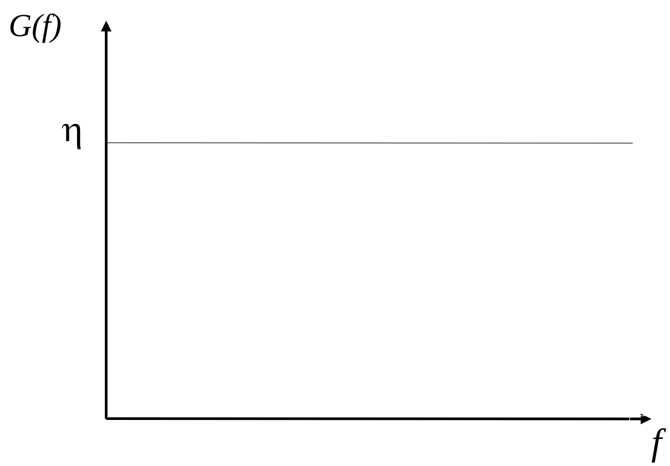
Main Sources of Electronic Noise

Thermal Noise

- It is always associated to dissipation phenomena produced by currents and voltages. It is represented by a voltage or current sources randomly variable in time.
- It is analytically described by a stationary process
- Amplitude distribution: GAUSSIAN with zero mean value
- Power Spectral Density: constant (white noise)



η = Power spectral density $= K \cdot T$

Main Sources of Electronic Noise

Shot noise

- It arises typically in *PN junctions* forwardly biased; it is due to the discrete nature of current through the junction, which results randomly variant around the imposed bias value
- Amplitude distribution: GAUSSIAN with zero mean value
- Constant spectrum (white noise)

η = Power spectral density ($=2q\cdot I$)

Main Sources of Electronic Noise

Flicker noise (1/f)

1. It arises in semiconductor devices, due to impurities and defects in the crystal structure).
2. Its spectrum is not constant (energy is concentrated at low frequency).

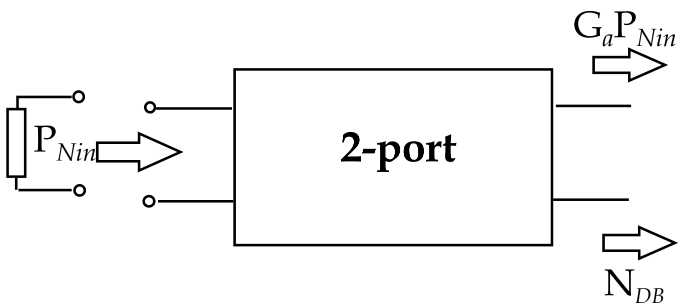
Power Spectrum:

$$G(f) = \bar{K}\frac{I^a}{f^b} \qquad W/Hz$$

K: depends on the fabrication process
I: DC current through the device
a: typically 0.5 ÷ 2
b: $\cong 1$

Nota: The amplitude distribution is not always GAUSSIAN

Noise Characterization in microwave devices



$\mathbf{N}_{DB}$ = Added Power (from the 2-port)

$\mathbf{P}_{Nout} = G_a P_{Nin} + N_{DB}$ = Total noise Power

The noise figure <u>*NF*</u>: define the attitude of the 2-port of adding noise at the output:

$$\boxed{NF = \frac{P_{Nout}}{G_a \cdot P_{Nin}}}$$

P_{Nout} is the actual noise power at output while $G_a \cdot P_{Nin}$ is the noise power at output if the 2-port would not add noise power (G_a is the available power)

Actually NF is a function of frequency, so the above powers must be assumed per unit band (i.e. they represent actually power densities). Moreover NF depends also on the source impedance Γ_S(being G_a depending on Γ_S)

NF dependance on Γ_S

$$NF = (NF)_{min} + 4r_n \frac{\left|\Gamma_s - \Gamma_{min}\right|^2}{\left|1+\Gamma_{min}\right|^2 \cdot \left(1-\left|\Gamma_s\right|^2\right)}$$

- ***(NF)$_{min}$*** = Minimum value of NF
- Γ_{min} value of Γ_S which determines NF=NF$_{min}$
- r_n = Normalized noise resistance

All these parameters are frequency dependent. Typically, they are made available by the manufacturers of commercial devices (directly into .s2p data files).

Constant Noise Figure Circles

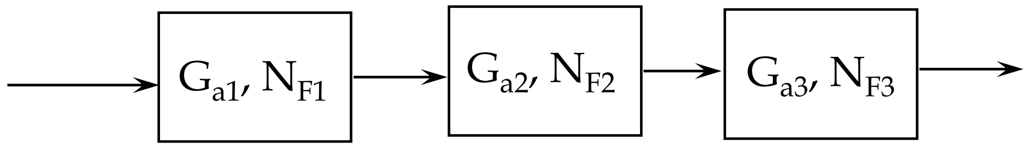
If we plot the equation expressing NF as function of Γ_S on the Smith Chart (representing Γ_S), we obtain a circle with the following center and radius:

$$C_F = \frac{\Gamma_{min}}{1+N_i}, \qquad r_F = \frac{1}{1+N_i}\sqrt{N_i^2 + N_i\left(1-\left|\Gamma_m\right|^2\right)}$$

N_i is given by:

$$N_i = \frac{NF-(NF)_{min}}{4r_n}\cdot\left(\left|1+\Gamma_{min}\right|^2\right)$$

Noise Figure for cascaded stages



$$\left(NF\right)_{TOT} = NF_1 + \frac{NF_2 - 1}{G_{a1}} + \frac{NF_3 - 1}{G_{a1}G_{a2}} + \cdots$$

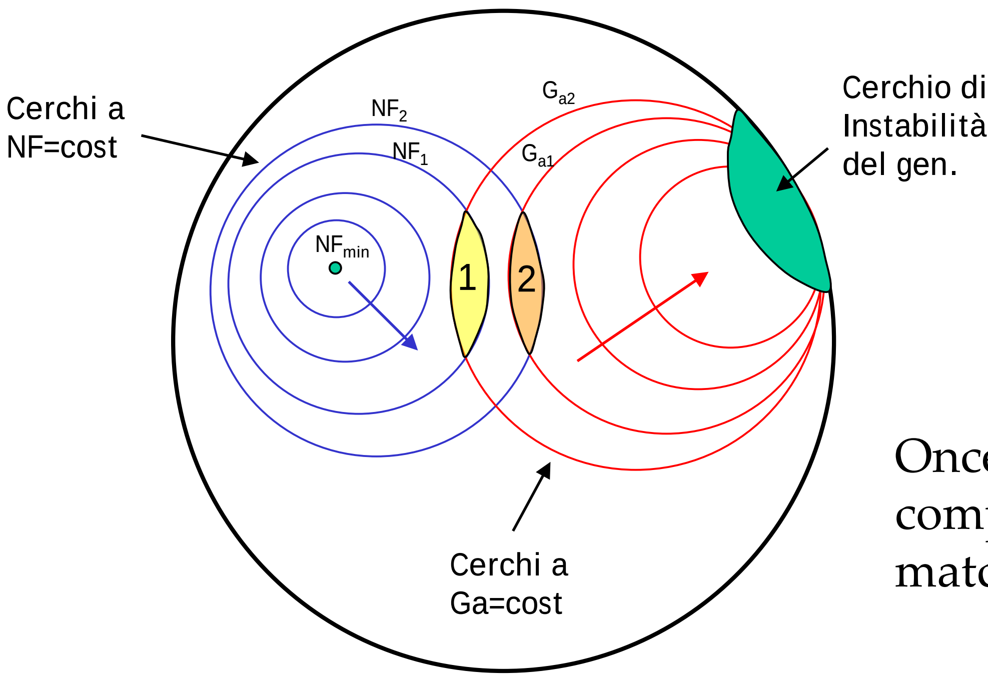
The noise figure is mainly determined by the first stage

NOTE: In general the value of Γ_S *that* determines the minimum value of NF is different by the one that maximizes G_T; the choice of Γ_S for the first stage is then the result of a compromise between the *noise figure* and gain

Design of a low noise amplifier

The choice of Γ_s is a compromise between G_T and NF.

Some circles with NF=cost and G_a=cost are first plotted on the Smith chart (Γ_s). The value of Γ_s is selected within the common area of two circles. Considering that NF increases with the radius while Ga decreases with it, we have (with reference to the zones 1 and 2):



In zone **1:** NF≤ NF_1 and $G_a \geq G_{a2}$
NF is previliged

In zone **2:** NF≤ NF_2 and $G_a \geq G_{a1}$
G_a is previliged

Once assigned $\Gamma_{s,opt}$, $\Gamma_{L,opt}$ is computed by imposing the matching at output (then G_T=G_a)

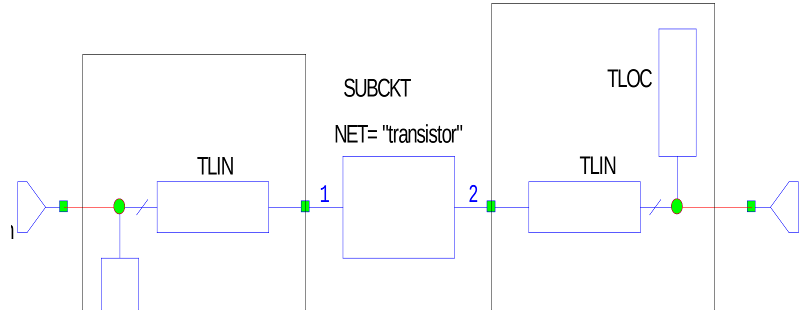
Example of design

Amplifier Requirements
Frequency Band: 6.2 – 6.8 GHz
Minimum Transducer Gain: 10.5 dB
Maximum Noise Figure: 1.5 dB

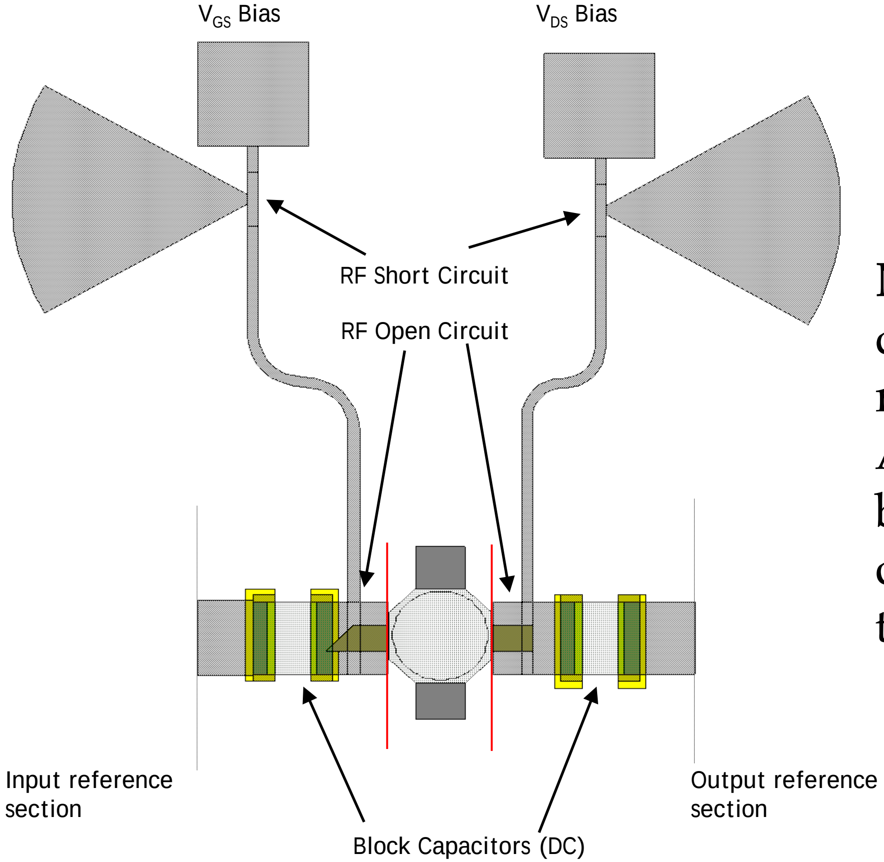
Substate: Duroid
$\varepsilon_r = 2.54$
H= 0.508 mm
t = 35 μ

Active Device
MGF1923 Mitsubishi (GaAs Mesfet)
MSG (6.8 GHz): 15.16 dB (with NF=3.1 dB)
Minimum NF (6.8 GHz): 1.13 dB (with Gt=8.06 dB)

Topology:

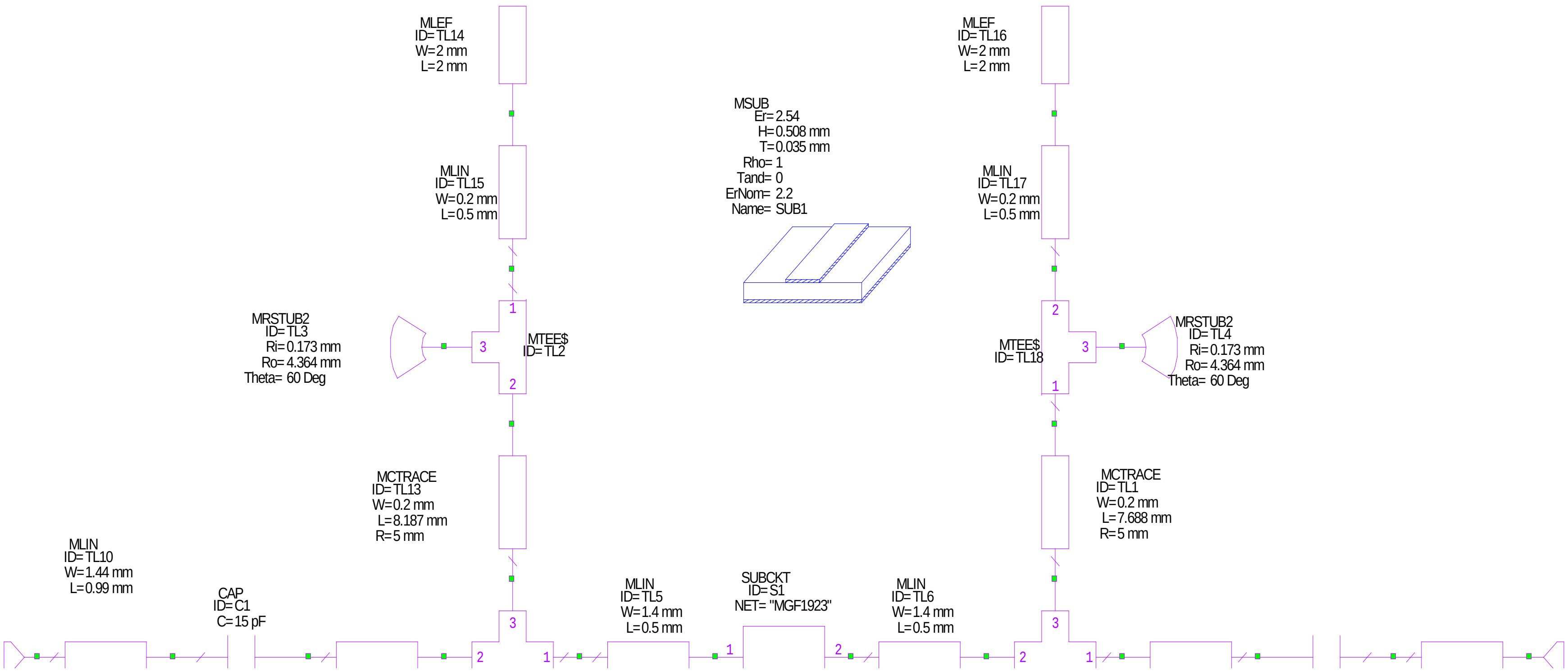


Biasing network of the active device



NOTE: The S parameters delivered by the manufacturer refers to the red sections.
After the biasing network has been assigned, the S parameters changes to the ones referred to the black sections

Evaluation of the S parameters of the biased active device



S parameters for the design

Frequency: 6.8 GHz

S11 (Mag, Phase deg) 0.80304 , -164.13

S12 (Mag, Phase deg) 0.07083 , -32.616

S21 (dB, Phase deg) 7.3143 , 23.642

S22 (Mag, Phase deg) 0.58075 , -130.99

Potentially INSTABLE

Maximum Stable Gain (dB): 15.155

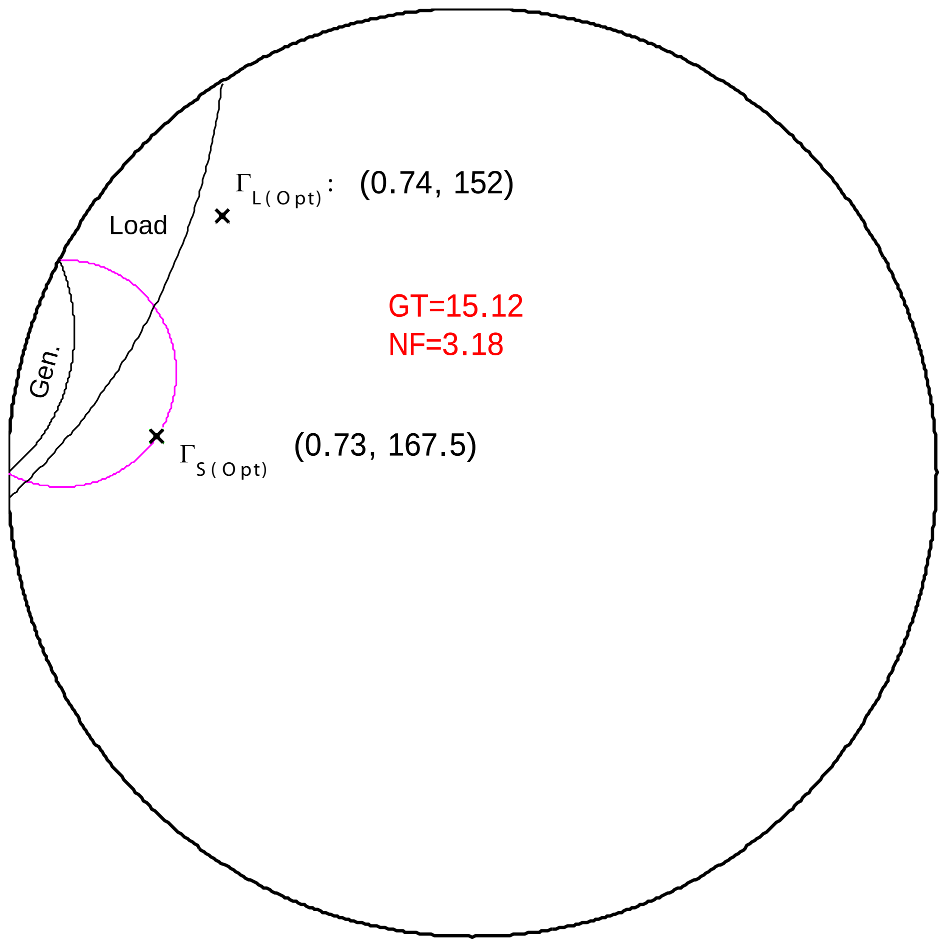
Stability Coefficient K: 0.66826

Minimum Noise Figure (dB): 1.1651

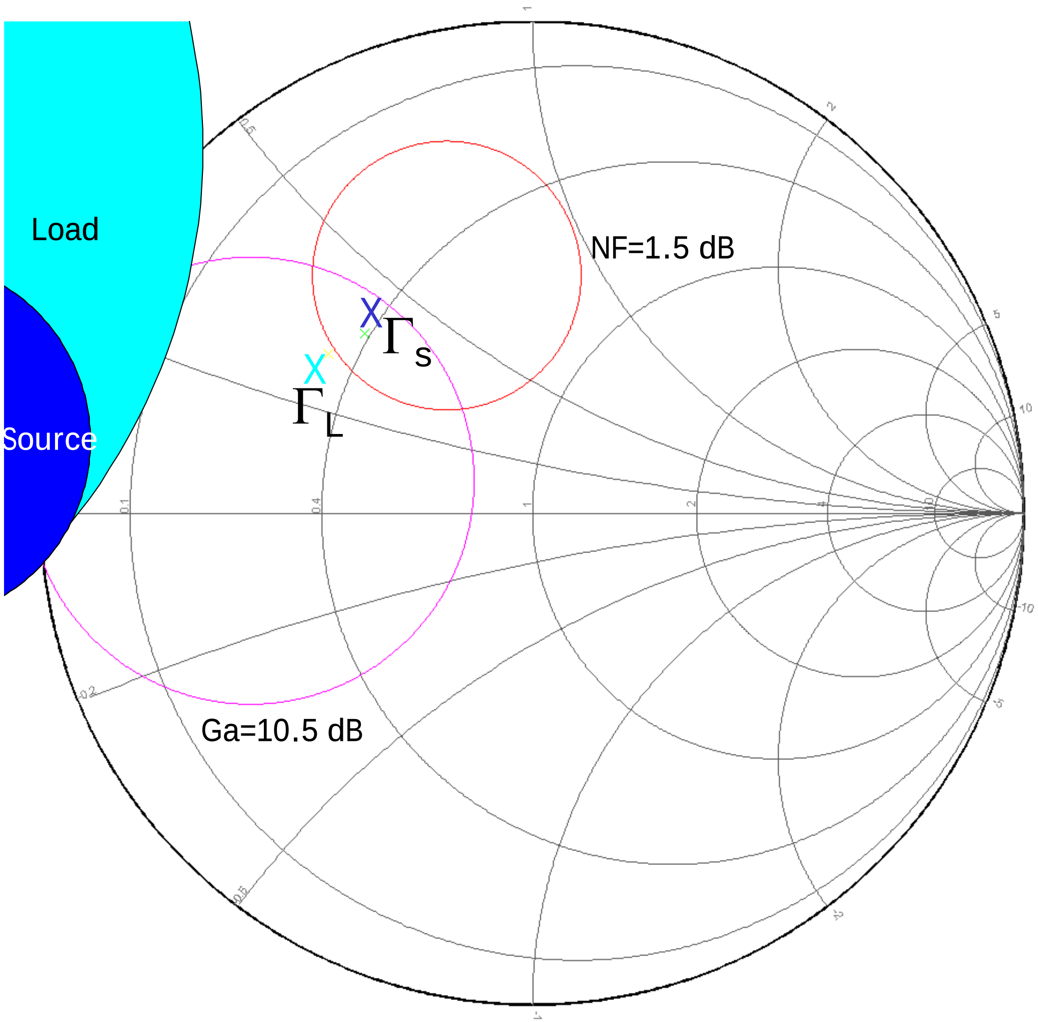
Optimum Gamma Source (Mag, Phase deg): 0.57413 , 109.84

Normalized Noise Resistance: 0.20827

Γ_S selected for maximum gain



Selection of Γ_S as a compromise between G_T and NF



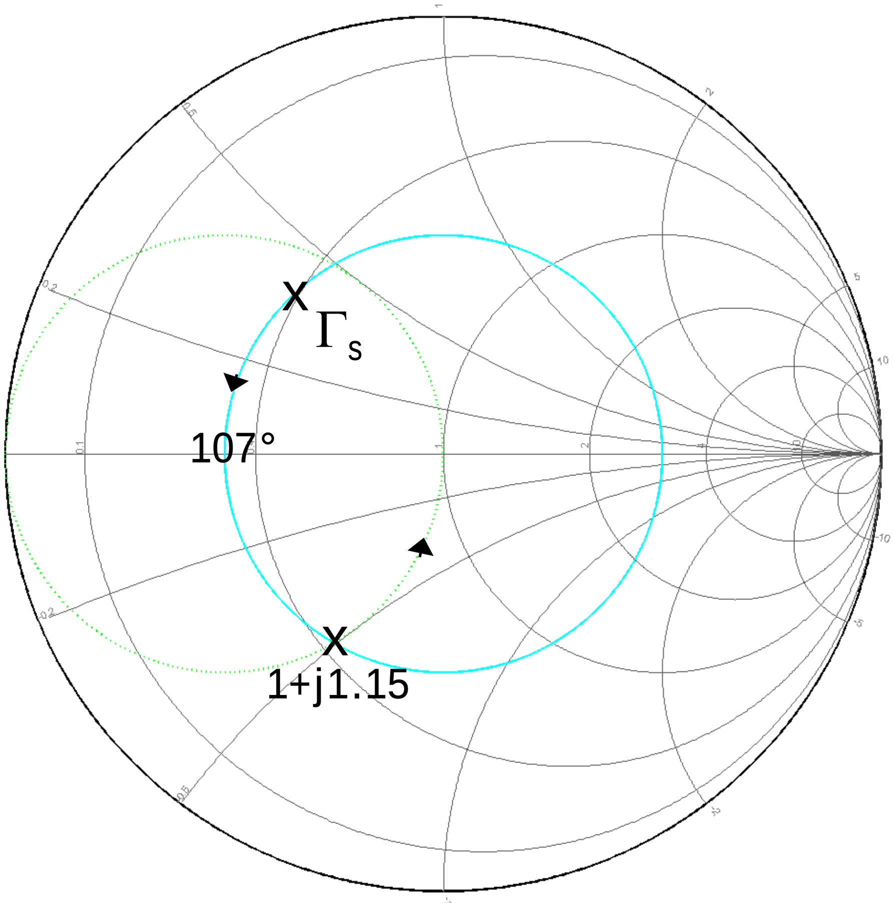
$G_T = 10.7$ dB

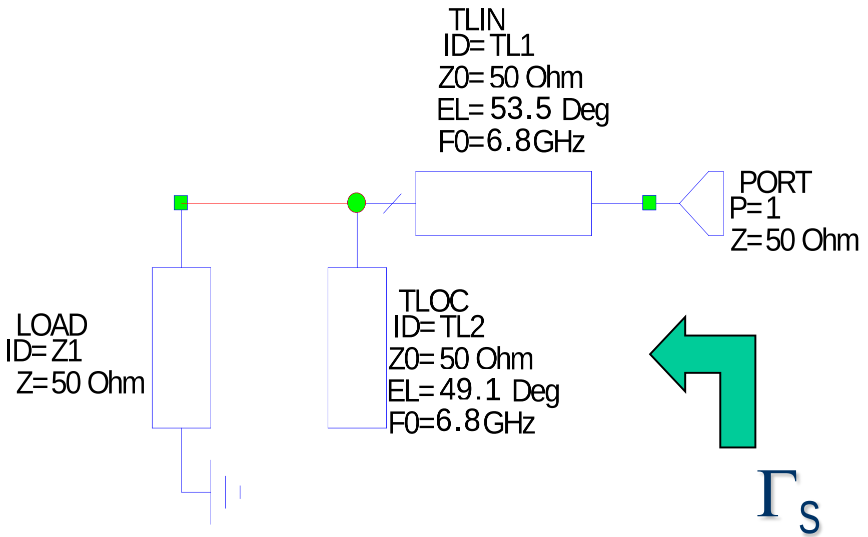
NF=1.36 dB

$\Gamma_L = 0.524\angle 142.1°$

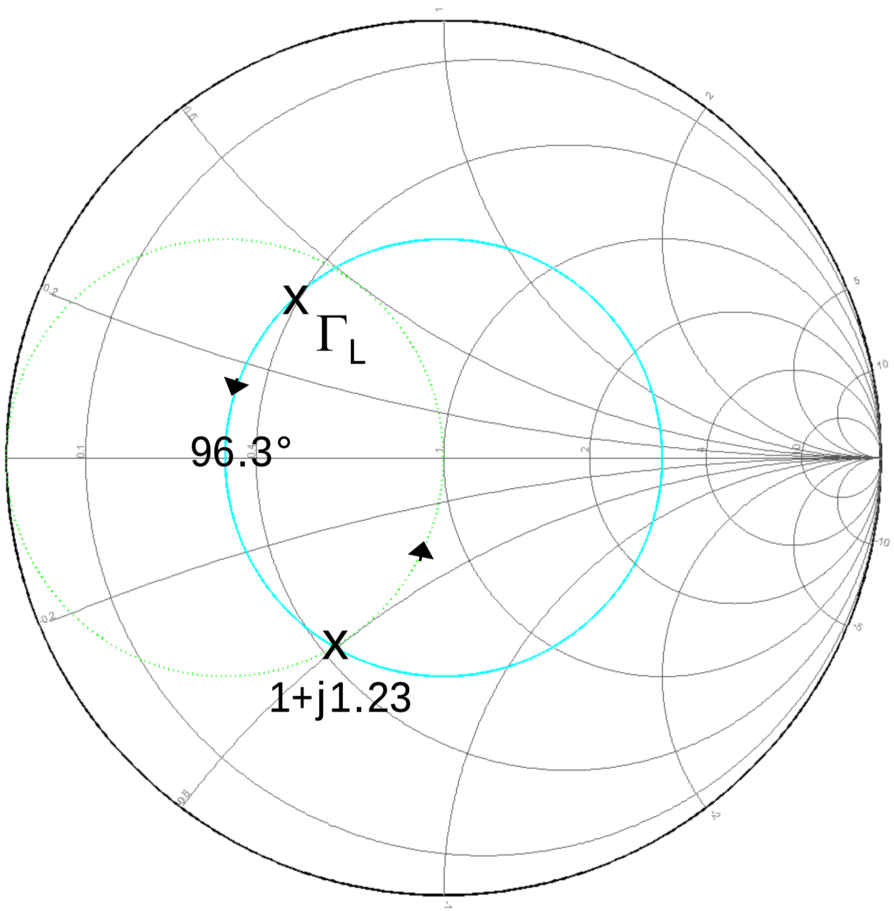
$\Gamma_S = 0.5 \angle 133°$

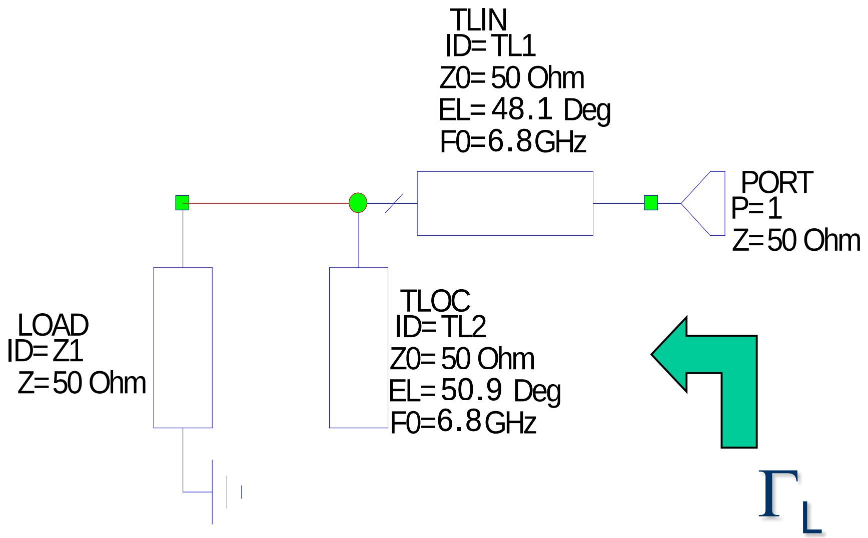
Input matching network



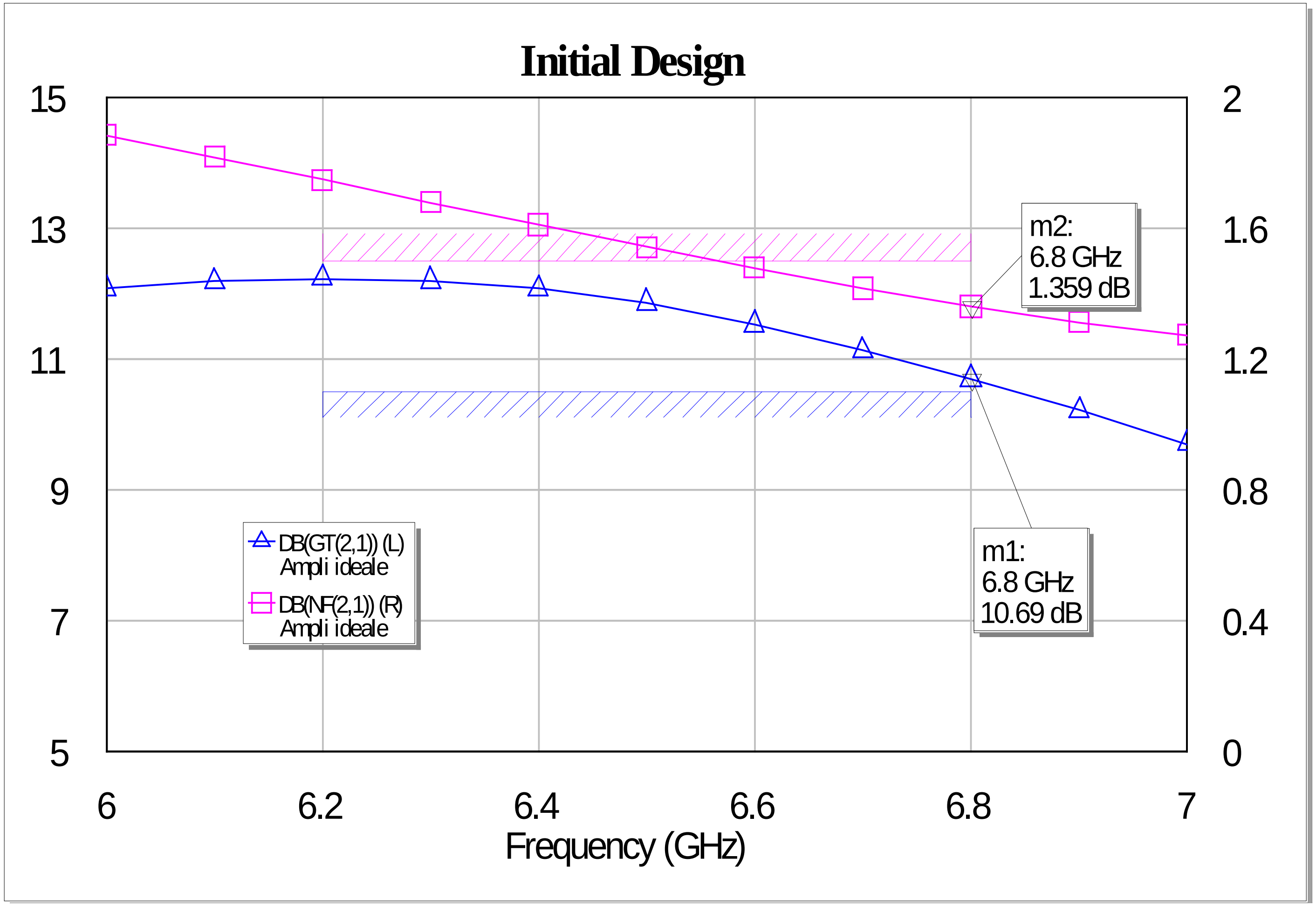


Output matching network

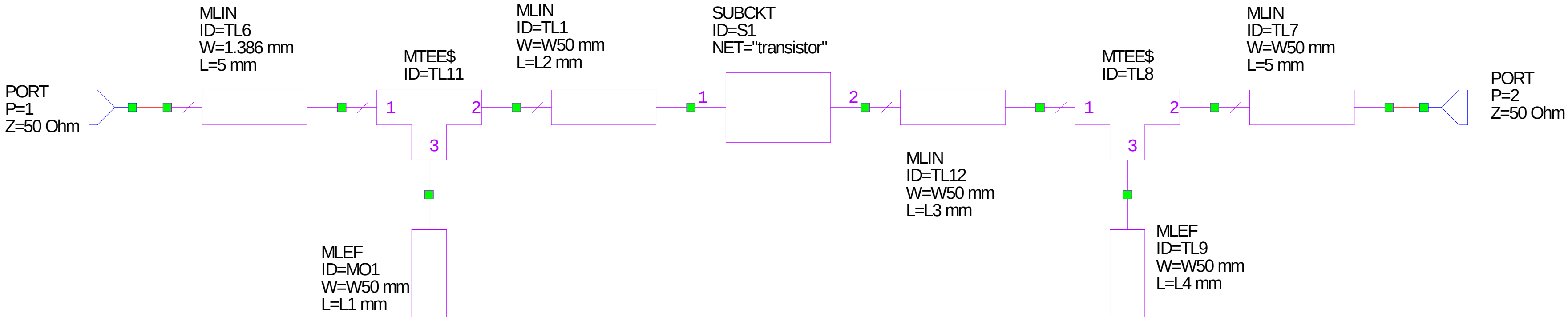




Initial design response (ideal lines)

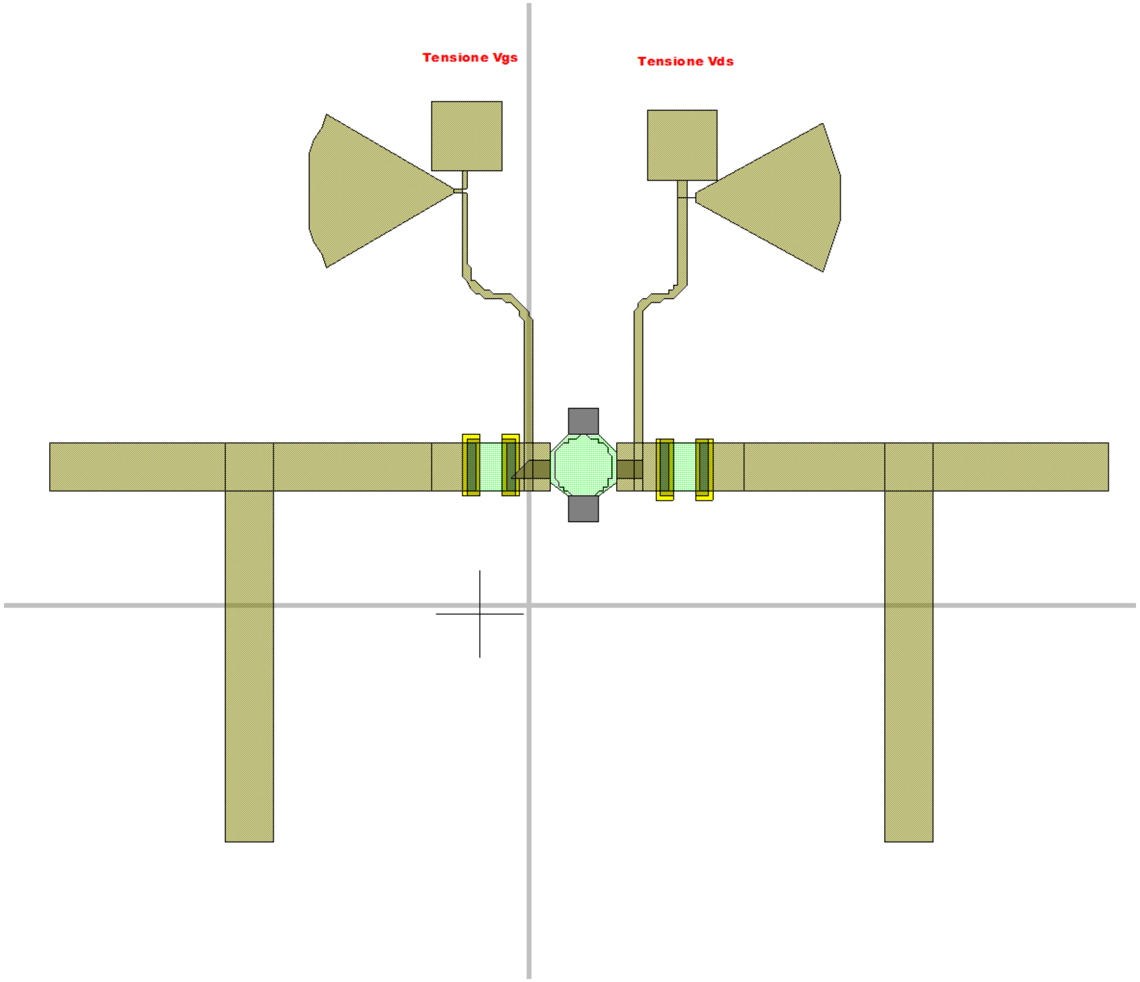


Microstrip implementation (including discontinuities)

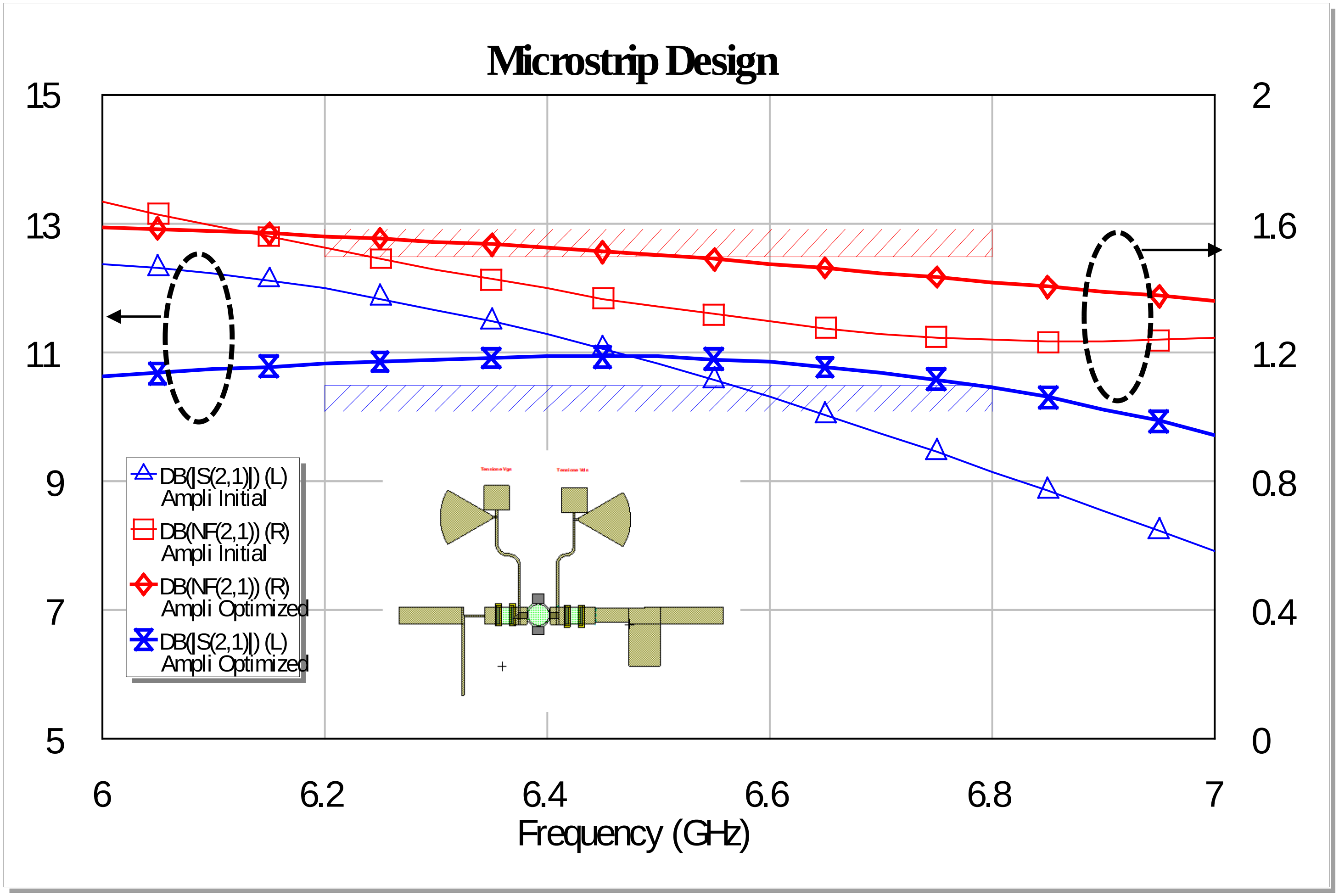


W50=1.38645
L1=4.13952
L2=4.51048
L3=4.05522
L4=4.29128

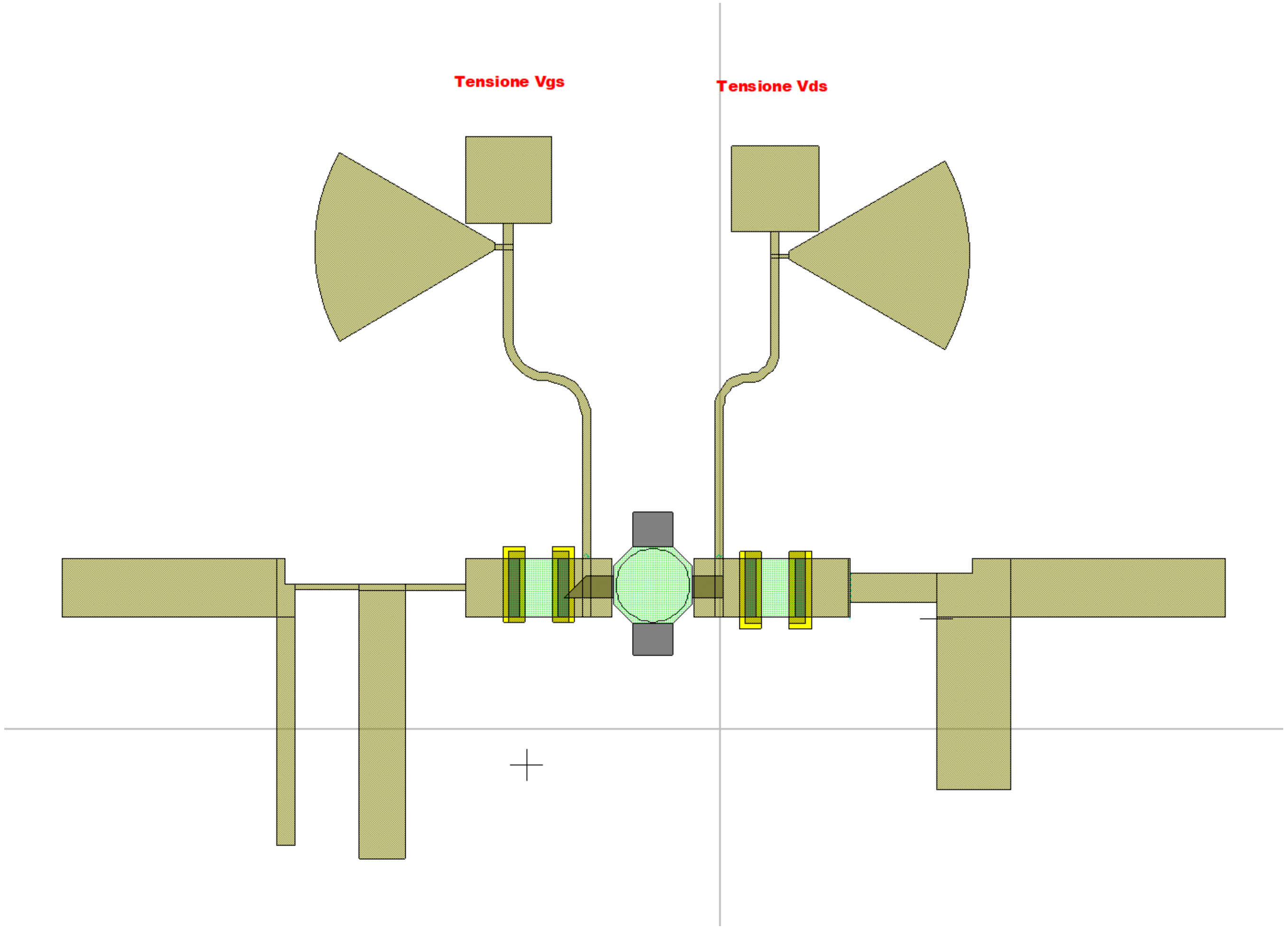
Amplifier Layout (initial)



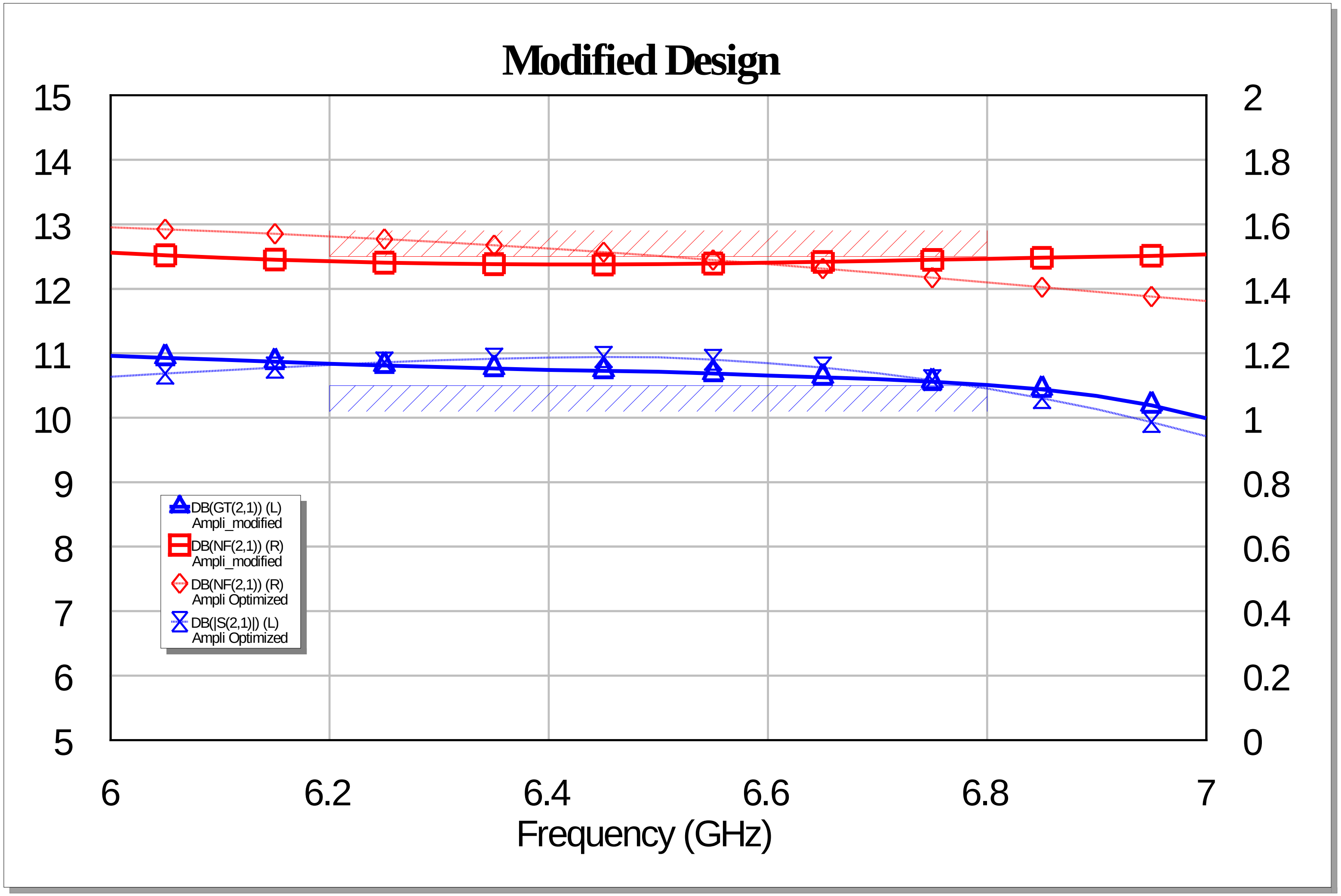
Amplifier Response (initial/optimized)



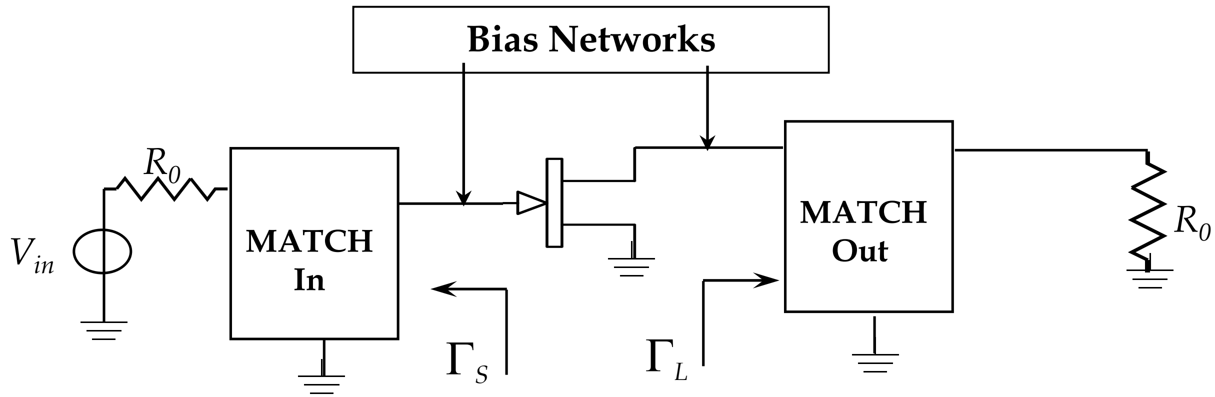
Alternative Scheme (optimized)



Final response



Scheme of a power microwave amplifier



The concept is identical to the ones seen before. In this case however the main design goals concern the output power, linearity and efficiency. The values of Γ_S and Γ_L should determine the best compromise among the main goals (but taking also into account the gain and the matching conditions).

Instability issues (also potential) have to be absolutely avoided (commercial devices are generally pre-matched internally for unconditional stability).

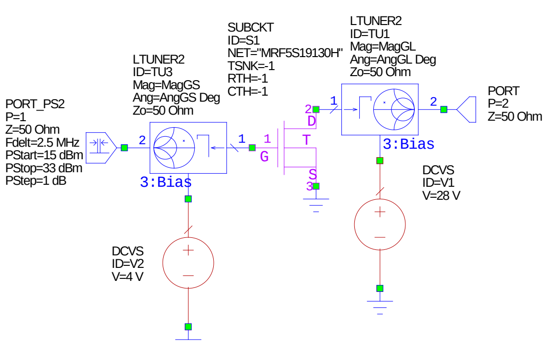
The active device must be characterized for large signal operation

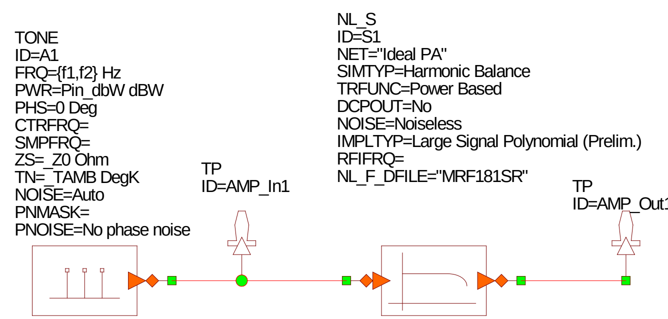
Device characterization

- The device model should allow the evaluation of absolute quantities (Pout, Intermodulation, Efficiency), to be used in non-linear Harmonic Balance simulators.
- Unfortunately non-linear circuit models of commercial devices are rarely available
- As a much less accurate alternative, behavioral models based on the information delivered by the manufacturers can be employed also in circuit simulations
- Such models are memoryless and usually adopt a polynomial model together with the saturated Pout. They are defined by P_{1dB} and IP_3, which however are not related by the simple relation valid for 3th order nonlinearity.

Selection of the active device

- In this phase the behavioral models are most suited
- If the requirements concern modulated (RF) signals, it could be convenient to use system simulators
- Some system simulators (e.g. VSS), other than models of PA based on polynomial characterization, also include models based on concurrent circuit simulation (Harmonic Balance). However, also these models adopt a polynomial characterization derived from the circuit response

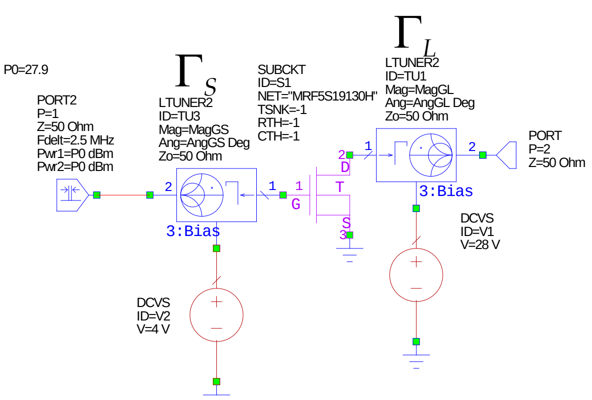


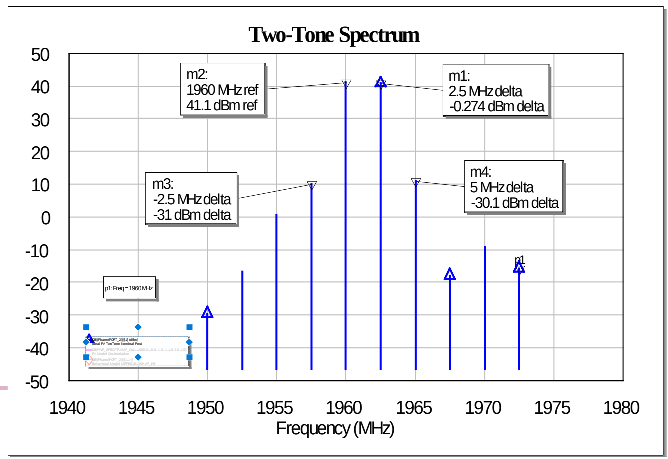


Amplifier Design From Microwave Office

Optimum Γ_S and Γ_L

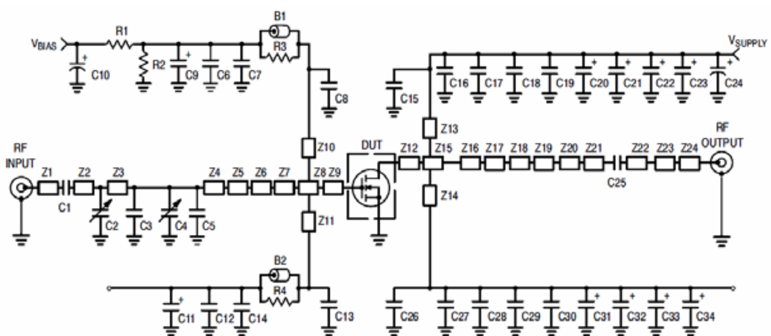
- These parameters are generally specified by the manufactured in the typical operating conditions (exciting signal and output average power)
- If a non-linear model of the device is available, the optimum values can be searched through non-linear simulations (starting with those suggested in the device data sheet)
- In most cases, the topology of the matching networks must be those provided by the manufactured (especially when the optimum impedances are very low)

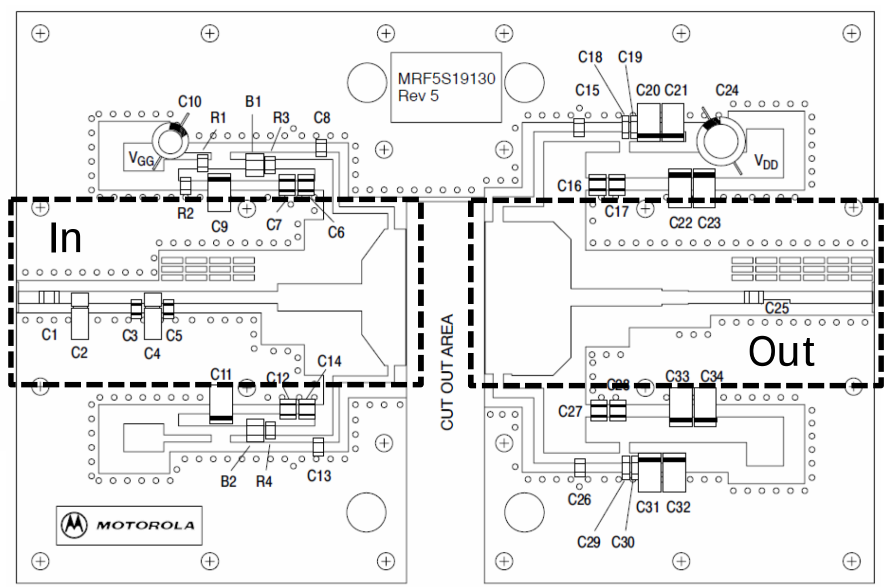




Matching Networks Layout

- The most suited topology is generally suggested by the manufacturer (also the biasing circuits are specified).
- If a non-linear model is available, a fine tuning of the network dimensions can be carried out





Design of a “line –up” for N-CDMA

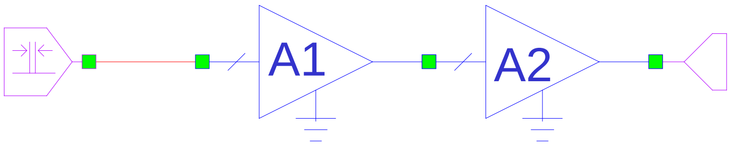
General Specifications

- Center Frequency: 1960 MHz,   Band : 1930-1990 MHz
- Channel band: 1.2288 MHz (IS-95) Channels spacing: 2.5 MHz
- Output power: ≥ 100 W PEP (2-tone)
- Gain: ≥ 27 dB (max input power 200 mW PEP (23 dBm))
- Linearity: C/I ≥ 30 dB (2-tone)

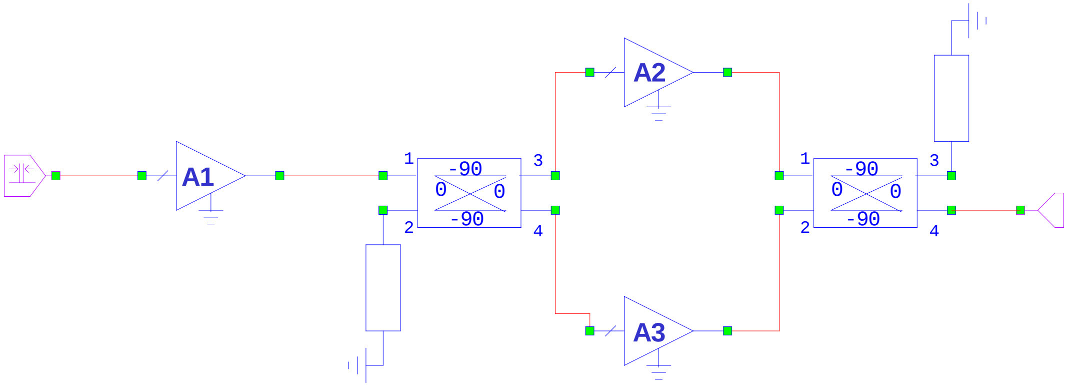
PAE > 25% with 2-tone at rated PEP

Possible topologies

- Due cascaded stages



- Single stage followed by a balanced pair



Devices choice

Manufacturer: <u>**Freescale**</u>

<u>Final stage</u>

MRF5S19130 (P1dB=125W, Vdd=28V, Gt=13 dB, η=33%) → Topology 1

MRF7S19100 (P1dB=100W, Vdd=28V, Gt=17.5 dB, η=30%) → Topology 1

MRF6S19060 (P1dB=60W, Vdd=28V, Gt=16 dB, η=35%, IM3=-35 dBc) → Topology 2

MRF19045 (P1dB=45W, Vdd=26V, Gt=14.5 dB, η=36%, IM3=-30 dBc) → Topology 2

<u>Driver</u>

MRF6S20010 (P1dB=20W, Vdd=28V, Gt=16 dB, η=41%, IMD=33 dBc) → Topology 1/2

MRF282 (P1dB=10W, Vdd=26V, Gt=12 dB, η=33%, IMD=31 dBc) → Topology 1/2

2 cascaded stages

Chosen final device: MRF5S19130 (IP3=61.5 dBm, G_{final}=13 dB).
Evaluation of IP3 of the driver (imposing the overall CI3):

$$P_{\omega 1} = PEP - 6\text{ dB} = 44\text{ dBm}$$

$$IP_{3,tot} = \frac{CI + 2P_{\omega 1}}{2} = 59\text{ dBm}$$

Sum in power of distortion

$$IP_{3,tot} = IP_{3,final} - 10\log\sqrt{\left(1 + 10^{(IP_{3,final} - G_{T,final} - IP_{3,driver})/5}\right)}$$

$$IP_{3,driver} = IP_{3,final} - G_{T,final} - 5\log\left(10^{\left(IP_{3,final} - IP_{3,tot}\right)/5} - 1\right)$$

→ $IP_{3,driver}$ = 46.82 dBm

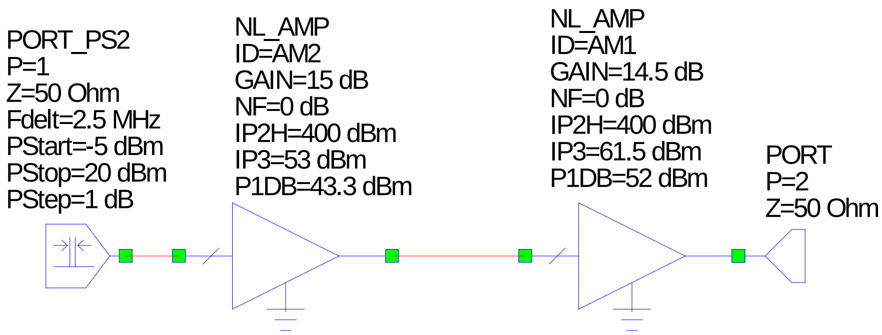
Sum in voltage of distortion

$$IP_{3,tot} = IP_{3,final} - 10\log\left(1 + 10^{(IP_{3,final} - G_{T,final} - IP_{3,driver})/10}\right)$$

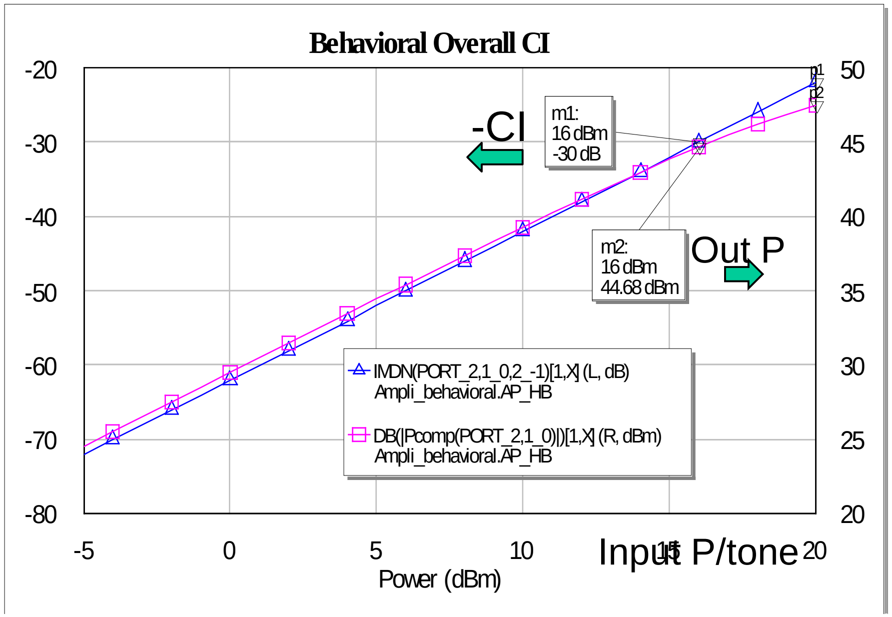
$$IP_{3,driver} = IP_{3,final} - G_{T,final} - 10\log\left(10^{\left(IP_{3,final} - IP_{3,tot}\right)/10} - 1\right)$$

→ $IP_{3,driver}$ = 49.58 dBm

Verification with behavioral models

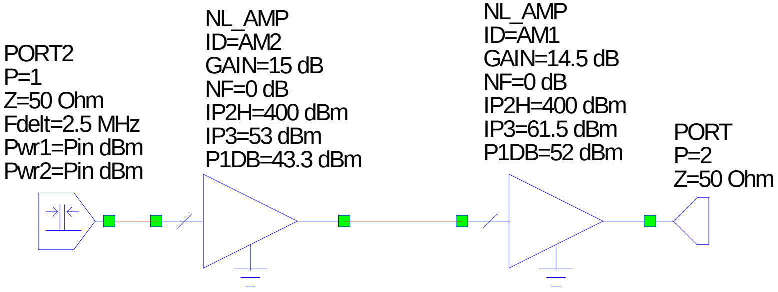


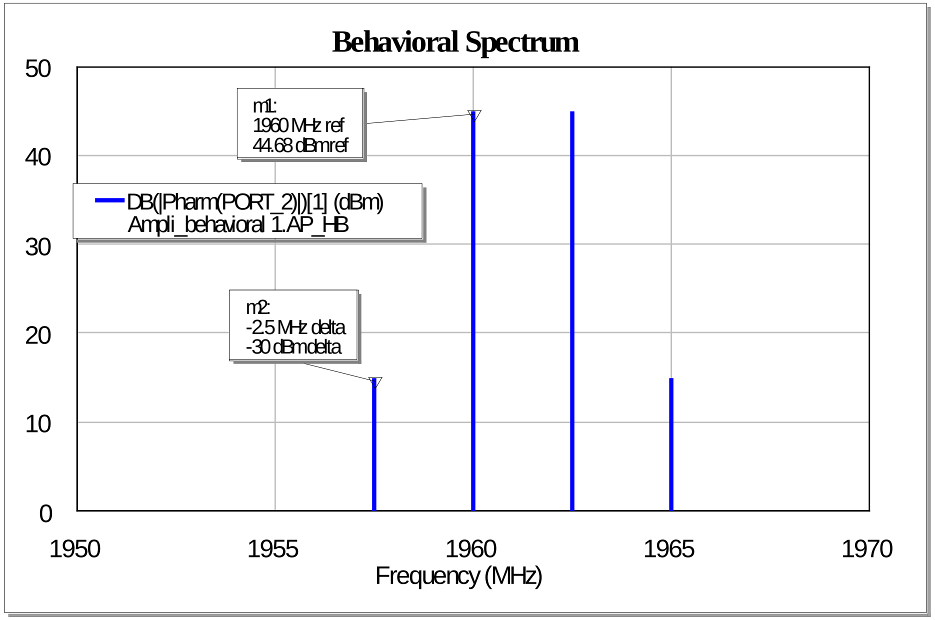
Requested outP/tone: 44 dBm



Chosen device for the driver:
MRF6S20010 (**IP3=53 dBm**)

Spectrum evaluation for CI=30 dBm

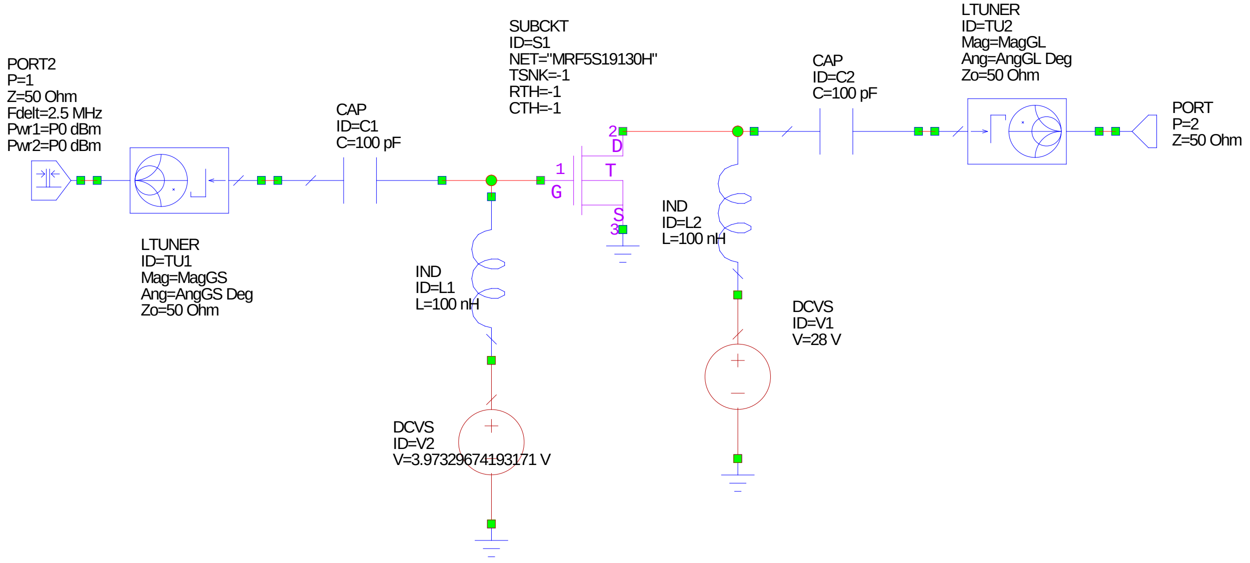




Potenza out: 116.95 W (PEP)
C/I: 30 dBc
Gt=28.68 dB (Pin/tone=16 dBm)

Evaluation of the optimum loads

- Starting points: optimum impedances reported on datasheets).
- Topology of the networks suggested by the manufacturer
- Biasing point reported on datasheets for optimum performances
- Tuning of the networks for maximize Pout e C/I

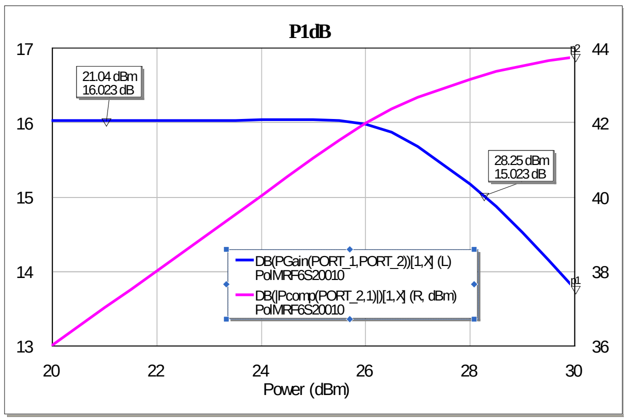


Result of the simulations (Harmonic balance)

Driver: MRF6S20010

Bias: Vdd=28, Id=130 mA

$Zs=9.52+j2.14 \quad Z_L=2.75+j3.67$



P1dB,driver=43.3 dBm

IP3,driver=50.5 dBm

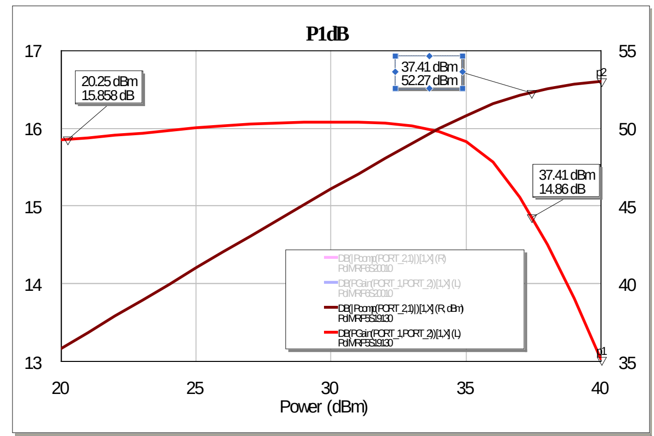
CI=34.4 (PEP=39.44 dBm)

G=17.44 dB

Final: MRF5S19130

Bias: Vdd=28, Id=1200 mA

Zs=2.35 - j7.6 ZL=1.28 - j1.5



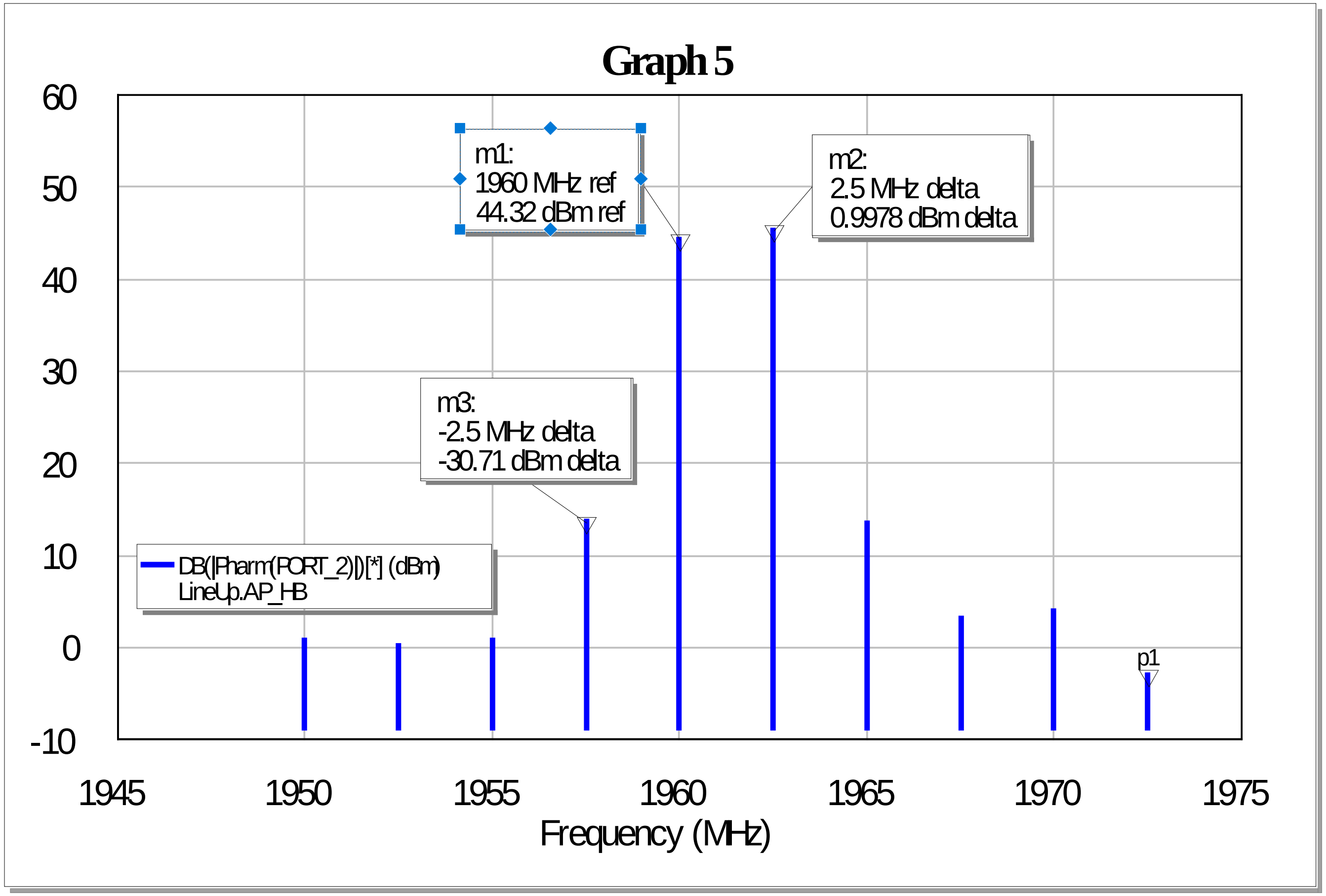
P1dB,final=52.3 dBm

IP3,final=60.4 dBm

CI=32. (PEP=50.33 dBm)

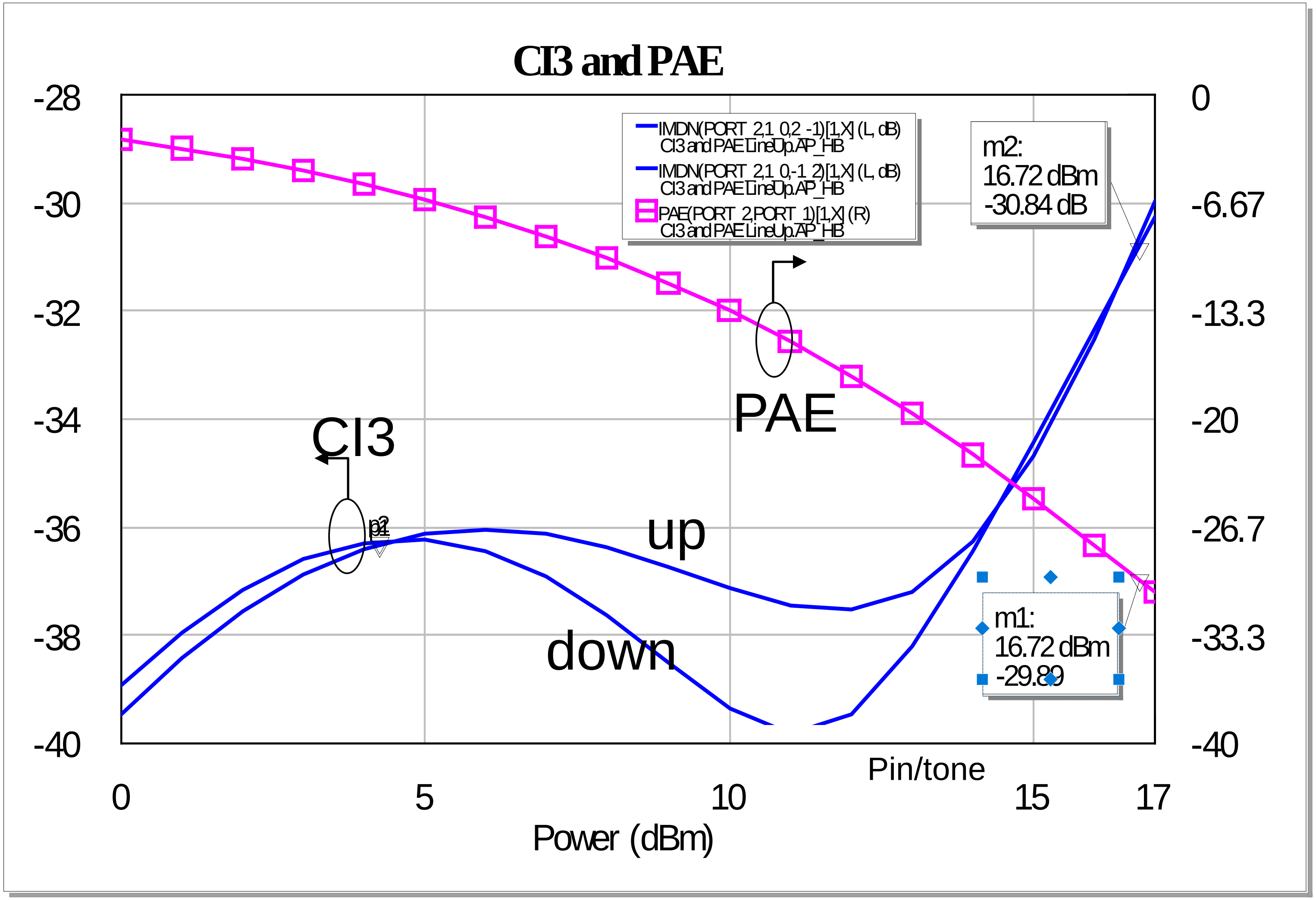
G=12.33 dB

Overall line-up: Spectrum for Pout max



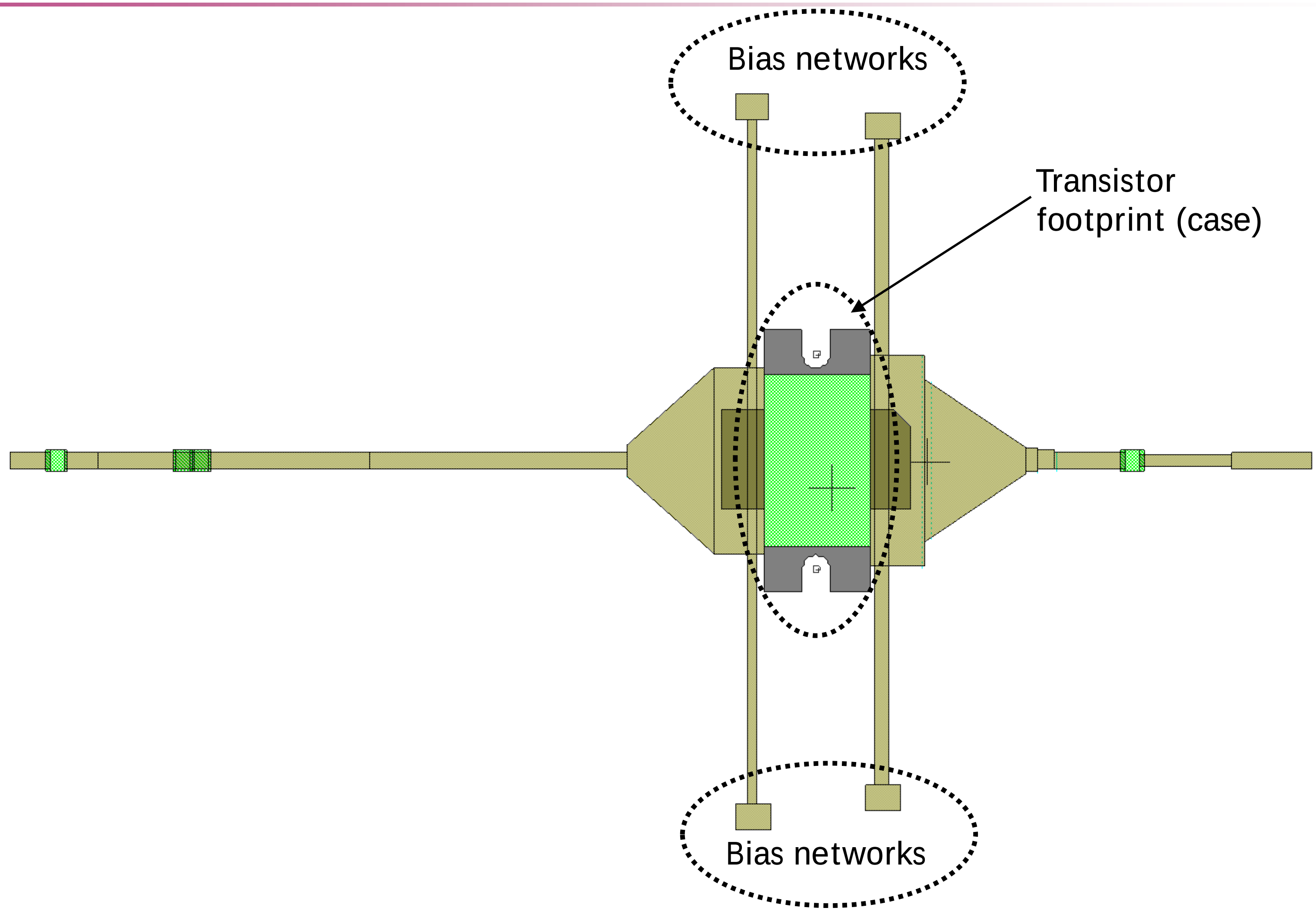
Pout: 107.9 W (PEP), C/I: 30.7 dBc Gt=27.6 dB (Pin=187 mW PEP)

Overall line-up : CI3 e PAE vs Pin



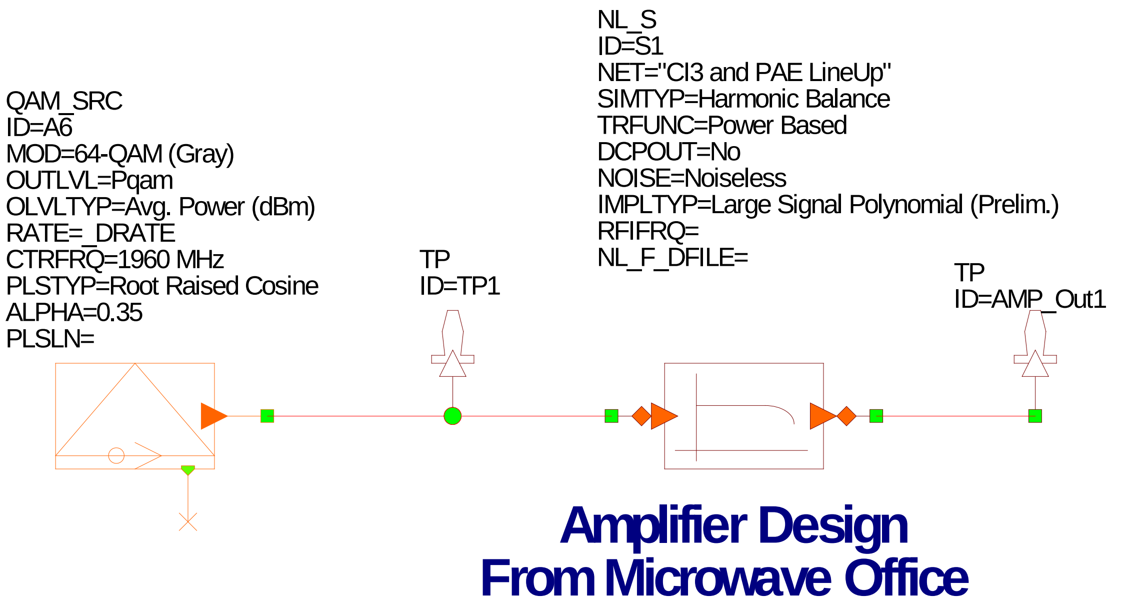
Asymmetry CI3 → Memory effects

Actual (possible) layout of final PA

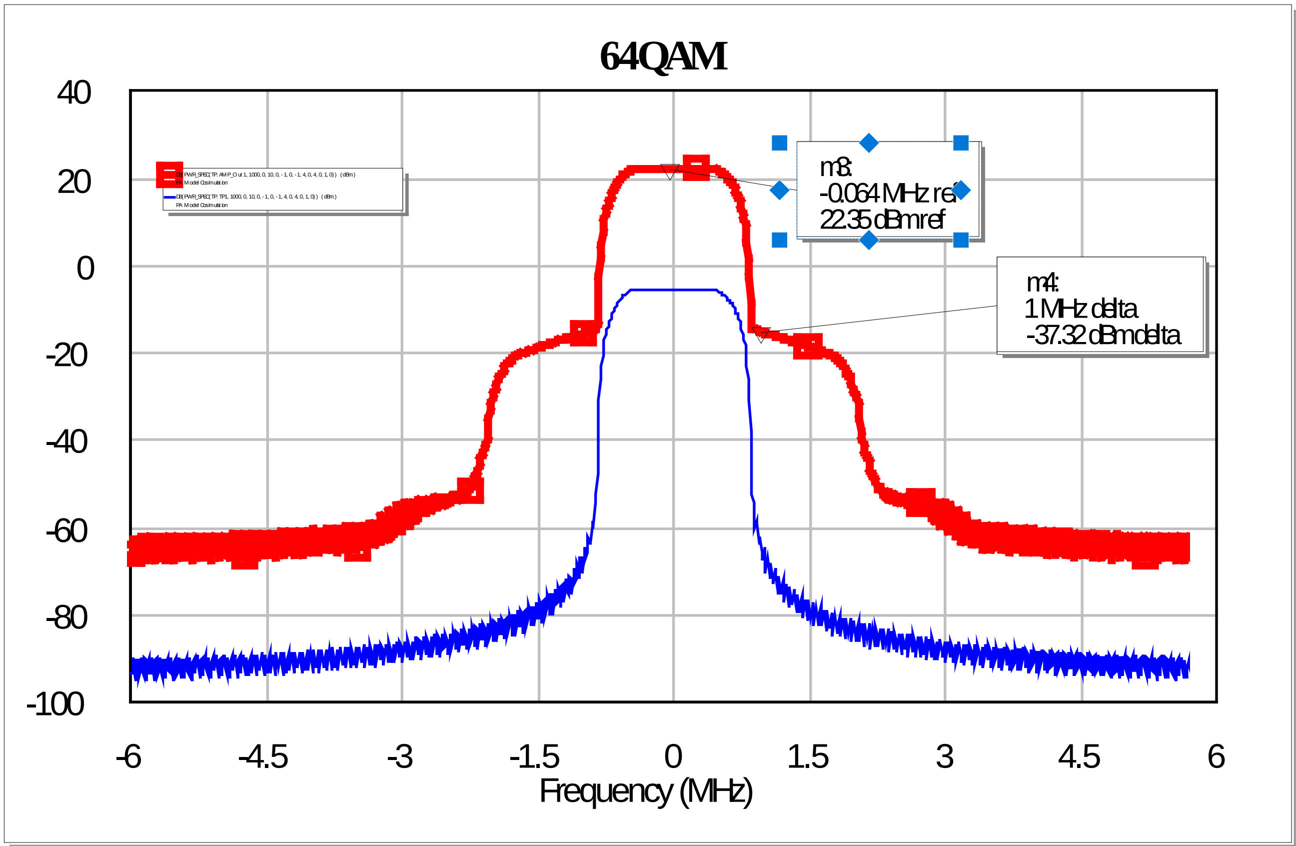


Co-simulation with VSS

- To obtain the estimated response with a RF signal (e.g. 64-QAM) we can use VSS with the block relating to circuit simulation:

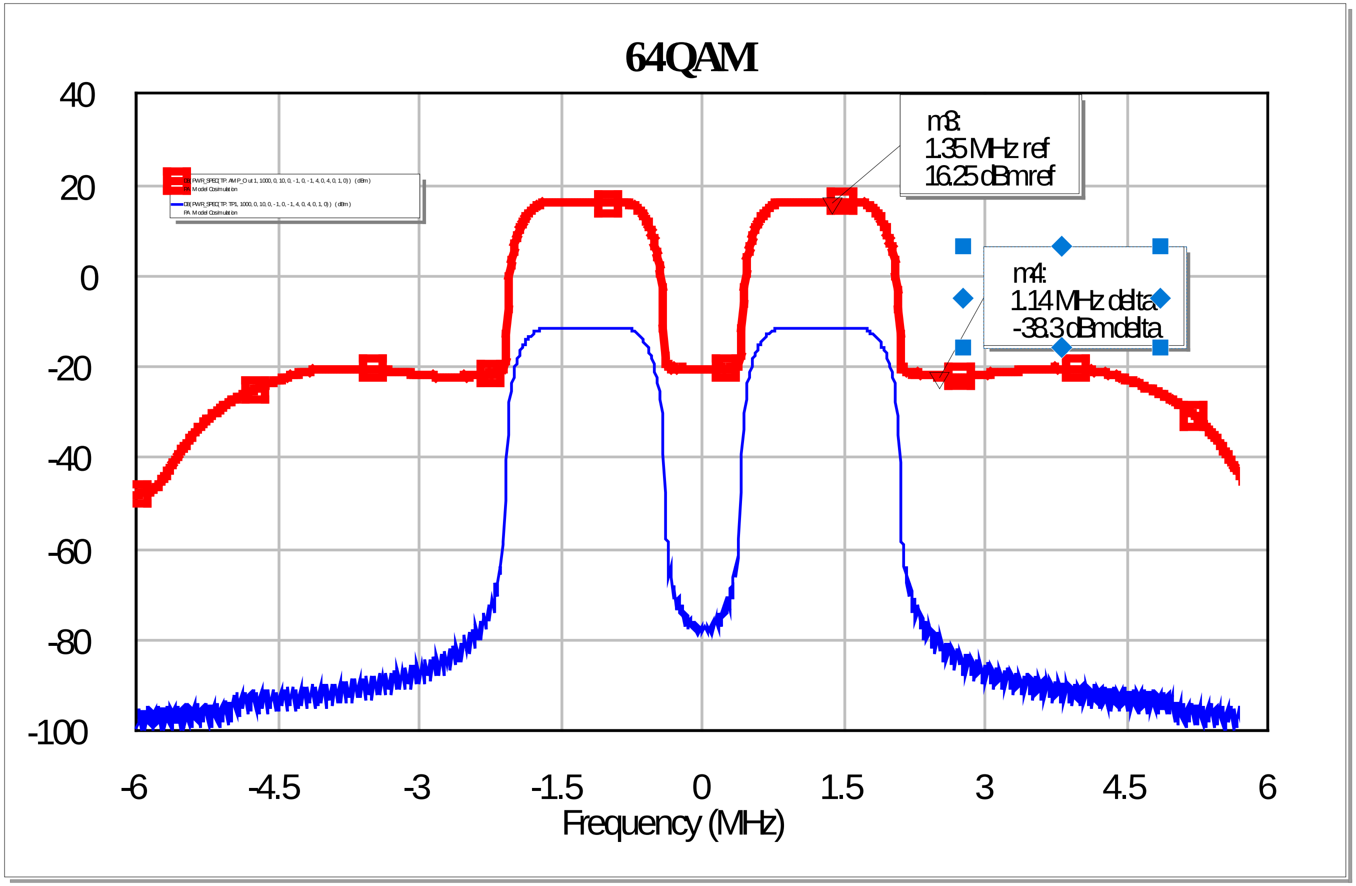


Input-output spectra (64-QAM)



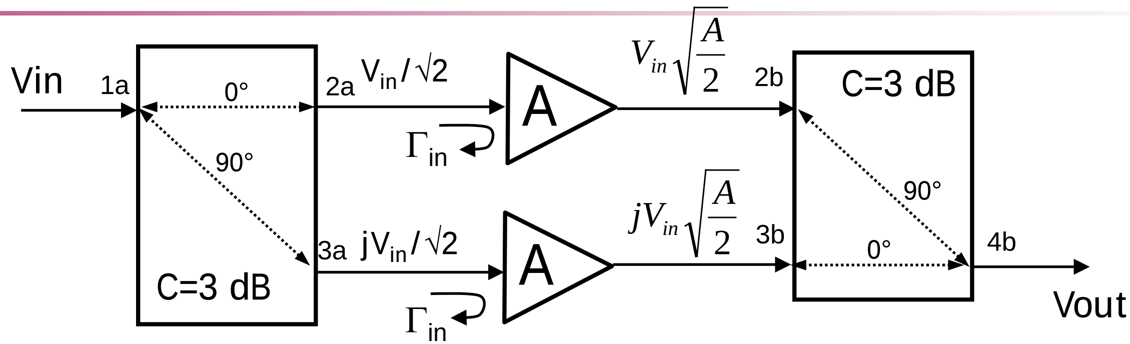
Output power: 40 dBm (average)

Two Modulated carriers (64-QAM)



Pout av. : 37 dBm (C/I=38 dB)

Balanced Amplifier



Gain:

$$V_{2b} = \sqrt{A}\cdot V_{in}\big/\sqrt{2}, \quad V_{3b} = j\sqrt{A}\cdot V_{in}\big/\sqrt{2}$$

$$V_{out} = V_{4b} = j\frac{V_{2b}}{\sqrt{2}} + \frac{V_{3b}}{\sqrt{2}} = j\sqrt{A}\cdot V_{in} \quad \Rightarrow \quad \frac{P_{out}}{P_{in}} = A$$

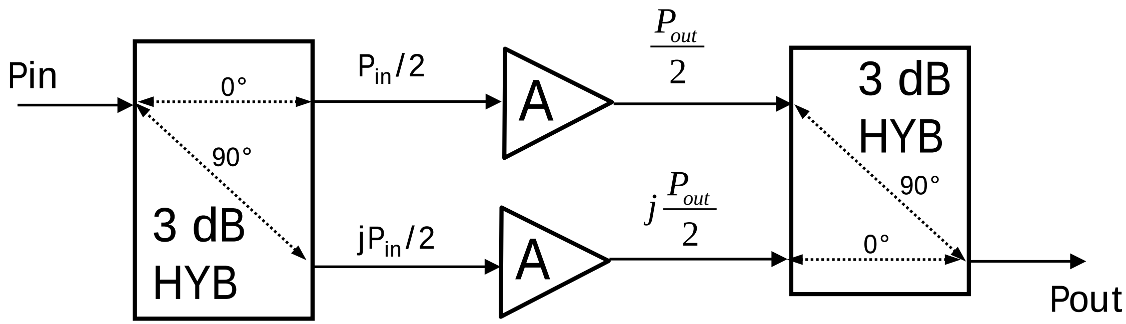
Reflection:

$$V_{1a}^{+} = V_{in}, \quad V_{2a}^{-} = \sqrt{A}\cdot V_{in}\big/\sqrt{2}, V_{2a}^{+} = \Gamma_{in}\sqrt{A}\cdot V_{in}\big/\sqrt{2}, \quad V_{3a}^{-} = j\sqrt{A}\cdot V_{in}\big/\sqrt{2},$$

$$V_{3a}^{+} = j\Gamma_{in}\sqrt{A}\cdot V_{in}\big/\sqrt{2}, \quad V_{1a}^{-} = \frac{1}{\sqrt{2}}\left(jV_{3a}^{+} + V_{2a}^{+}\right) = -\Gamma_{in}\sqrt{A}\cdot V_{in}\big/2 + \Gamma_{in}\sqrt{A}\cdot V_{in}\big/2$$

$$\Gamma_{in} = \frac{V_{1a}^{-}}{V_{1a}^{+}} = 0$$

IP3 in balanced amplifiers



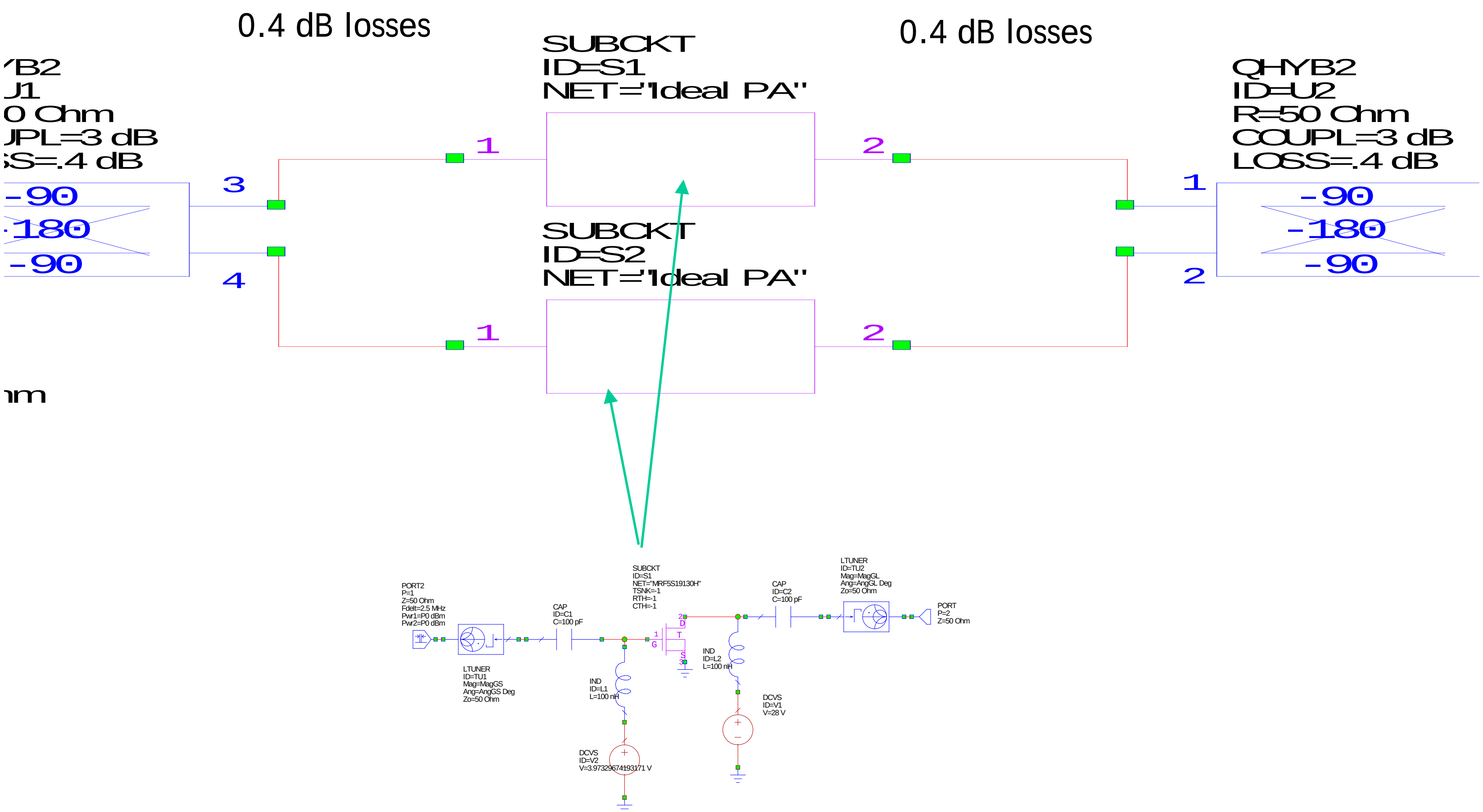
$$P_{\text{int},1} = 3(P_{out} - 3) - 2IP_3, \qquad P_{\text{int},2} = \left[3(P_{out} - 3) - 2IP_3\right]$$

$$P_{\text{int}} = 3(P_{out} - 3) - 2IP_3 + 3 = 3P_{out} - 2IP_3 - 6 = 3P_{out} - 2IP_3'$$

$$IP_3' = IP_3 + 3$$

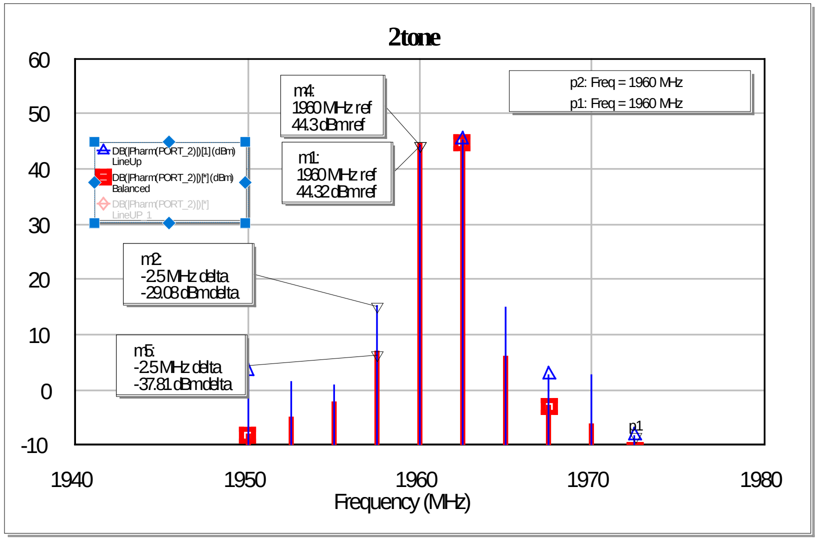
<u>Result</u>: the equivalent IP3 of the overall amplifier is doubled with respect the one of the single amplifiers. This means that for the same overall output power the power of the intermodulation products is 6 dB lower.

Balanced PA with two MRF5S19130

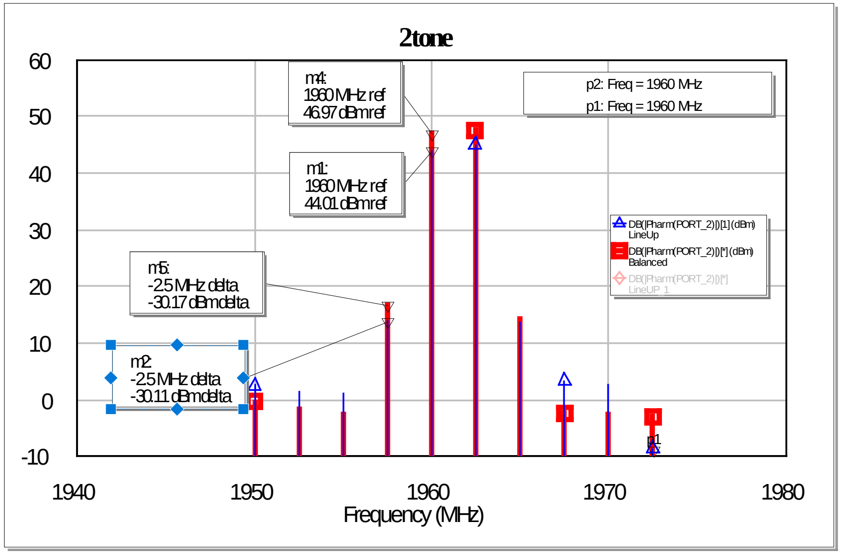


Comparisons with cascade topology

Same output power



Same IM3



IM3 reduces by 8.8 dB
PAE: from 30% to 18%

Output power increases by 3 dB
PAE: from 30% to 27%

<u>Gain reduces by 15 dB</u>