots G_{n-1}}$$

- Called "Friis' equation": the noise contributed by each stage decreases as the total gain preceding that stage increases, implying that the first few stages in a cascade are the most critical.

Noise Figure of Cascaded Stages



Friis formula:

$$F = F_{BPF} + \frac{F_{LNA} - 1}{G_{BPF}} + \frac{F_{Mix} - 1}{G_{BPF} G_{LNA}} + \frac{F_{LPF} - 1}{G_{BPF} G_{LNA} G_{Mix}}$$

$$= 2 + \frac{2-1}{0.5} + \frac{10-1}{0.5 \times 50} + \frac{8-1}{0.5 \times 50 \times 20} = 4.37 = 6.4\,\text{dB}$$

Noise temperature (pp. 54-58)

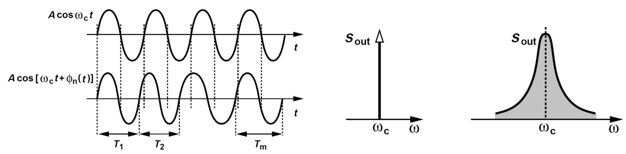
- Sometimes you can see the term "Noise Temperature". What is this?
- Definition: "the equivalent temperature of a source impedance into a perfect (noise-free) device that would produce the same added noise".

$$F = 1 + \frac{T_N}{T_{ref}} \Rightarrow T_N = T_{ref}(F - 1)$$

- Ex. NF=3 dB => T_N = 289 K (T_{ref} = 290 K)
  NF=1 dB => T_N = 75 K

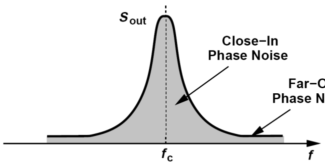
Phase noise

- The phase of the oscillator varies as $A\cos(\omega_c(t)+\Phi_n(t))$.
- The term $\Phi_n(t)$ is called the "phase noise."
- Can also be viewed as a random frequency variation, leading to a broadening of the spectrum called phase noise.



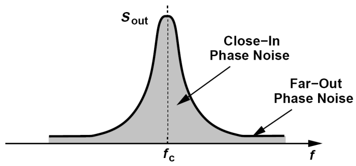
Phase noise

- Since the phase noise falls at frequencies farther from ω_c, it must be specified at a certain "frequency offset," i.e., a certain difference with respect to ω_c.
- We consider a 1-Hz bandwidth of the spectrum at an offset of Δf, measure the power in this bandwidth, and normalize the result to the "carrier power", called "dB with respect to the carrier".



Phase noise

- In practice, the phase noise reaches a constant floor at large frequency offsets (beyond a few megahertz).
- We call the regions near and far from the carrier the "close-in" and the "far-out" phase noise, respectively.



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