# TSEK38: Radio Frequency Transceiver Design
# Lecture 6: Receiver Synthesis (I)


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## Lecture schedule

w4:

- Le1: Introduction (Ch 1)
- Le2: Fundamentals of RF system modeling (Ch 2)
- Le3: Superheterodyne TRX design (Ch 3.1)

w6:

- Le4: Homodyne TRX design (Ch 3.2)
- Le5: Low-IF TRX design (Ch 3.3)
- Le6: Systematic synthesis (calculations) of RX (Ch 4)

w7:

- Le7: Systematic synthesis (continued)
- Le8: Systematic synthesis (calculations) of TX (Ch 5)

w8:

- Le9: Systematic synthesis (continued)





## Lab schedule

w6:

- We: Lab1a (after Le6): 15-19 (ASGA)
- Th: Lab1b: 17-21 (SOUT)

w7:

- We: Lab1c (after Le8): 15-19 (SOUT)
- Th: Lab1d: 17-21 (EGYP)

Instructions:

- One long lab (4 x 4 h).
- Lab manual in the Lisam *Course Documents/2019/Lab* folder.
- Supervision (Ted) available at the times above.
- To pass: complete and document the exercises in the lab manual, go through with Ted.





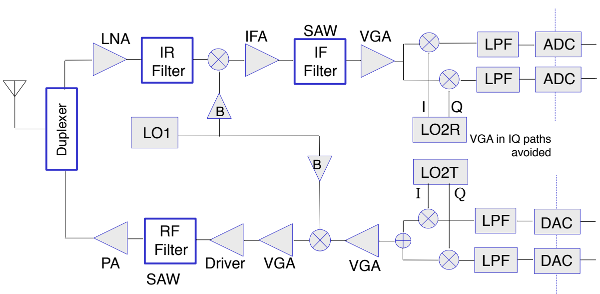
## Repetition of lectures 3 - 5

- Heterodyne Architecture
- Homodyne Architecture
- Low-IF Architecture





## **Heterodyne** with analog IF architecture
## FDD, one antenna, shared LO1




RF, IF filters and duplexer not integrated → matching issues.

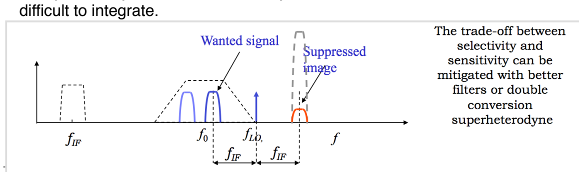
Most of gain at IF (75 %) and RF. IF gain is more power efficient





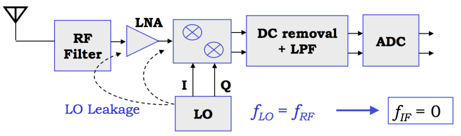
## Heterodyne receiver

- Careful frequency planning required (IF frequencies).
- Many possible intermodulation effects must be considered.
- Pulling of LO by TX avoided.
- Trade-off between sensitivity and selectivity in RX reduced by 2nd stage.
- LO in first stage can be shared, giving different $IF_{TX}$ and $IF_{RX}$.
- Trade-off between sensitivity and linearity and power in RX.
- Most of RX gain at IF or BB (after removing blockers).
- Today heterodyne architecture mostly for TX rather than for RX since IR filters difficult to integrate.



# 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
- IP2 of mixer critical
- 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

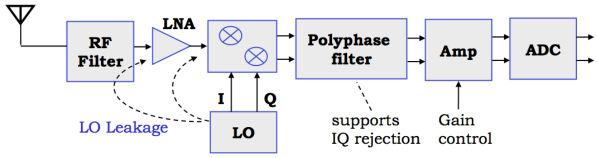







# Low-IF receiver

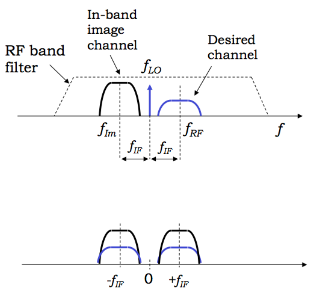
- Tradeoff between heterodyne and homodyne
- Blocking profile dictates IF = BW/2.
- DC offset and 1/f do not corrupt the signal (cf. homodyne).
- Image problem reintroduced - close image!
- Still even-order distortions can result in low-freq. beat: differential circuits useful but not sufficient








# Close-image problem




Tough requirements for IQ match if image is large, otherwise signal strongly corrupted

$f_{IF}$ = ½ BW typical



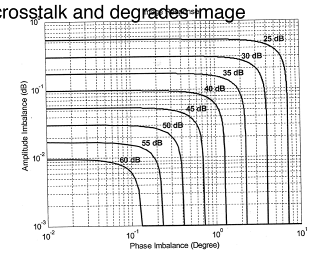


# IQ imbalance and Image rejection

- IQ imbalance causes image crosstalk and degrades image rejection.

$$IR = 10\log\frac{P_{Signal}}{P_{Image}}$$

$$IR = 10\log\frac{1+2(1+\delta)\cos\varepsilon+(1+\delta)^2}{1-2(1+\delta)\cos\varepsilon+(1+\delta)^2}$$




δ - amplitude (gain) imbalance
ε - phase imbalance

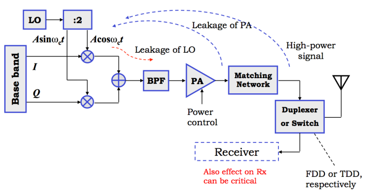
IR is partly insensitive,
e.g. for 0.3d B gain imbalance, IR=30dB
keeps up to 2° phase imbalance





# 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








# Systematic Receiver Synthesis (1)

- 4.1 Introduction
- 4.2 Sensitivity, Noise Figure
  - Receiver Desensitization from transmitter emissions
  - Mismatch between antenna and Rx
- 4.3 Intermodulation characteristics
  - Phase noise degradation

# 4.1 Introduction

- Many different ways to realize the receiver. Somewhat fewer for transmitter.
- TDD half-duplex: GSM, WLAN, DECT, …, 5G,
- FDD full-duplex: WCDMA, LTE, …
- Usually different frequency bands for RX and TX.
- The TX signal may be 120 dB stronger than the RX signal.
- FDD receivers are generally trickier to implement because of TX leakage into RX, VCO pulling from LO, ...





# Introduction

- Key design parameters, RX:
  - sensitivity: overall noise figure (4.2)
  - intermodulation: 3rd order distortion, IP3 (4.3) also IP2
  - single-tone desensitization (4.4, not!)
  - adjacent/alternate channel selectivity: channel filter, phase noise, (4.5)
  - interference blocking: channel filter, phase noise, (4.5)
  - dynamic range: AGC, ADC (4.6).
- Book: Full-duplex, but can be applied half-duplex (some not relevant for half-duplex).





# 4.2 Sensitivity and NF

- Sensitivity = weakest detectable signal to obtain a minimum SNR for achieving required BER.
- Simple model: noise at input is white noise. Directly related to the overall NF of the RX.
- Varies because: signal modulation and characteristics, signal propagation in the channel, external noise level.
- In this book: Carrier-to-Noise Ratio (CNR) for the analog signal. SNR is for the digital BB. But we will use S and SNR!
- Calculation of sensitivity and NF: pp. 230-232.





# AWGN

Additive white Gaussian noise (AWGN) is a basic noise model to mimic the effect of many random processes that occur in nature:

- Additive because it is added to any noise that might be intrinsic to the information system.
- White refers to the idea that it has uniform power across the frequency band for the information system. It is an analogy to the color white which has uniform emissions at all frequencies in the visible spectrum.
- Gaussian because it has a normal distribution in the time domain with an average time domain value of zero.

[Wikipedia]





# $E_b/N_0$ vs. Signal-to-Noise Ratio

- A better measure of signal-to-noise ratio for digital data is the ratio of energy per bit transmitted ($E_b$) to the noise power density ($N_0$).
- SNR (a quantity which can be measured) is related to $E_b/N_0$ (an artificial quantity used in comparisons) by

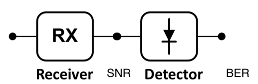
$$SNR = \frac{SignalPower}{NoisePower} = \frac{E_b}{N_0}\frac{R_b}{B}$$

$\frac{R_b}{B}$ is the spectral density (bitrate / bandwidth).

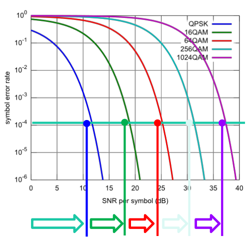




# BER vs SNR




- BER depends on SNR of the received signal, i.e. the signal after the receiver block.
- More complicated modulation schemes require higher SNR for the same error (trade off between BW and BER)
- It may be possible to correct errors with advanced Forward Error Correction (FEC) Coding (reduce BER for the same SNR)



# BER versus SNR in demodulation

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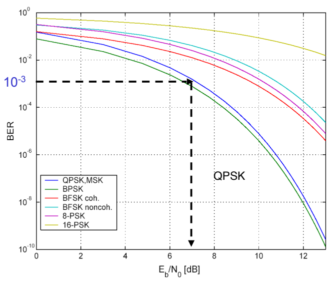


Dependent on modulation and demodulator implementation.

$(S/N) = P_{sig}/N = (E_b R_b)/(N_0 B)$
$(S/N) = E_b/N_0 * R_b/B$
$R_b/B \approx 0.5 - 1.5$ (typically)

$E_b$ = energy per transmitted byte
$N_0$= noise power density
$R_b$ = bit rate
B = channel bandwidth

$SNR_{min} = (E_b/N_0)_{dB} + 10\log(R_b/B)$

For spread-spectrum (e.g. CDMA) Rb/B<<1 => SNRmin < 0





# BER vs. $E_b/N_0$

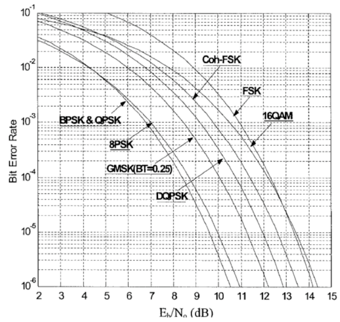


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Figure 2.37. The bit error rate and $E_b/n_o$ of different modulation signals





# 4.2.1 Reference sensitivity and NF

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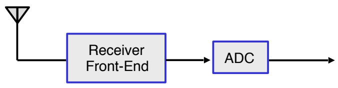
Input signal:
$S_{min}$: Sensitivity
$P_{Ni}$ ($N_{ref}$): integrated thermal noise power within Rx BW.
$SNR_{in}$ ($CNR_{in}$): signal-to-noise




Output signal:
To demodulator
$SNR > SNR_{min}$

Receiver noise factor: $F_{RX} = (S/N)_i / (S/N)_o = (P_S/P_{Ni}) / (S/N)_o$ (4.2.1)

Integrated thermal noise power within Rx BW: $P_{Ni} = k * T_0 * B$ (4.2.2)
Called *Reference noise*. Ground floor for noise.

=> RX input signal power: $P_s = kT_0 * B * F_{RX} * (S/N)_o$ (4.2.3)





# Reference sensitivity and NF

Min $(S/N)_o$ required for BER for the sensitivity level = $(S/N)_{min}$ = $SNR_{min}$.
=> Sensitivity = $S_{min}$[dB] = $10\log(P_{S,min}) = -174 + 10\log(B) + NF_{RX} + SNR_{min}$ (4.2.4)

=> $NF_{RX}$ [dB] = $SNR_{in} - SNR_{out} = SNR_{in} - SNR_{min}$ =
$S_{min}$ - (-174 dBm/Hz + 10log(B)) - $SNR_{min}$ (4.2.7)

Reference noise or noise floor

Calculation flow:
1. BER requirement.
2. Use graph to convert to $E_b/N_0$.
3. Calculate $SNR_{min}$.
4. Calculate $S_{min}$ (sensitivity).
5. Calculate NF.





# Reference sensitivity and NF

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Table 2.4. Estimated maximum CNR for reference static sensitivity of different protocol mobile station receivers

| System | Maximum CNR (dB) | Reference Sensitivity (dBm) | Modulation | Symbol Rate (ksps) | Maximum Error Rate (%) | Note |
|---|---|---|---|---|---|---|
| AMPS | 3.5 | -116 | FM | | | |
| TDMA | 10 | -110 | π/4DQPSK | 24.3 | 3 (BER) | |
| PHS | 11 | -97 | π/4DQPSK | 192 | 1 (BER) | |
| GSM | 8 | -102 | GMSK | 270.833 | 4 (RBER) | TCH/FS Class II |
| CDMA | -1 | -104 | QPSK | 1,228.8 ( kcps) | 0.5 (FER) | 9.6 kbps voice data rate |
| GPRS | 10 | -102/-99 | GMSK | 270.833 | 10 (BLER) | CS1 code/ Uncoded |
| Edge | 12 | -98 | 8PSK | 271 | 10 (BLER) | |
| WCDMA | -7.5 | -106.7 | QPSK | 3,840.0 (kcps) | 0.1 (BER) | 12.2 kbps voice data rate |

Examples of $P_{Ni}$ (reference noise):
GSM: -174+10log(200 kHz) = -121 dBm → $V_n$ = 6.3 µV
BT: -174+10log(1 MHz) = -114 dBm → $V_n$ = 14.1 µV
WCDMA: -174+10log(5 MHz) = -107 dBm → $V_n$ = 31.6 µV





# Reference sensitivity and NF

3

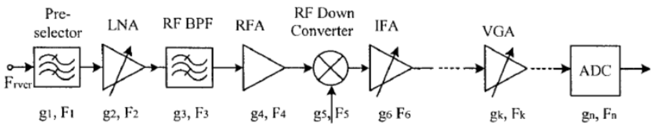
GSM (calc or from the table):
$S_{min}$= -102 dBm, $N_{ref}$ = -121 dBm, $BER_{max}$= $4\times10^{-2}$ (4 %)

=> $SNR_{min} = 6.5 + 10*\log(271/250) \approx 6.8$ dB (book 8 dB)

=> $NF_{Rx}$ = -102 - (-174 - 54) - 6.8 = 11 dB (total max noise)

Specify $NF_{Rx}$ = 7 dB:
- 3 dB margin because of ~3 dB loss in TDD switch and RF filter,
- 1 dB additional margin.

# 4.2.2 Cascaded Noise Figure

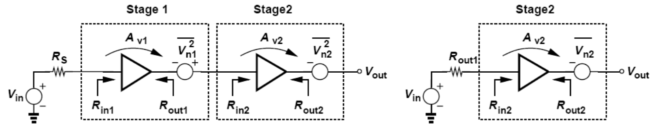



$$NF_{Rx} = 10\log\left(10^{\frac{NF_1}{10}} + \sum_{i=2}^{n}\frac{10^{\frac{NF_i}{10}}-1}{\prod_{j=1}^{i-1}10^{\frac{G_j}{10}}}\right). \quad (4.2.12c)$$





# 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.





# Cascaded Noise Figure: matching

$$NF_{Rx} = 10\log\left(10^{\frac{NF_1}{10}} + \sum_{i=2}^{n}\frac{10^{\frac{NF_i}{10}}-1}{\prod_{j=1}^{i-1}10^{\frac{G_j}{10}}}\right). \quad (4.2.12c)$$

Equation (4.2.12) is also called a *Friis equation* [2]. The gains in this equation are *available power gain* based on the assumption that there is conjugate matching between stages. The assumption of conjugate match may be in reality true for the RF blocks of the receiver. However, in the IF and the analog BB blocks of the receiver the input impedance of the *m*th stage $Z_{i,m}$ may not match with the output impedance of its preceding stage $Z_{o,m-1}$ and usually $Z_{i,m} >> Z_{o,m-1}$. The voltage gain $g_v$ instead of the power gain is usually used for the stages in the IF and the analog BB blocks.

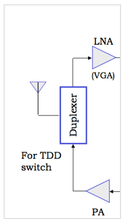
- In our calculations, we will assume matching





# 4.2.3 Receiver desensitization, TX leakage

- We want to find the receiver desensitization caused by TX leakage into the receiver band for an FDD TRX.
- This emission noise passes through the duplexer.
- It can be modelled as:
  - Approach #1: Equivalent Duplexer Noise Figure,
  - Approach #2: Equivalent Antenna Temperature.








# 4.2.3.1 Duplexer noise figure

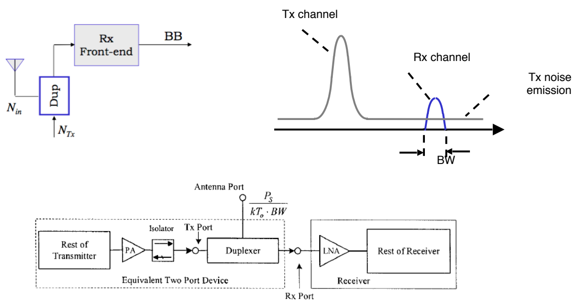



Figure 4.3. Simplified configuration of full-duplex transceiver





# Duplexer noise figure

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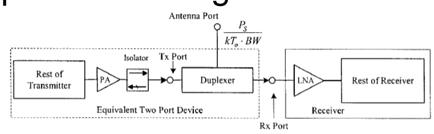


Antenna port: $(S/N)_{input} = \frac{P_S}{kT_o \cdot BW}$ (4.2.24)

Noise at Rx port: $(kT_o + P_{N,Tx\_ant,RxBand} \cdot g_{ant\_Rx}) \cdot BW$

$P_{N,TX\_ant,RxBand}$ = excess transmitter noise...

$g_{ant\_Rx}$ = antenna insertion gain

$$(S/N)_{output} = \frac{P_S \cdot g_{ant\_Rx}}{(kT_o + P_{N,Tx\_ant,RxBand} \cdot g_{ant\_Rx}) \cdot BW}. \quad (4.2.25)$$

# Duplexer noise figure

Noise factor from antenna to Rx (4.2.24 / 4.2.25):

$$F_{e\_dplx} = \frac{(S/N)_{input}}{(S/N)_{output}} = \frac{1}{g_{ant\_Rx}} + \frac{P_{N,Tx\_ant,RxBand}}{kT_o} \qquad (4.2.26a)$$

Noise figure from antenna to Rx:

$$NF_{e\_dplx} = 10\log\left(\frac{1}{g_{ant\_Rx}} + \frac{P_{N,Tx\_ant,RxBand}}{kT_o}\right) \text{dB}. \qquad (4.2.26b)$$





# 4.2.3.2 Antenna Temperature

- Different antenna temp compared to TRX temp

$$kT_e = kT_o + P_{N,Tx\_Rx}. \qquad (4.2.29)$$

effective noise temperature

$$NF_{Rx} = 10\log\left[\frac{P_{N,Tx\_ant,RxBand}}{kT_o} + F_{Rx,o}\right]. \qquad (4.2.32b)$$

- (4.2.32b) is basically the same as (4.2.26b)





# Duplexer noise figure

5

- NF of the duplexer can be largely degraded by TX leakage.

Example: emission noise density in the RX band

$$N_{Tx}\big|_{dB} = -130\text{dBm/Hz}, \quad A = 44\text{dB}$$ attenuation in the duplexer

$$G_{Dup}\big|_{dB} = -2.5\text{dB}$$

$$NF_{Dup} = 10\log\left(10^{\frac{-G_{Dup}|_{dB}}{10}} + 10^{\frac{N_{Tx}|_{dB} - A + 174\text{dBm/Hz}}{10}}\right) = 4.4\text{dB}$$

- If Tx is off, then $NF_{Dup}$ = 2.5 dB





# Comparison: NF of TDD and FDD Rx

6

TDD: $NF_{BPF+Switch} = L_{BPF+Switch} = -G_{BPF+Switch}$

FDD: $NF_{Dup} = L_{Dup} = -G_{Dup}$

$$NF_{Rx} = 8\text{dB}, \quad L_{BPF+Switch}\big|_{dB} = 2.5\text{dB}$$
$$NF_{Rx} = L_{BPF+Switch}\big|_{dB} + NF_{F-E}$$

TDD: $NF_{F-E}$ =5.5 dB

$$NF_{Rx} = 8\text{dB}, \quad L_{Dup}\big|_{dB} = 2.5\text{dB}$$
$$NF_{Dup} = 4\text{dB}$$
$$F_{Rx} = F_{Dup} + \frac{F_{F-E} - 1}{G_{Dup}}$$

FDD: $NF_{F-E}$ =5 dB





# 4.2.4 Mismatch between antenna and Rx

- VSWR = Voltage standing wave ratio
- Impedance mismatch
- Causes standing waves and high voltage peaks
- Two waves, forward and reflected:

$$\text{VSWR} = \frac{|V_{max}|}{|V_{min}|} = \frac{1+|\Gamma|}{1-|\Gamma|} \qquad \Gamma = \frac{V_r}{V_f}$$

- Varies between 1.5 and 6 may (lower is better).
- Special case when load $R_L$ is purely resistive:

$$\text{SWR} = \left(\frac{R_L}{Z_0}\right)^{\pm 1}$$





# Mismatch between antenna and Rx

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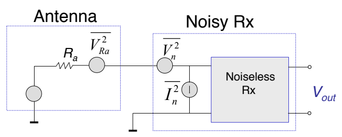



$$F_{Rx} = \frac{SNR_{in}}{SNR_{out}} = 1 + \frac{\overline{(V_n + R_a I_n)^2}}{\overline{V_{Ra}^2}}$$

$$F_{Rx} = 1 + \frac{\overline{V_n^2} + R_a^2 \overline{I_n^2}}{4kTR_a}$$

when uncorrelated

$$R_{opt}^2 = \overline{V_n^2}/\overline{I_n^2} = R_n/G_n$$
$$F_{Rx,\min} = 1 + 2R_n/R_{opt}$$

$$F_{Rx} = 1 + \frac{F_{Rx,\min} - 1}{2} \times \left(\frac{R_a}{R_{opt}} + \frac{R_{opt}}{R_a}\right)$$

$$F_{Rx} = F_{Rx,\min} + R_n R_a (G_a - G_{opt})^2$$
$$= 1 + \sqrt{R_n G_n}\left(\frac{R_a}{R_{opt}} + \frac{R_{opt}}{R_a}\right)$$

For $R_a/R_{opt}$ = 2 (3)

$NF_{min}$= 3 dB, $NF_{Rx}$= 3.5 dB (4.3 dB)
$NF_{min}$= 6 dB, $NF_{Rx}$= 6.8 dB (7.8 dB)

# 4.3 Intermodulation characteristics

4.3.1 Intermodulation products and intercept points

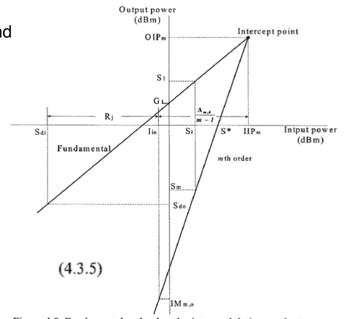



Figure 4.8. Fundamental and *m*th order intermodulation product response

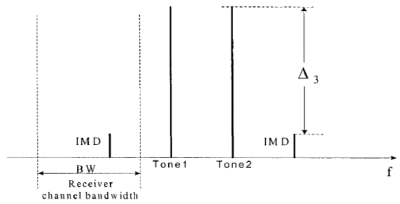
General expression:

$$IM_{mi,k} = mI_{i,k} - (m-1)IIP_{m,k} \ \text{dBm}. \qquad (4.3.5)$$





# Third-order intermodulation product




$$IIP_3 = \frac{1}{2}(3I_{in} - IM_3) = I_{in} + \frac{\Delta_3}{2}, \qquad (4.3.10)$$





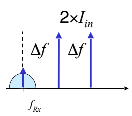
# Intermodulation characteristics

IM3 is the main problem, close to the carrier.

$IIP_3 = (3I_{in} - IM_3)/2 = I_{in} + \Delta_3/2$ (4.3.10)

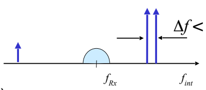
using $\Delta_3 = I_{in} - IM_3$ (see prev. slide)

$IM_3 = 3I_{in} - 2IIP_3$




IP3 test

IM2 is a minor problem, except for the direction conversion receiver.




IP2 test

$IIP_2 = 2I_{in} - IM_2$ [dBm] (4.3.9)

$IM_2 = 2I_{in} - IIP_{2,Rx}$

4.3.2 Cascaded IP…
Can be rather complicated when frequency selectivity and matching are considered.





# Intermodulation characteristics (4.3.3)

- The RX linearity (cascaded IIP for the whole receiver) is the main cause of intermodulation distortion + LO phase noise + …
- *Allowed maximum degradation* $D_{max,in}$ caused by noise and interference:
  - $D_{max,in} = S_{d,i} - SNR_{min}$
  - $S_{d,i}$ = receiver input desired signal = (usually) = $S_{min\_ref}$ + 3 dB = reference sensitivity level + 3 dB.
  - $SNR_{min}$ = minimum signal to noise/interference ratio





# 4.3.3.1 Degradation because of inherent noise

- $N_{nf}$ = receiver inherent noise (from NF, converted to the input port):
  $N_{nf} = -174 + 10\log(B) + NF_{RX}$.
- $N_{nf}$ reduces the $D_{max,in}$:

$$D_a = 10\log\left(10^{\frac{D_{max,in}}{10}} - 10^{\frac{N_{nf}}{10}}\right), \qquad (4.3.39)$$

- Example: assume GSM with $S_{min}$ = reference sensitivity = -102 dBm => $S_{d,i}$ = -99 dBm, $SNR_{min}$ = 8 dB, $NF_{RX}$ = 10 dB, B = 250 kHz.

=> $D_a$ = 10log(1.9953E-11 - 9.9527E-12) = -110 dBm (-107 with Dm only)

```
(Dm=10^((-99-8)/10);Nf=10^((-174+10+10*log10(250E3))/10);Da=10*log10(Dm-Nf))
```





# Alternative calculations (not in the book)

- Calculations of $SNR_{min}$ when degraded by noise, linearity, etc.
- $P_{IM\#}$ is the cascaded $IIP_{RX}$ (mostly $IIP_{3,RX}$ and $IIP_{2,RX}$).
- $N_{other}$ = other noise sources, usually assumed <= -100 dBm

$$(S/N)_{min} = \frac{P_{in}}{N_{in} + N_{ref}(F_{Rx} - 1)}, \quad P_{in} > P_{in,min} \ (+3dB)$$

$$N_{in} = N_{ref} + P_{IM\#} + N_{Other}$$

$$P_{IM\#} = \frac{P_{in}}{(S/N)_{min}} - \underbrace{F_{Rx}N_{ref}}_{\text{Rx noise floor}} - N_{Other}$$

Headroom for $P_{IM\#}$ limited by $SNR_{min}$ and $N_{Other}$ esp. LO phase noise + spurious

Additionally $P_{IM2}$ limited by self-mixing of the test signal (interferer) as discussed for Zero-IF Rx

# $N_{in}$ estimation



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$$N_{in} = N_{ref} + P_{IM\#} + N_{Other} \qquad P_{IM\#} = \frac{P_{in}}{(S/N)_{min}} - F_{Rx}N_{ref} - N_{Other}$$

$$10\log\left(10^{\frac{P_{IM\#}|_{dBm}}{10}} + 10^{\frac{N_{Other}|_{dBm}}{10}}\right)$$

$$= 10\log\left(10^{\frac{S_{in}-SNR_{min}}{10}} - 10^{\frac{-174+10\log BW+NF_{Rx}}{10}}\right) \cong -94.5\,\text{dBm}$$

$I_{in}$ = -38 dBm
$S_{in}$ = -86 dBm
$BW$ = 1.25 MHz
$SNR_{min}$ = 8 dB
$NF_{Rx}$ = 10 dB

+ headroom for extra noise and distortion



# IP3/IP2 requirement estimation



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$I_{in}$ = -38 dBm
$S_{in}$ = -86 dBm
$BW$ = 1.25 MHz
$SNR_{min}$ = 8 dB
$NF_{Rx}$ = 10 dB
$P_{bl}$ = -30 dBm

$$P_{IM\#}\big|_{dBm} = 10\log\left(10^{\frac{S_{in}-SNR_{min}}{10}} - 10^{\frac{-174+10\log BW+NF_{Rx}}{10}} - 10^{\frac{N_{Other}|_{dBm}}{10}}\right)$$

$$N_{Other}\big|_{dBm} = -100\,\text{dBm} \;\Rightarrow\; P_{IM\#}\big|_{dBm} \cong -96\,\text{dBm}$$

$$IIP_{3,Rx} = \frac{3(-38)-(-96)}{2} = -9\text{dBm}$$

When all distortion is attributed to IP3.

$$IIP_{3,Rx} = \frac{3(-38)-(-99)}{2} = -7.5\text{dBm}$$

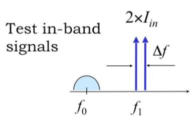
$$IIP_{2,Rx} = 2(P_{bl}-3)-(-99) = 33\text{dBm}$$

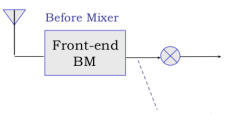
When distortion is attributed both to IP3 and IP2 by half each.



# IIP2 of RX






$$IIP_{2\_Mix} = 2I_{in\_Mix} - P_{IM2\_Mix}\big|_{dBm}$$
$$= 2I_{in} - P_{IM2\_in}\big|_{dBm} + G_{BM}$$

$$I_{in\_Mix} = I_{in} + G_{BM}$$
$$P_{IM2\_Mix}\big|_{dBm} = P_{IM2\_in}\big|_{dBm} + G_{BM}$$

$$(S/N)_{min} = \frac{P_{in}}{N_{in} + N_{ref}(F_{Rx}-1)}$$

**SNR budget**

$$= \frac{P_{in}}{(N_{ref} + N_{other} + P_{IM2\_in}) + N_{ref}(F_{Rx}-1)}$$

Noise + IM2 Distortion — Inherent Noise

⇨ Tradeoff between IM2 and NF + other noise effects (PN+spurious)

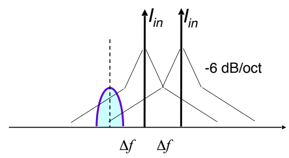


# 4.3.3.3 Phase noise and spur degradation



- Phase noise (PN) and spurious emissions (SP) from LO contaminate the signal. Will increase the $IIP_{x,min}$.
- Intermodulation interference tones will mix with the LO PN ($N_{PN}$ = average phase noise density) and/or spurios ($N_{SP}$ at ) and generate in-band noise and spurs, which degrades the CNR.
- Covered by $N_{other}$: $N_{PN} + N_{Sp} <= N_{other}$

$$PN(\Delta f) = PN(2\Delta f)\text{+6dB}$$
$$\Delta f = n \times BW$$






# Phase noise and spur degradation



- PN(Δf) = average PN density over Rx BW at offset Δf

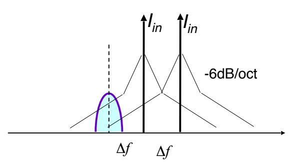
$$PN(\Delta f) = \underbrace{N_{PN}(\Delta f) - 10\log BW}_{P_{PN}|_{\text{dBm/Hz}}} - I_{in} \quad [\text{dBc/Hz}] \qquad (4.3.37)$$

$$N_{PN} = 10^{\frac{PN(\Delta f)+I_{in}+10\log BW}{10}} + 10^{\frac{PN(2\Delta f)+I_{in}+10\log BW}{10}}$$

$$N_{Sp} = 10^{\frac{N_{Sp}(\Delta f)+I_{in}}{10}} + 10^{\frac{N_{Sp}(2\Delta f)+I_{in}}{10}}$$

- In practice, we set $N_{Sp} \approx N_{PN}$

$$PN(\Delta f) = PN(2\Delta f)\text{+6dB}$$
$$\Delta f = n \times BW$$






# Phase noise estimation in two-tone test



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$$N_{PN} + N_{Sp} \approx 2\times10^{\frac{PN(\Delta f)+I_{in}+10\log BW}{10}} + 2\times10^{\frac{PN(2\Delta f)+I_{in}+10\log BW}{10}}$$

$N_{Sp} \approx N_{PN}$

$$= 2\times10^{\frac{PN(\Delta f)+I_{in}+10\log BW}{10}}\left(1+10^{\frac{-6dB}{10}}\right)$$

$$PN(\Delta f) = PN(2\Delta f)\text{+6dB}$$

$$N_{PN+Sp}\big|_{dBm} = 10\log\left(2.5\times10^{\frac{PN(\Delta f)+I_{in}+10\log BW}{10}}\right)$$
$$= PN(\Delta f) + I_{in} + 10\log BW + 4\text{dB} \le N_{Other}\big|_{dBm}$$

$$PN(\Delta f) \le N_{Other}\big|_{dBm} - I_{in} - 10\log BW - 4\text{dB}$$
$$= -100 + 36 - 61 - 4$$
$$PN(\Delta f) \le -129\,\text{dBc/Hz} \;\rightarrow\; -135\,\text{dBc/Hz}\ @\Delta f$$

with 6 dB margin

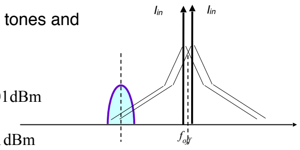
From this, the impact on IIP3 can be calculated (eq 4.3.51).

## Phase noise estimation, IP2/Blocking test



- SNR budget under IP2 test (homodyne Rx).
- $N_{PN+SP} + N_{TX,SM} \le N_{other}$
- The contribution by Tx leakage and self modulation is estimated to -107 dBm in this example (FDD).
- +6 dB is equivalent to 2x2 Pn (PN of 2 tones and 2x for LO spurs).

2x-30dBm = -27dBm




$$N_{PN+Sp}\Big|_{dBm} \le 10\log\left(10^{\frac{N_{Other}|_{dBm}}{10}} - 10^{\frac{-107dBm}{10}}\right) = -101\,\text{dBm}$$

$$PN(f_{off}) + (I_{in} + 3\text{dB}) + 10\log BW + 3\text{dB} \le -101\,\text{dBm}$$

$$PN(f_{off}) \le -138\,\text{dBm}$$

For $f_{off}$ >> Δf the PN approx. the same

$$N_{PN} + N_{Sp} \approx 2\times 10^{\frac{PN(f_{off}) + I_{in} + 3\text{dB} + 10\log BW}{10}}$$



## IIP2 of RX

additional calculations



$$SNR_{min} = S_{in} - 10\log(N_{ref}F_{Rx} + N_{other} + P_{IM2\_in})$$

$$P_{IM2\_in}\Big|_{dBm} = 10\log\left(10^{\frac{S_{in}-SNR_{min}}{10}} - 10^{\frac{-174+10\log BW + NF_{Rx}}{10}} - 10^{\frac{N_{other}}{10}}\right)$$

$$IIP_{2\_Mix} = 2I_{in} - P_{IM2\_in}\Big|_{dBm} + G_{BM}$$

Headroom for IM2 product is limited by SNR, thermal noise and contribution from other sources (PN, LO spurious and RF-to-LO leakage)



## Estimation of IIP2

additional calculations



**Test specs:**

$I_{in}$ = -30 dBm
$S_{in}$ = -86 dBm
$BW$ = 1.25 MHz
$SNR_{min}$ = 8 dB $NF_{Rx}$ = 10 dB
($S_{ref}$ = -95 dBm)
($N_{in\ mx}$ = -95 dBm)
$N_{other}$ = -100 dBm

$$P_{IM2\_in}\Big|_{dBm} = 10\log\left(10^{\frac{-86-8}{10}} - 10^{\frac{-174+10\log(1.25+10^6)+10}{10}} - 10^{\frac{-100}{10}}\right)$$

$$P_{IM2\_in}\Big|_{dBm} \cong -96\,\text{dBm}$$

$$IIP_{2\_Mix} = 2\times(-30) + 96 + G_{BM} = 36\,\text{dBm} + G_{BM}$$

$G_{BM}$ = 15.. 25 dB ⇨ $IIP_{2\text{-}Mix} \approx$ 51.. 61 dBm

**Exercise:** How much IIP2 should be increased for

1) $N_{other}$ increased by 3 dB ?
2) What is the upper bound for $N_{other}$?



## Estimation of Tx leakage vs IIP2

additional calculations



**We estimate contribution of Tx leakage to $N_{other}$**

$P_{Tx}$ = 27 dBm
$R_{Dupl}$ = -48 dB
$R_{BPF}$ = -30 dB
$G_{BM}$ = 15.. 25 dB
$N_{other}$ = -100 dBm
$IIP2_{in}$ = 36 dBm

**Path_1: (according to IM2)**

$$P_{Tx\_Mix} = P_{Tx} + R_{Dupl} + G_{BM} + R_{BPF} = -51 + G_{BM}$$

$$IIP_{2\_Mix} = 2I_{in\_Mix} - P_{IM2\_Mix}\Big|_{dBm}$$

and $P_{Tx\_Mix} \approx I_{in\_Mix} + 3dB$ crude !

$$P_{IM2\_Mix}\Big|_{dBm} = 2(P_{Tx\_Mix} - 3) - IIP_{2\_Mix} = -144\,\text{dBm} + G_{BM}$$

$$P_{IM2\_in}\Big|_{dBm} = P_{IM2\_Mix}\Big|_{dBm} - G_{BM} = -144\,\text{dBm}$$

To be compared with $N_{other}$= -100 dBm so less rejection for Tx would be sufficient ⇦



## Estimation of Tx leakage

additional calculations



**Contribution of Tx leakage to $N_{other}$**

$P_{Tx}$ = 27 dBm
$R_{Dupl}$ = -48 dB
$R_{BPF}$ = -30 dB
$G_{BM}$ = 15.. 25 dB
$N_{other}$ = -100 dBm
$P_{LO}$ = -3 dBm

**Path_2: (self mixing)**

$$P_{Tx\_Mix} = P_{Tx} + R_{Dupl} + G_{BM} + R_{BPF} = -51 + G_{BM}$$

$$P_{Tx\_LO} = P_{Tx} - ISL_{Tx\_LO}$$

Self mixing (referred to Mixer input )

$$P_{SM} \cong P_{Tx\_Mix} + (P_{Tx\_LO} - P_{LO}) = -51 + G_{BM} + 27 - ISL_{Tx\_LO} + 3 = -21\,\text{dBm} + G_{BM} - ISL_{Tx\_LO}$$

$P_{SM\_in} \cong P_{SM} - G_{BM}$ (referred to Rx input )

$$= -21\,\text{dBm} - ISL_{Tx\_LO} << N_{other}$$

$$P_{SM\_in} \cong N_{other} - 10\text{dB}$$

ISL needed: ⇦

$ISL_{Tx_LO} \cong 90$dB



## Estimation of Tx leakage

additional calculations



$ISL_{Tx\text{-}LO}$ needed is ~90dB, but

- PA is a separate chip – PCB board issue
- On-chip leakage reduced by ÷2 LO architecture
- All blocks differential - substrate coupling reduced

Other Tx leakage path are inferior for their lower signal powers www.liu.se

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