Lecture 10

2017/2018

Microwave Devices and Circuits for Radiocommunications

Materials

- RF-OPTO
  - **http://rf-opto.etti.tuiasi.ro**
- **David Pozar, "Microwave Engineering"**, Wiley; 4th edition , 2011
  - 1 exam problem ← Pozar
- Photos
  - sent by email: rdamian@etti.tuiasi.ro
  - used at lectures/laboratory

Software

- ADS ~~2016~~ **2017**
- EmPro ~~2015~~ **2017**
- based on IP from outside university or campus

Date:

Grupa	5601 (2017/2018)
Specializarea	Master Retele de Comunicatii
Marca	857

Acceseaza ca acest student | Cere acces la licente

Note obtinute

Disciplina	Tip	Data	Descriere	Nota	Puncte	Obs.
TMPAW	Tehnici moderne de proiectare a aplicatiilor web					
	N	29/05/2017	Nota finala	10	-	

Nume

[redacted]

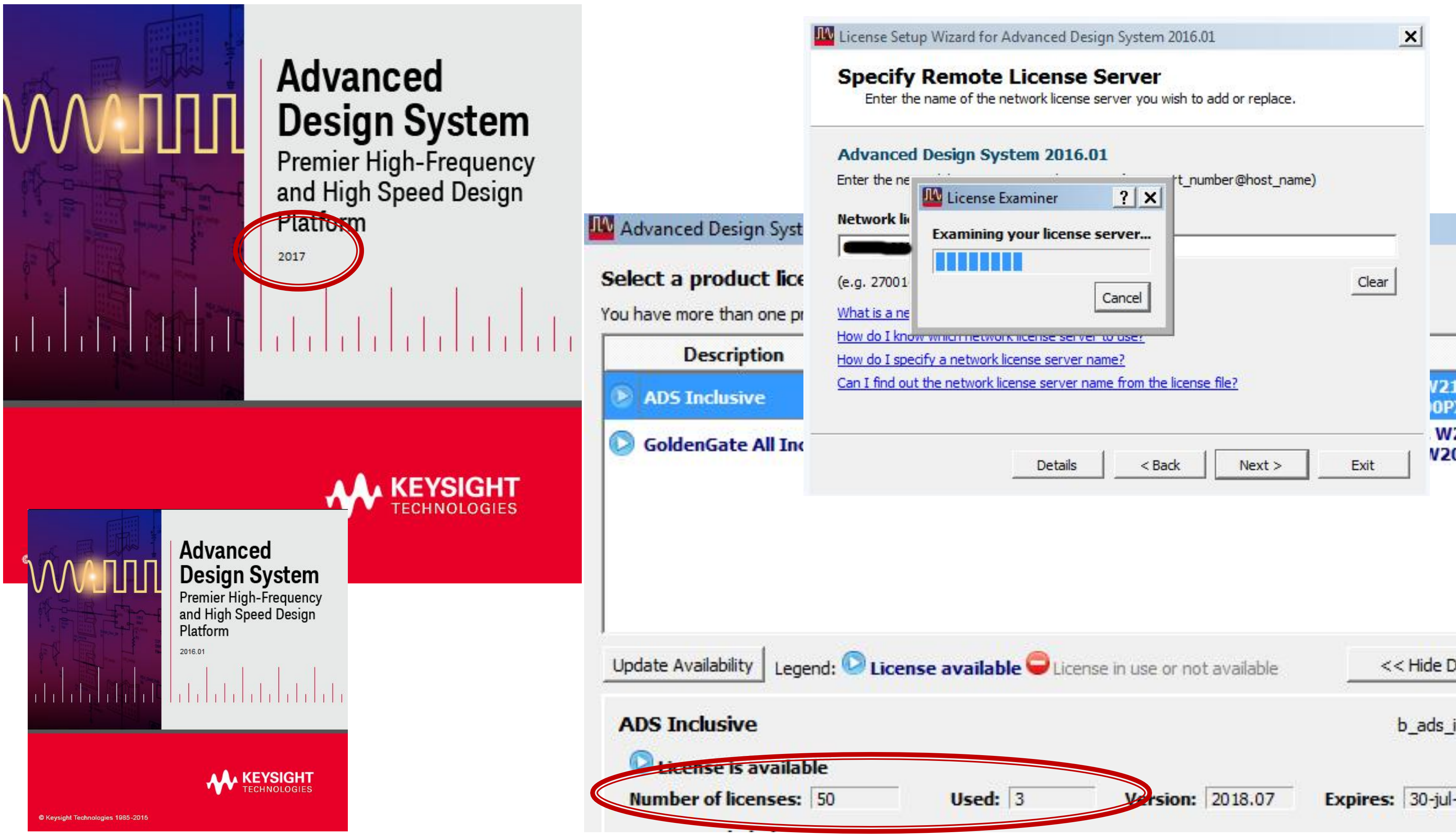
Email

Cod de verificare

344bd9f

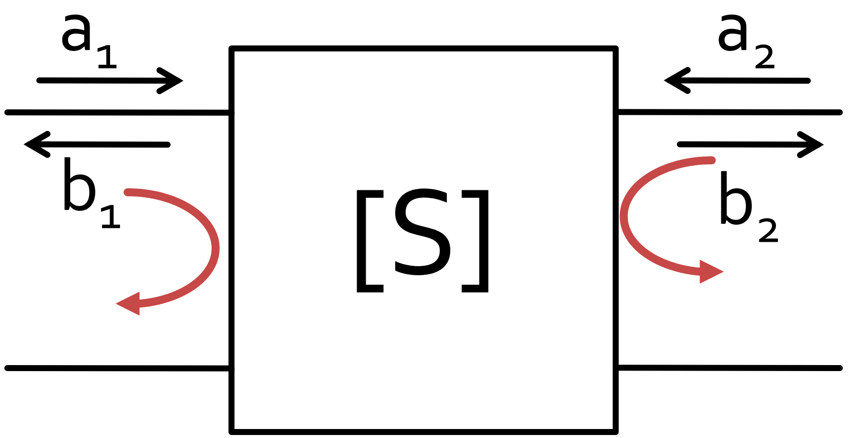
Trimite

Software


Advanced Design System
Premier High-Frequency and High Speed Design Platform
2017
KEYSIGHT TECHNOLOGIES
Advanced Design System
Premier High-Frequency and High Speed Design Platform
2016.01
KEYSIGHT TECHNOLOGIES
© Keysight Technologies 1985-2016
License Setup Wizard for Advanced Design System 2016.01
Specify Remote License Server
Enter the name of the network license server you wish to add or replace.
Advanced Design System 2016.01
License Examiner
Examining your license server...
Cancel
Clear
How do I specify a network license server name?
Can I find out the network license server name from the license file?
Details
< Back
Next >
Exit
Select a product lice
Description
ADS Inclusive
GoldenGate All Inc
Update Availability
Legend:
License available
License in use or not available
ADS Inclusive
License is available
Number of licenses: 50
Used: 3
Version: 2018.07
Expires:

Microwave Network Analysis

Scattering matrix – S

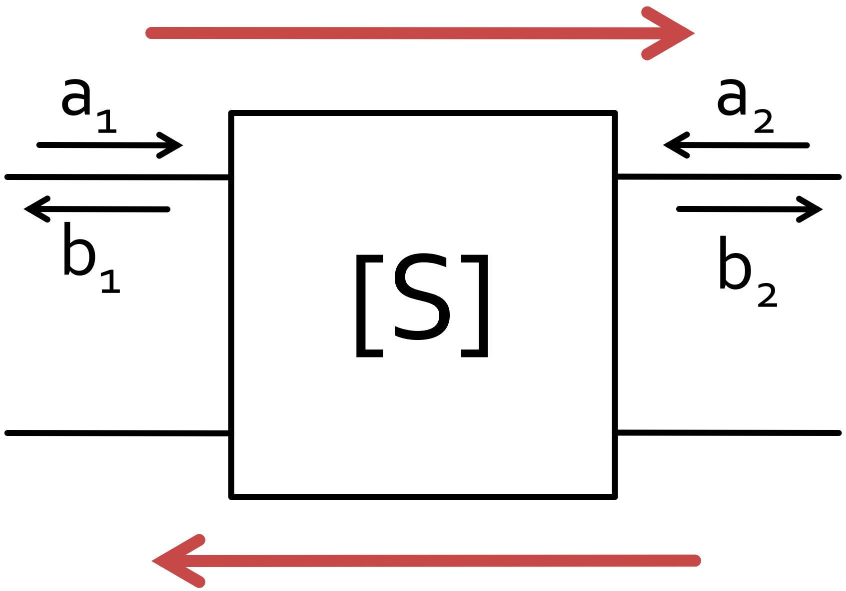


$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \cdot \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$$

$$S_{11} = \left. \frac{b_1}{a_1} \right|_{a_2=0} \qquad S_{22} = \left. \frac{b_2}{a_2} \right|_{a_1=0}$$

- S_{11} and S_{22} are reflection coefficients at ports 1 and 2 when the other port is matched

Scattering matrix – S

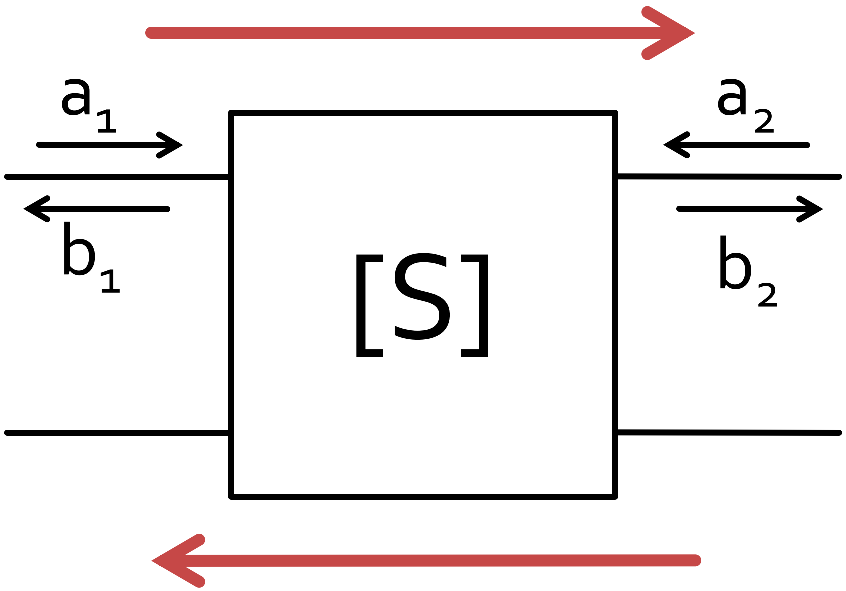


$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \cdot \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$$

$$S_{21} = \left. \frac{b_2}{a_1} \right|_{a_2=0} \qquad S_{12} = \left. \frac{b_1}{a_2} \right|_{a_1=0}$$

- S_{21} and S_{12} are signal amplitude gain when the other port is matched

Scattering matrix – S



$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \cdot \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$$

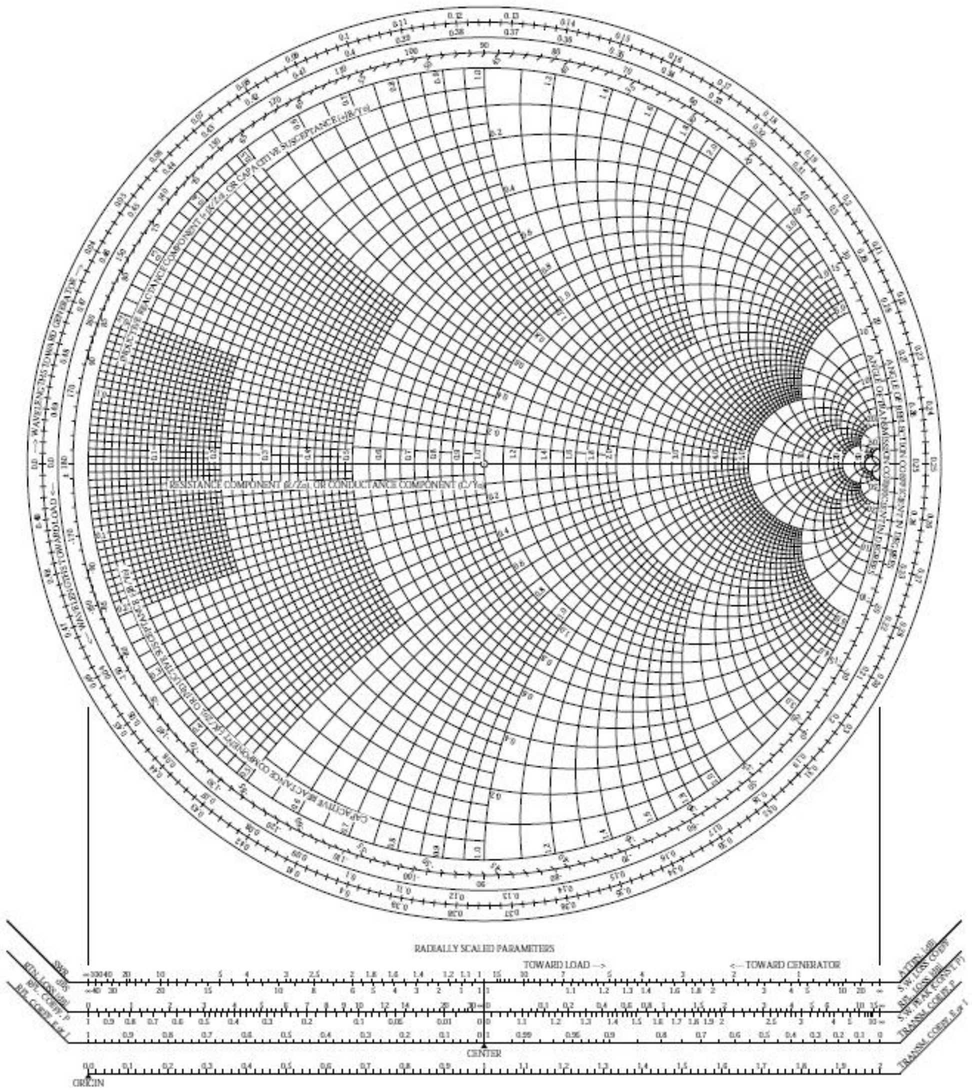
$$|S_{21}|^2 = \frac{Power\ in\ Z_0\ load}{Power\ from\ Z_0\ source}$$

- a,b
  - information about signal power **AND** signal phase
- S_{ij}
  - network effect (gain) over signal power **including** phase information

Impedance Matching

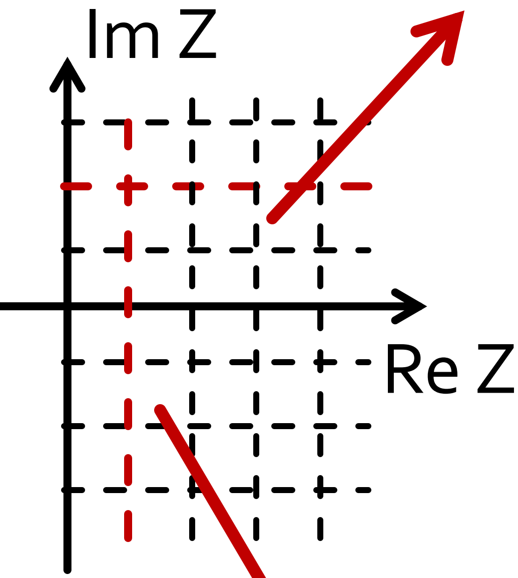
The Smith Chart

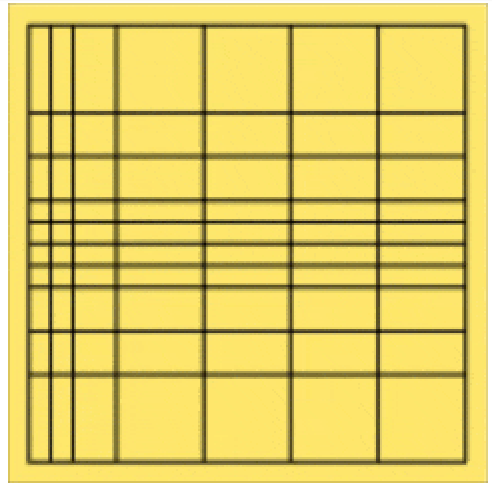
The Smith Chart

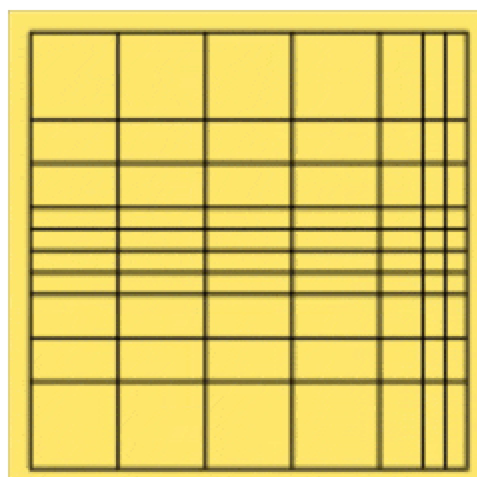


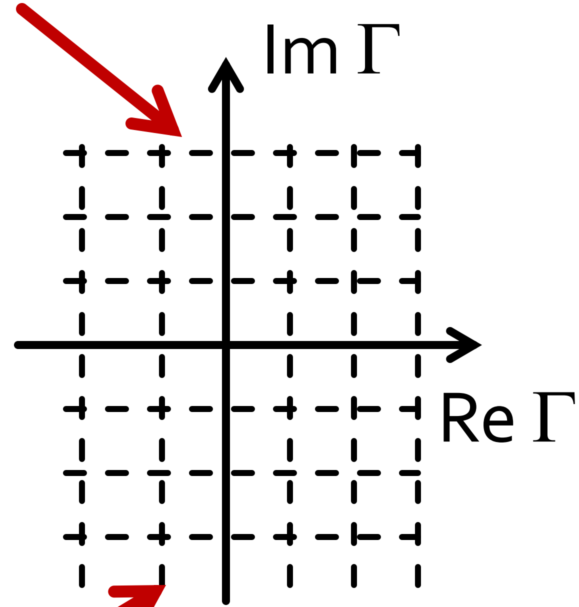
The Smith Chart

$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0} = \frac{z_L - 1}{z_L + 1}$$



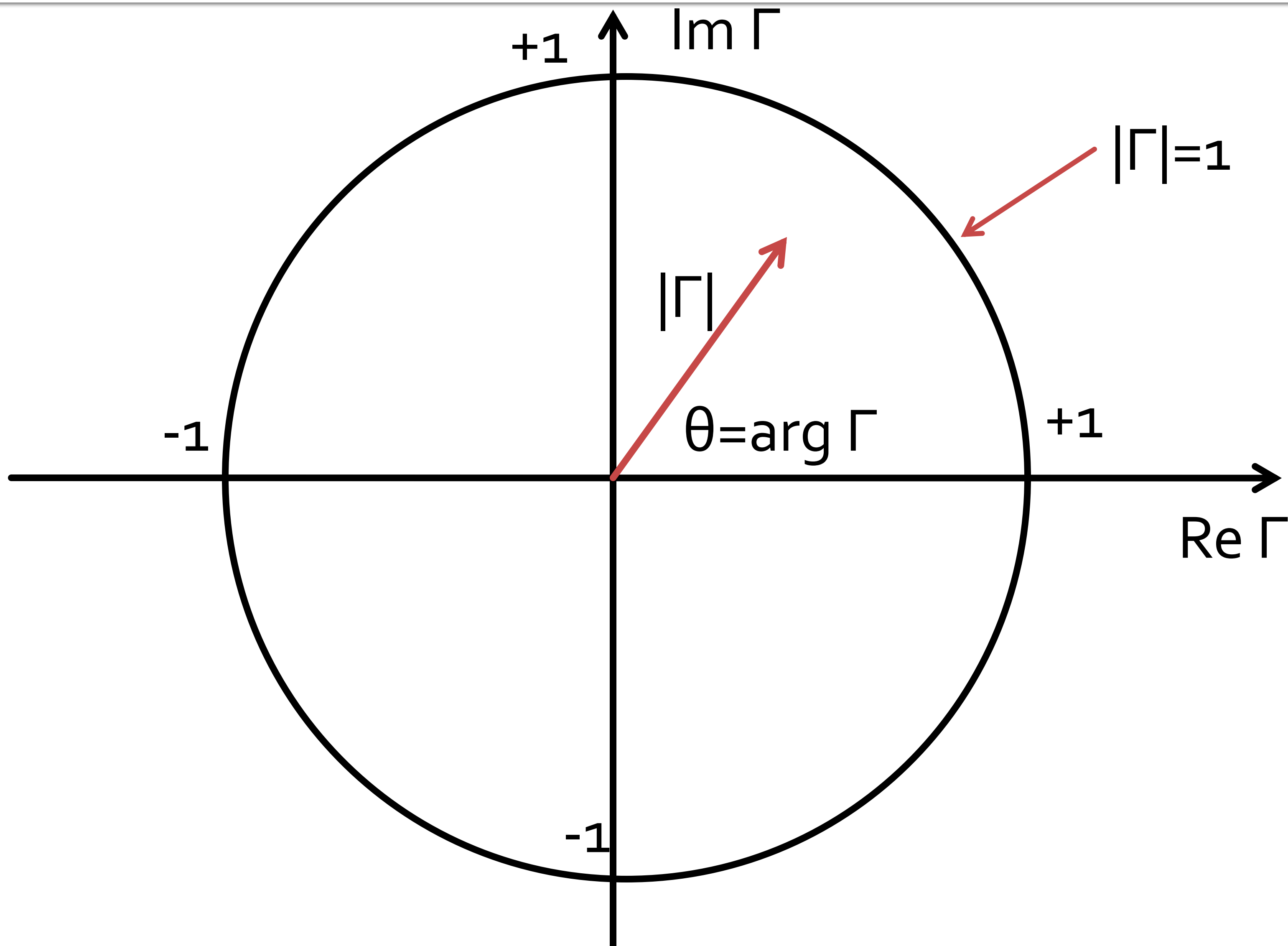






$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0} = \frac{Y_0 - Y_L}{Y_0 + Y_L} = \frac{1 - y_L}{1 + y_L}$$

The Smith Chart

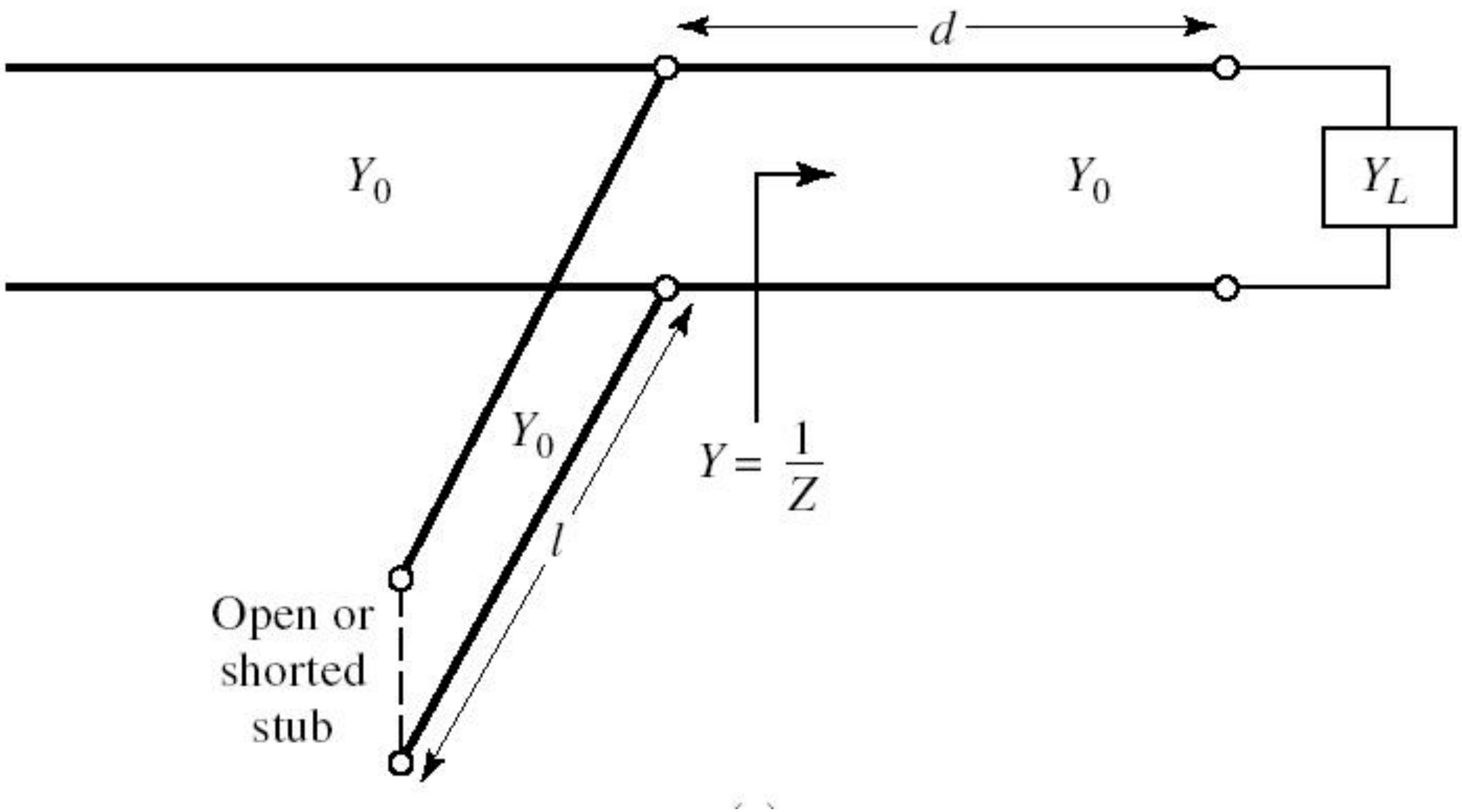


Impedance Matching with Stubs

Impedance Matching

Shunt Stub

- Shunt Stub



Analytical solution, Γ

$$\cos(\varphi + 2\theta) = -|\Gamma_S|$$

$$\theta_{sp} = \beta \cdot l = \tan^{-1} \frac{\mp 2 \cdot |\Gamma_S|}{\sqrt{1 - |\Gamma_S|^2}}$$

$$\Gamma_S = 0.593\angle 46.85°$$

$$|\Gamma_S| = 0.593; \quad \varphi = 46.85° \qquad \cos(\varphi + 2\theta) = -0.593 \Rightarrow (\varphi + 2\theta) = \pm 126.35°$$

- The **sign** (+/-) chosen for the **series line** equation imposes the **sign** used for the **shunt stub** equation
  - **"+" solution**

    $$(46.85° + 2\theta) = +126.35° \qquad \theta = +39.7° \qquad \operatorname{Im} y_S = \frac{-2 \cdot |\Gamma_S|}{\sqrt{1 - |\Gamma_S|^2}} = -1.472$$

    $$\theta_{sp} = \tan^{-1}(\operatorname{Im} y_S) = -55.8°(+180°) \rightarrow \theta_{sp} = 124.2°$$

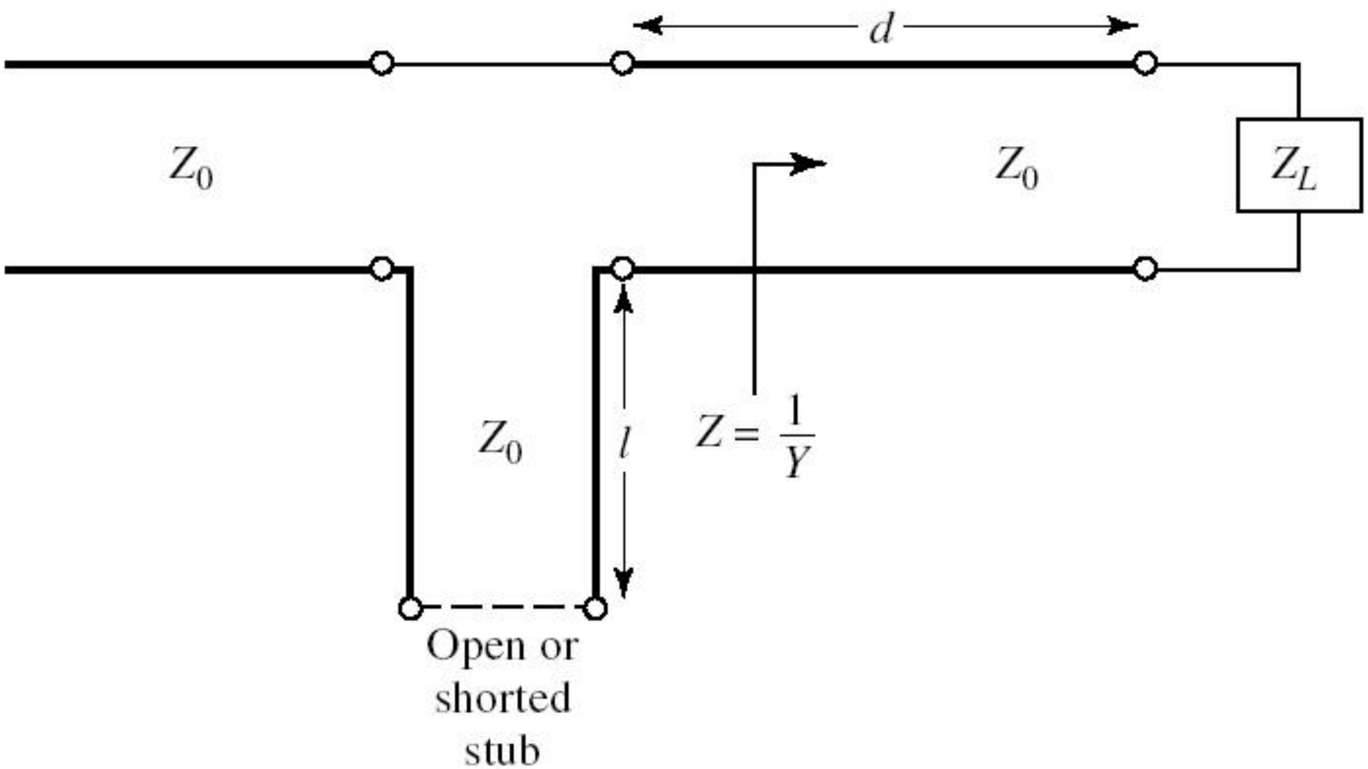
  - **"-" solution**

    $$(46.85° + 2\theta) = -126.35° \qquad \theta = -86.6°(+180°) \rightarrow \theta = 93.4°$$

    $$\operatorname{Im} y_S = \frac{+2 \cdot |\Gamma_S|}{\sqrt{1 - |\Gamma_S|^2}} = +1.472 \qquad \theta_{sp} = \tan^{-1}(\operatorname{Im} y_S) = 55.8°$$

Series Stub

- Series Stub
- difficult to realize in single conductor line technologies (microstrip)



Analytical solution, Γ

$$\cos(\varphi + 2\theta) = |\Gamma_S|$$

$$\theta_{ss} = \beta \cdot l = \cot^{-1} \frac{\mp 2 \cdot |\Gamma_S|}{\sqrt{1 - |\Gamma_S|^2}}$$

$$\Gamma_S = 0.555\angle -29.92°$$

$$|\Gamma_S| = 0.555; \quad \varphi = -29.92° \qquad \cos(\varphi + 2\theta) = 0.555 \Rightarrow (\varphi + 2\theta) = \pm 56.28°$$

- The **sign** (+/-) chosen for the **series line** equation imposes the **sign** used for the **series stub** equation
  - **"+" solution**

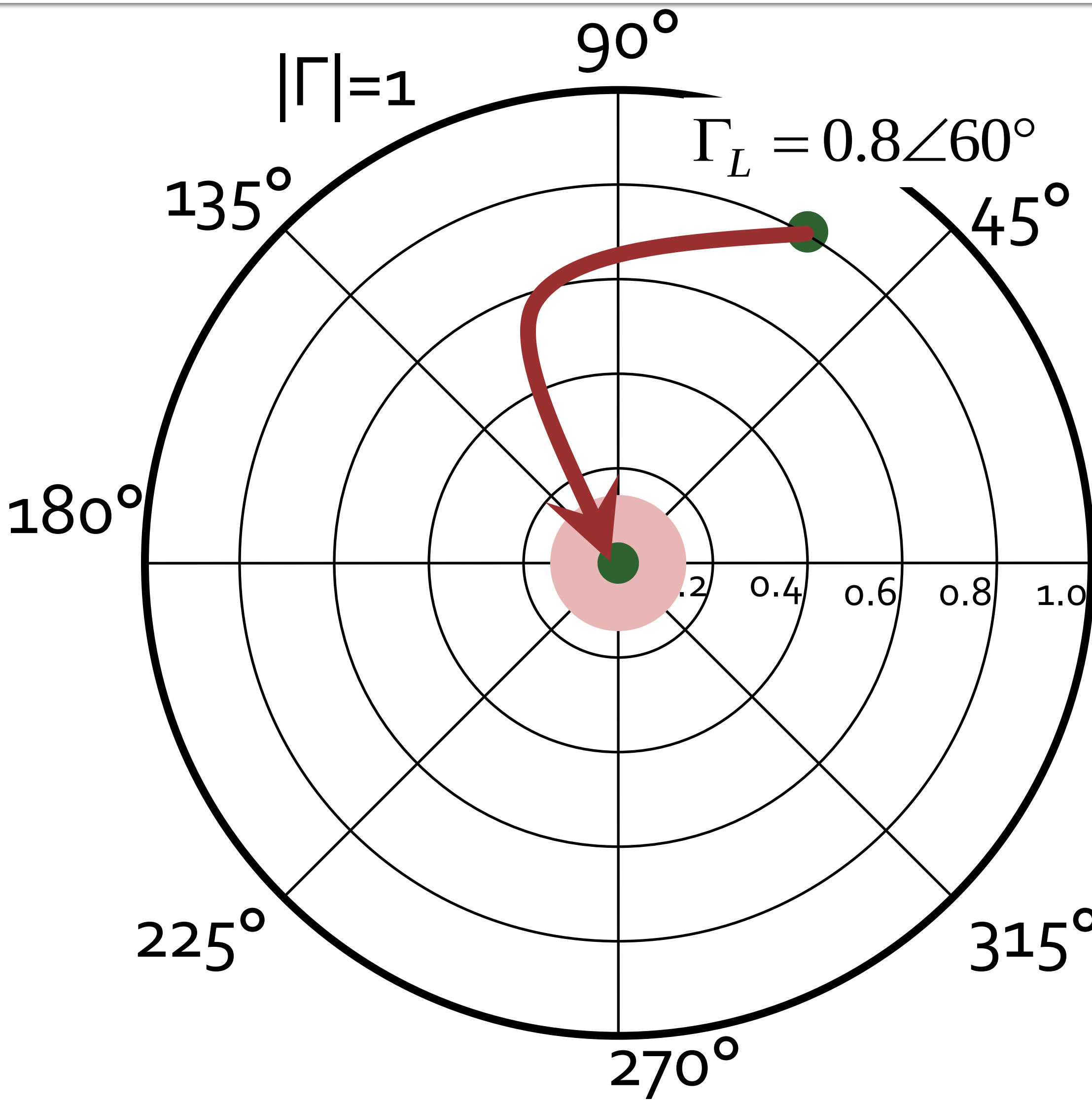
    $$(-29.92° + 2\theta) = +56.28° \qquad \theta = 43.1° \qquad \operatorname{Im} z_S = \frac{+2 \cdot |\Gamma_S|}{\sqrt{1 - |\Gamma_S|^2}} = +1.335$$

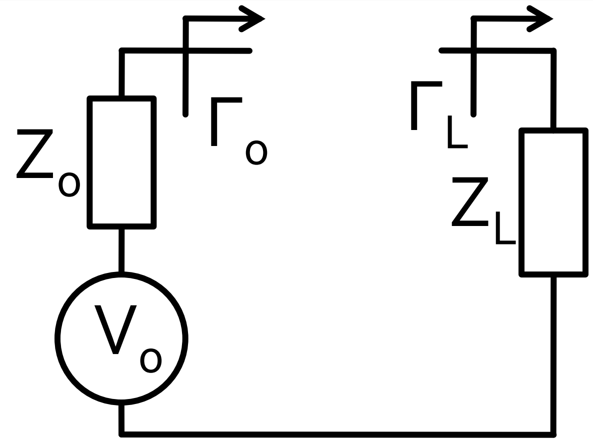
    $$\theta_{ss} = -\cot^{-1}(\operatorname{Im} z_S) = -36.8°(+180°) \rightarrow \theta_{ss} = 143.2°$$
  - **"-" solution**

    $$(-29.92° + 2\theta) = -56.28° \qquad \theta = -13.2°(+180°) \rightarrow \theta = 166.8°$$

    $$\operatorname{Im} z_S = \frac{-2 \cdot |\Gamma_S|}{\sqrt{1 - |\Gamma_S|^2}} = -1.335 \qquad \theta_{ss} = -\cot^{-1}(\operatorname{Im} z_S) = 36.8°$$

The Smith Chart, matching, $Z_L \neq Z_o$





Matching Z_L load to Z_o source.
We normalize Z_L over Z_o

$$Z_L = 21.429\Omega + j \cdot 82.479\Omega$$

$$z_L = 0.429 + j \cdot 1.65$$

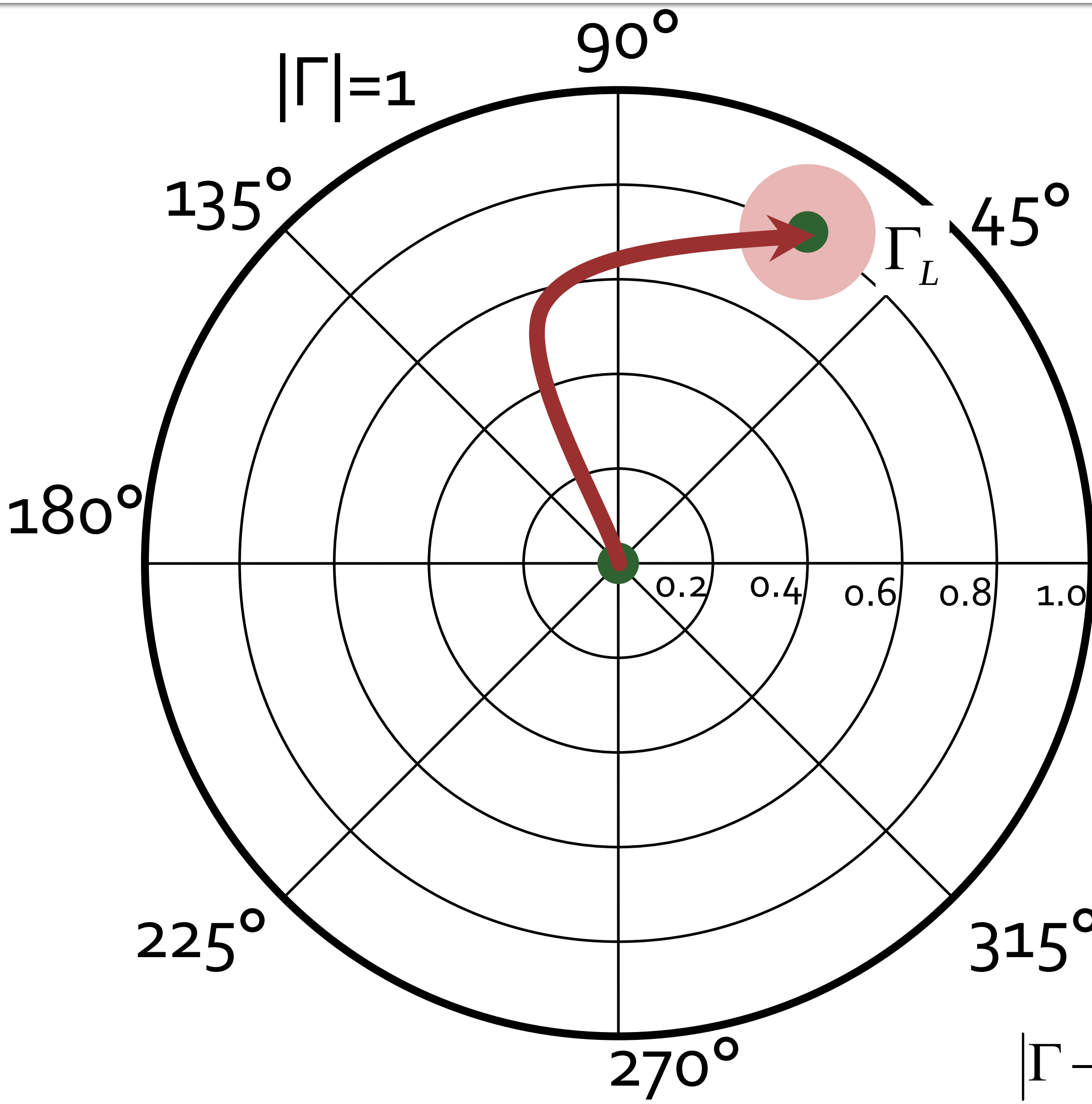
$$\Gamma_L = 0.8\angle 60°$$

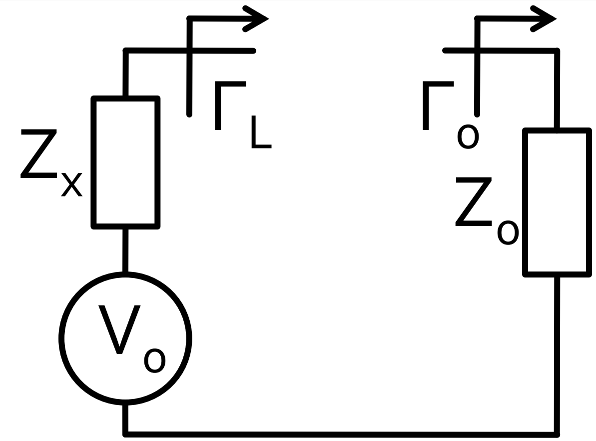
We must move the point denoting the reflection coefficient in the area where with a Z_o source we have:

$\Gamma_0 = 0$ perfect match

$|\Gamma_0| \leq \Gamma_m$ "good enough" match

The Smith Chart, matching, $Z_L = Z_o$





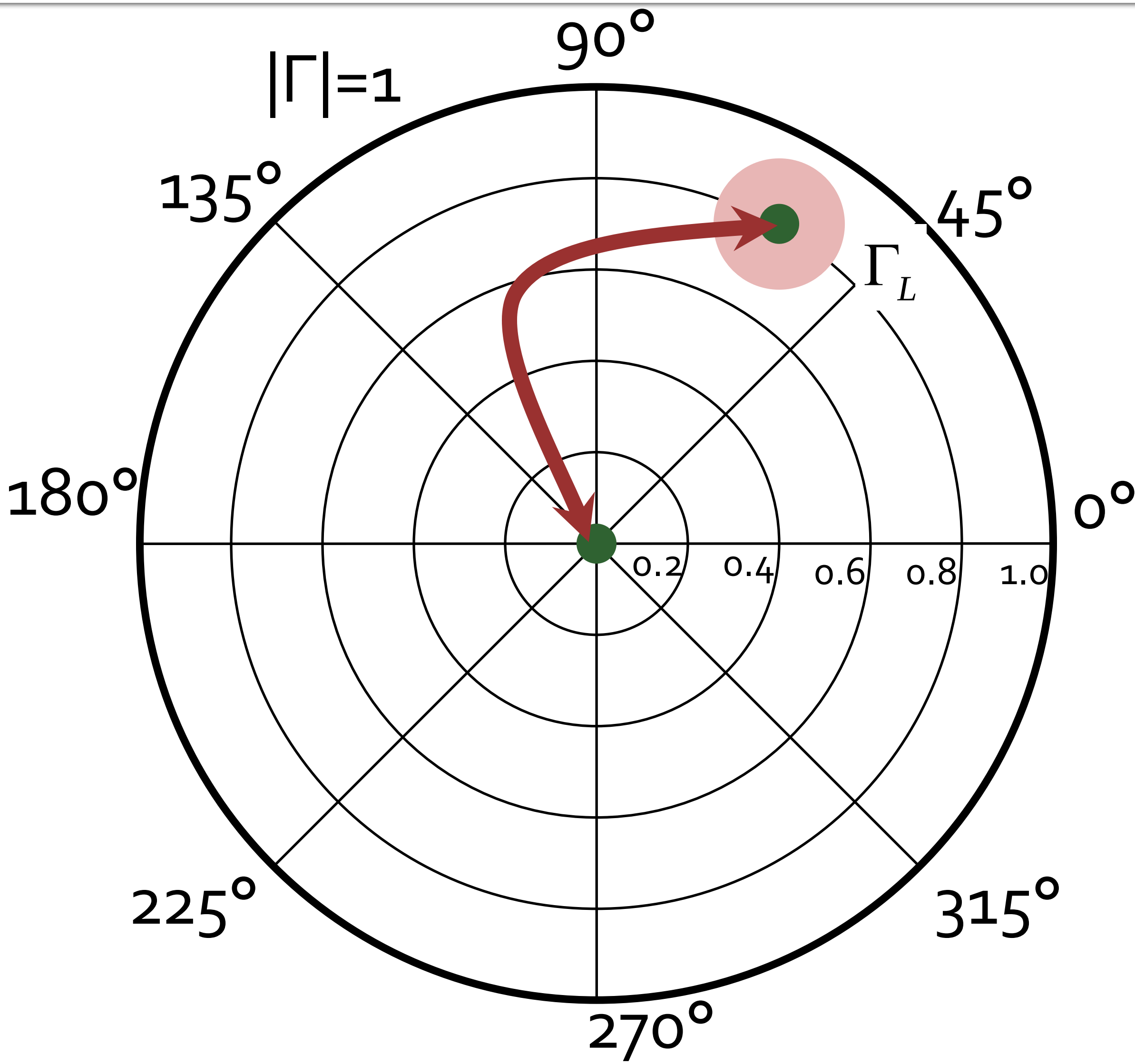
The source (eg. the transistor) having Z_X needs to see a certain reflection coefficient Γ_L towards the load Z_o

The matching circuit must move the point denoting the reflection coefficient in the area where for a Z_o load (Γ_o=0) we see towards it:

$\Gamma = \Gamma_L$ perfect match

$|\Gamma - \Gamma_L| \le \Gamma_m$ "good enough" match

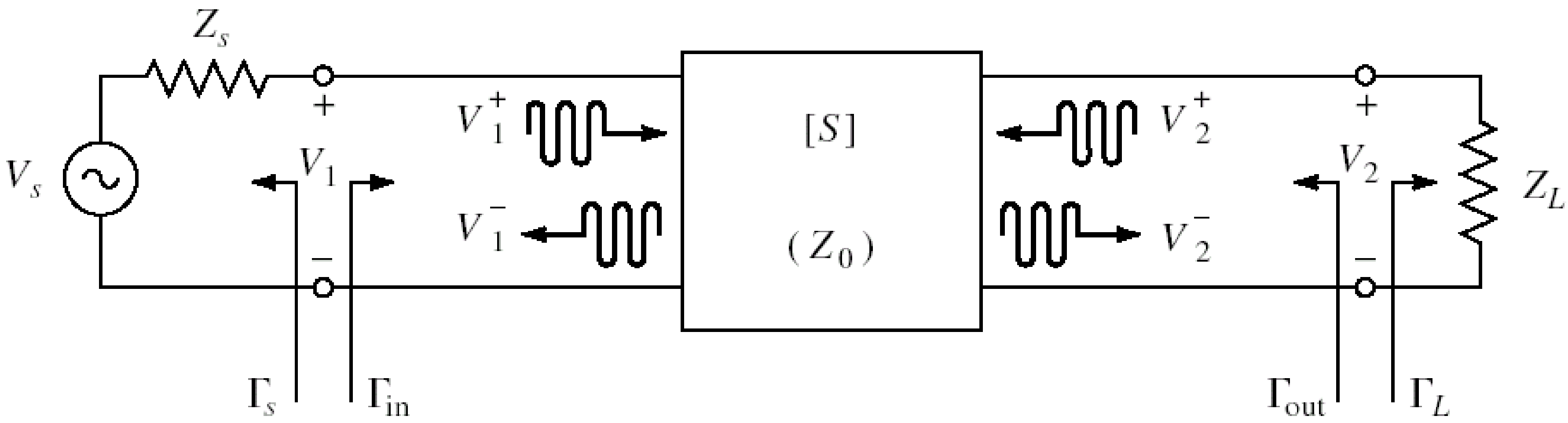
The Smith Chart, matching , $Z_L \neq Z_o$, $Z_L = Z_o$



- The matching sections needed to move
  - Γ_L in Γ_o
  - Γ_o in Γ_L
- are **identical**. They differ only by the **order** in which the elements are introduced into the matching circuit
- As a result, we can use in match design the same:
  - **methods**
  - **formulae**

Microwave Amplifiers

Amplifier as two-port



- Charaterized with S parameters
- normalized at Zo (implicit 50Ω)
- Datasheets: S parameters for specific bias conditions

Datasheets

NE46100

VCE = 5 V, IC = 50 mA

FREQUENCY (MHz)	S11 MAG	S11 ANG	S21 MAG	S21 ANG	S12 MAG	S12 ANG	S22 MAG	S22 ANG	K	MAG[2] (dB)
100	0.778	-137	26.776	114	0.028	30	0.555	-102	0.16	29.8
200	0.815	-159	14.407	100	0.035	29	0.434	-135	0.36	26.2
500	0.826	-177	5.855	84	0.040	38	0.400	-162	0.75	21.7
800	0.827	176	3.682	76	0.052	43	0.402	-169	0.91	18.5
1000	0.826	173	2.963	71	0.058	47	0.405	-172	1.02	16.3
1200	0.825	170	2.441	66	0.064	47	0.412	-174	1.08	14.0
1400	0.820	167	2.111	61	0.069	47	0.413	-176	1.17	12.4
1600	0.828	165	1.863	57	0.078	54	0.426	-177	1.15	11.4
1800	0.827	162	1.671	53	0.087	50	0.432	-178	1.14	10.6
2000	0.828	159	1.484	49	0.093	50	0.431	-180	1.17	9.5
2500	0.822	153	1.218	39	0.11	48	0.462	177	1.18	7.8
3000	0.818	148	1.010	30	0.135	46	0.490	174	1.16	6.3
3500	0.824	142	0.876	21	0.147	44	0.507	170	1.16	5.3
4000	0.812	137	0.762	13	0.168	38	0.535	167	1.14	4.3

VCE = 5 V, IC = 100 mA

100	0.778	-144	27.669	111	0.027	35	0.523	-114	0.27	30.2
200	0.820	-164	14.559	97	0.029	29	0.445	-144	0.42	27.0
500	0.832	-179	5.885	84	0.035	38	0.435	-166	0.81	22.2
800	0.833	175	3.691	76	0.048	45	0.435	-173	0.95	18.8
1000	0.831	172	2.980	71	0.056	51	0.437	-176	1.05	16.0
1200	0.836	169	2.464	67	0.061	52	0.432	-178	1.11	14.0
1400	0.829	166	2.121	61	0.072	53	0.447	-180	1.12	12.6
1600	0.831	164	1.867	58	0.080	54	0.445	179	1.14	11.4

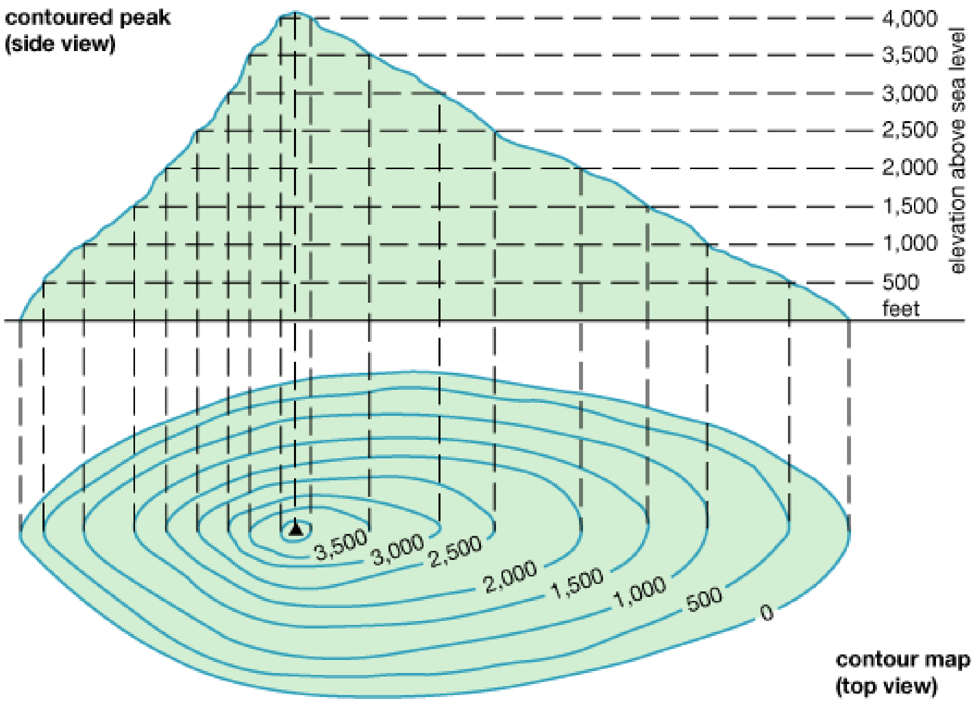
S2P - Touchstone

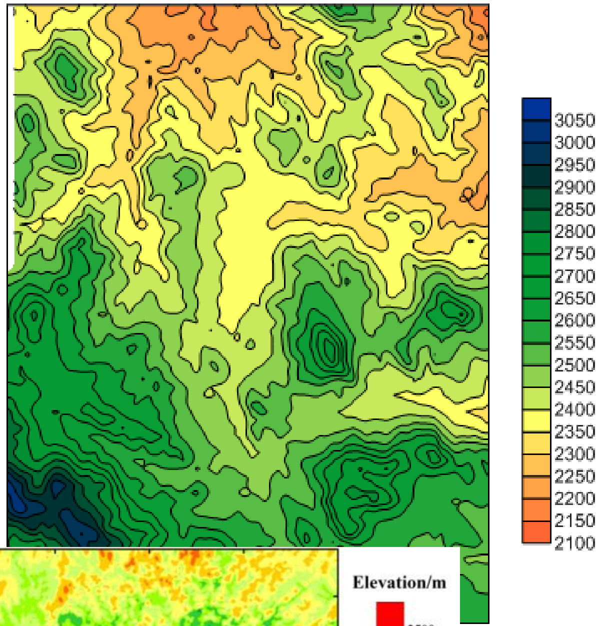
- Touchstone file format (*.s2p)

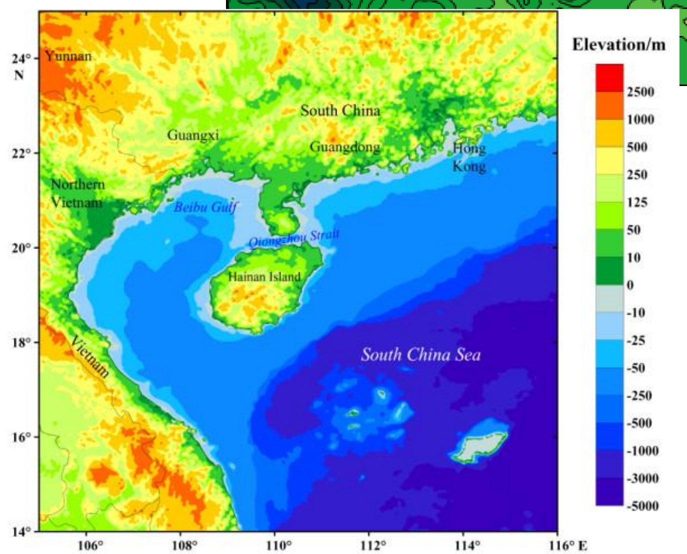
```
! SIEMENS Small Signal Semiconductors
! VDS = 3.5 V     ID = 15 mA
# GHz  S  MA  R  50
!  f       S11         S21         S12         S22
! GHz    MAG   ANG    MAG   ANG    MAG   ANG    MAG   ANG
 1.000 0.9800  -18.0   2.230  157.0  0.0240   74.0  0.6900  -15.0
 2.000 0.9500  -39.0   2.220  136.0  0.0450   57.0  0.6600  -30.0
 3.000 0.8900  -64.0   2.210  110.0  0.0680   40.0  0.6100  -45.0
 4.000 0.8200  -89.0   2.230   86.0  0.0850   23.0  0.5600  -62.0
 5.000 0.7400 -115.0   2.190   61.0  0.0990    7.0  0.4900  -80.0
 6.000 0.6500 -142.0   2.110   36.0  0.1070  -10.0  0.4100  -98.0
!
!  f      Fmin   Gammaopt  rn/50
! GHz      dB    MAG  ANG    -
 2.000     1.00  0.72   27   0.84
 4.000     1.40  0.64   61   0.58
```

Contour map/lines

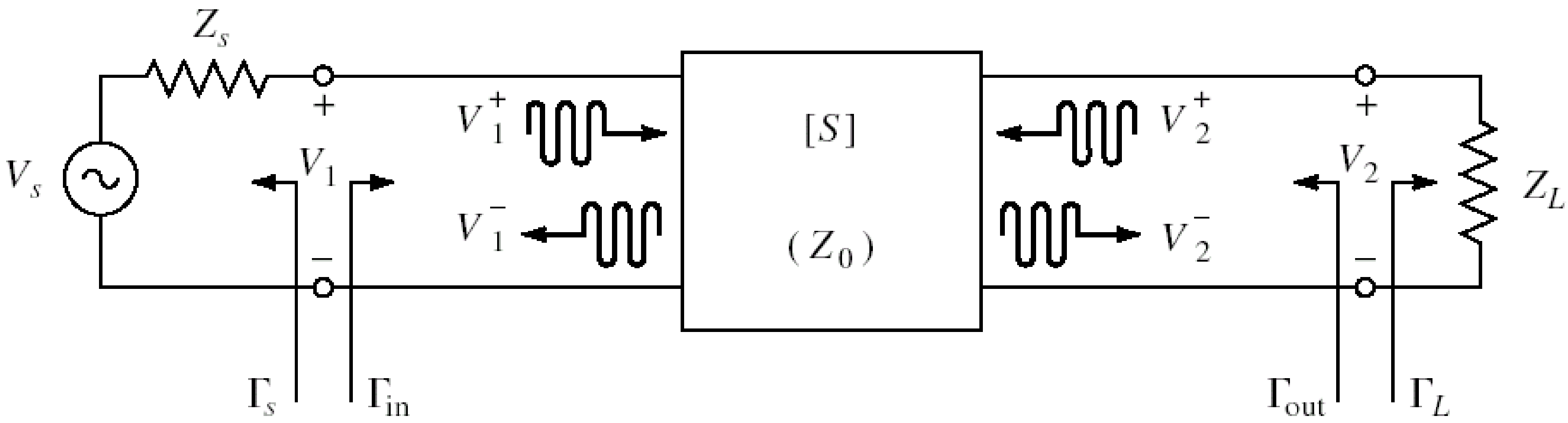
$$\begin{cases} F = f(x,y) & x,y \in \mathbf{R} \\ F = f(z) & z \in \mathbf{C} \end{cases}$$







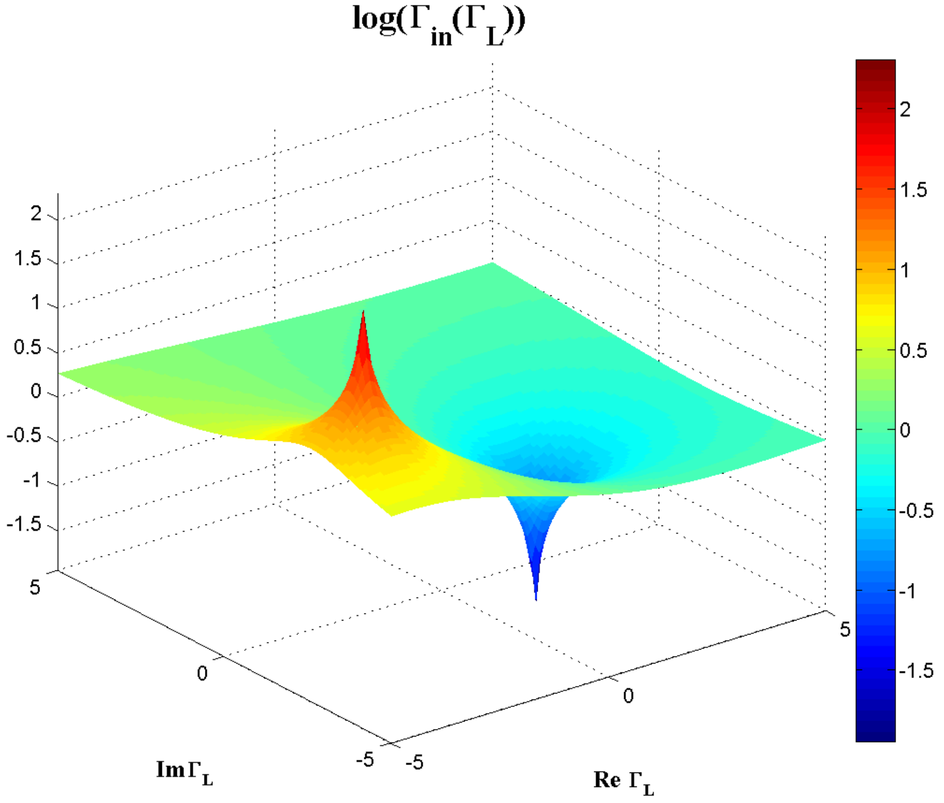
Amplifier as two-port

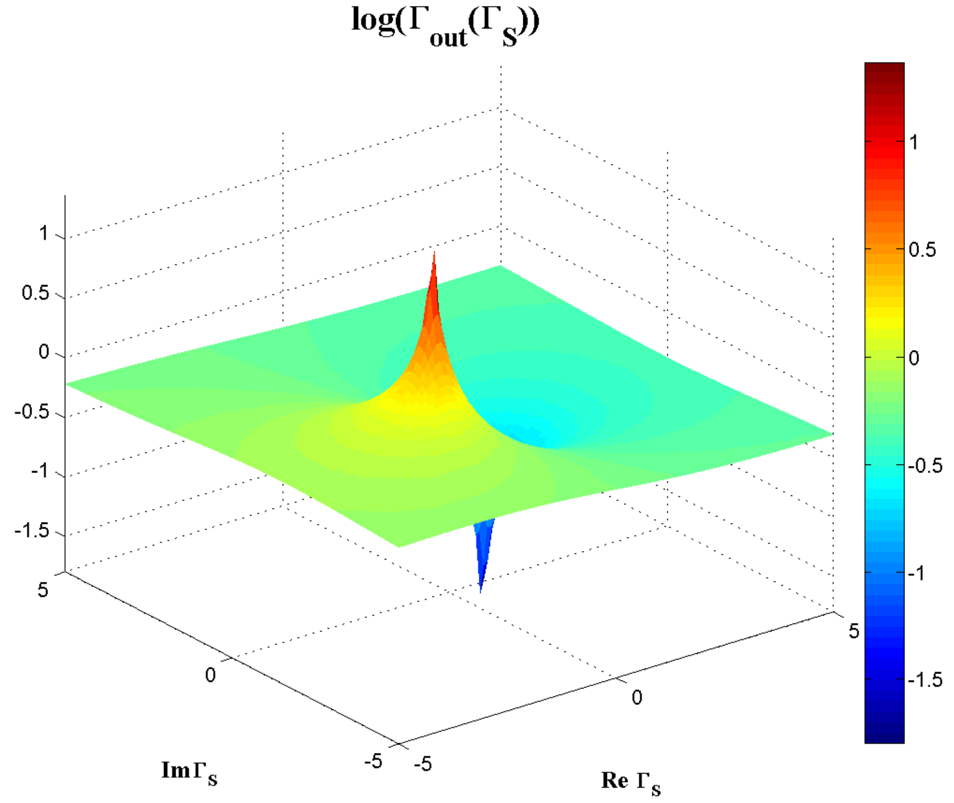


- For an amplifier two-port we are interested in:
  - **stability**
  - power gain
  - noise (sometimes – small signals)
  - linearity (sometimes – large signals)

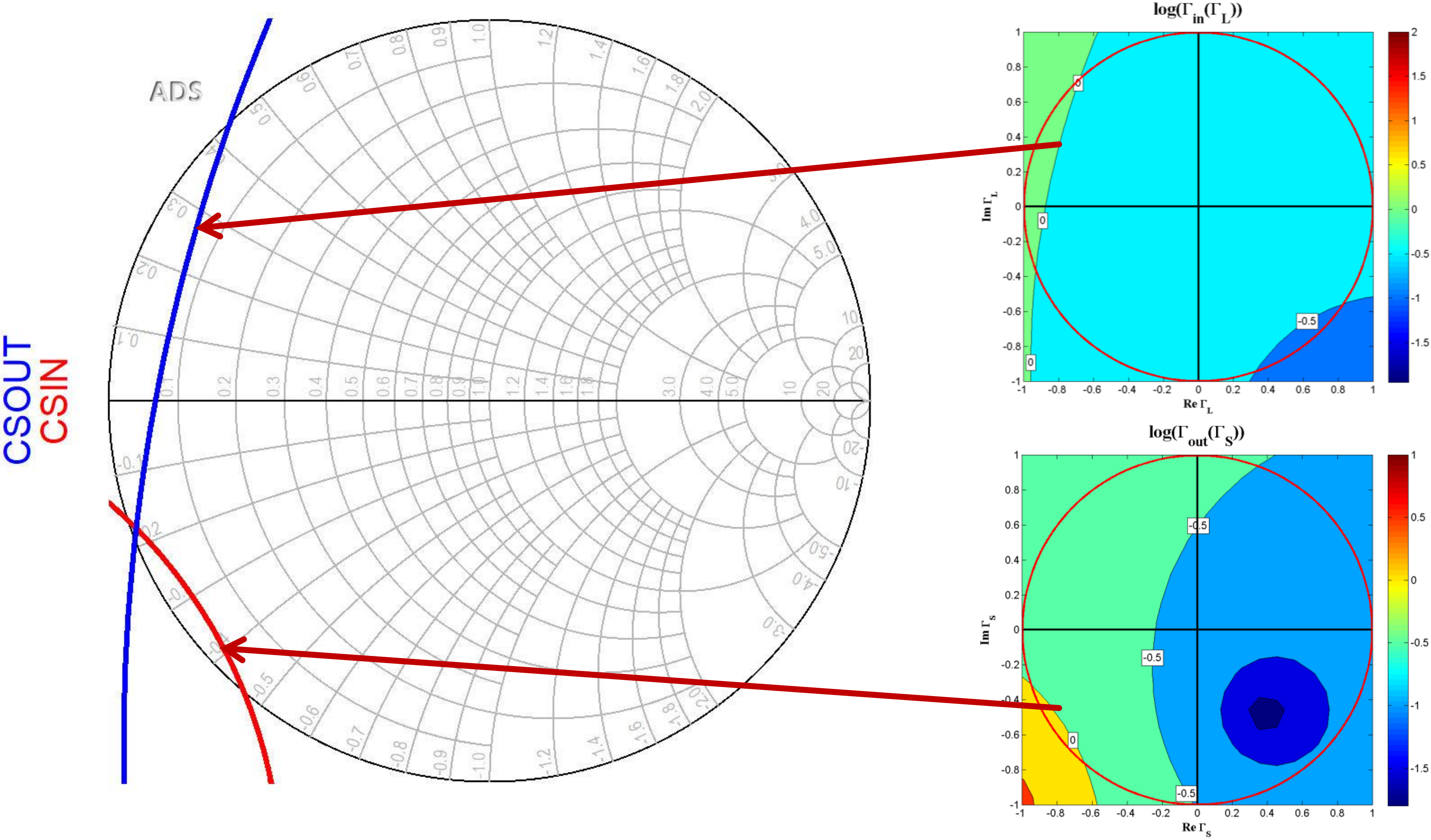
3D representation of $|\Gamma_{in}|$, $|\Gamma_{out}|$

- $\log_{10}|\Gamma_{in}|$, $\log_{10}|\Gamma_{out}|$





CSIN, CSOUT, $|\Gamma_{in}|=1$, $|\Gamma_{out}|=1$



Rollet's condition

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2 \cdot |S_{12} \cdot S_{21}|}$$

$$\Delta = S_{11} \cdot S_{22} - S_{12} \cdot S_{21}$$

- The two-port is **unconditionally stable** if:
- two conditions are simultaneously satisfied:
  - $K > 1$
  - $|\Delta| < 1$
- together with the implicit conditions:
  - $|S_{11}| < 1$
  - $|S_{22}| < 1$

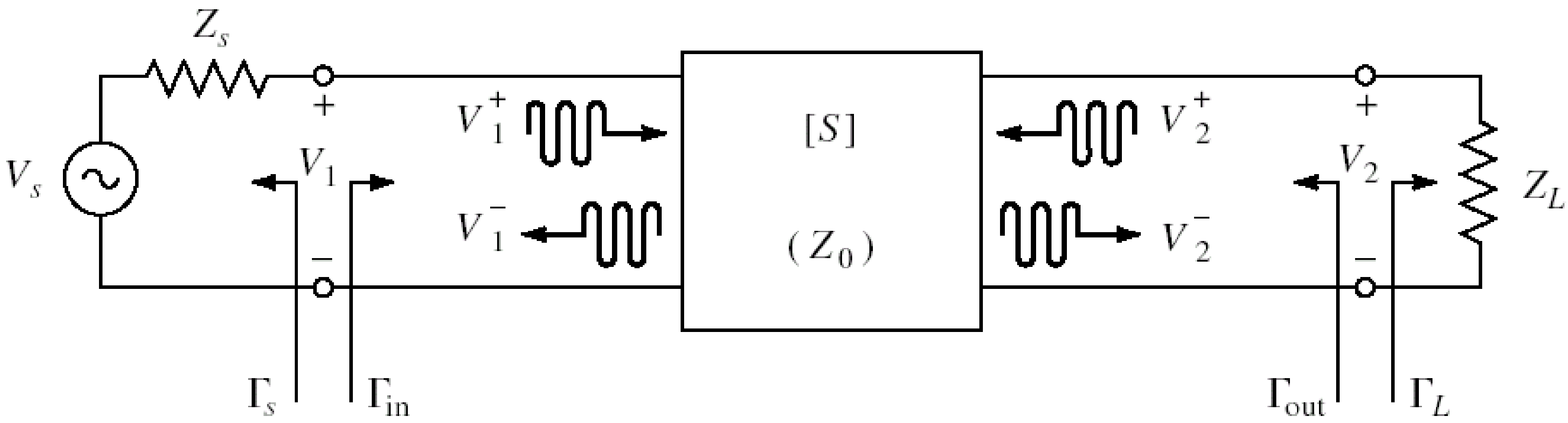
$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2 \cdot |S_{12} \cdot S_{21}|} > 1$$

$$|\Delta| = |S_{11} \cdot S_{22} - S_{12} \cdot S_{21}| < 1$$

Power Gain of Microwave Amplifiers

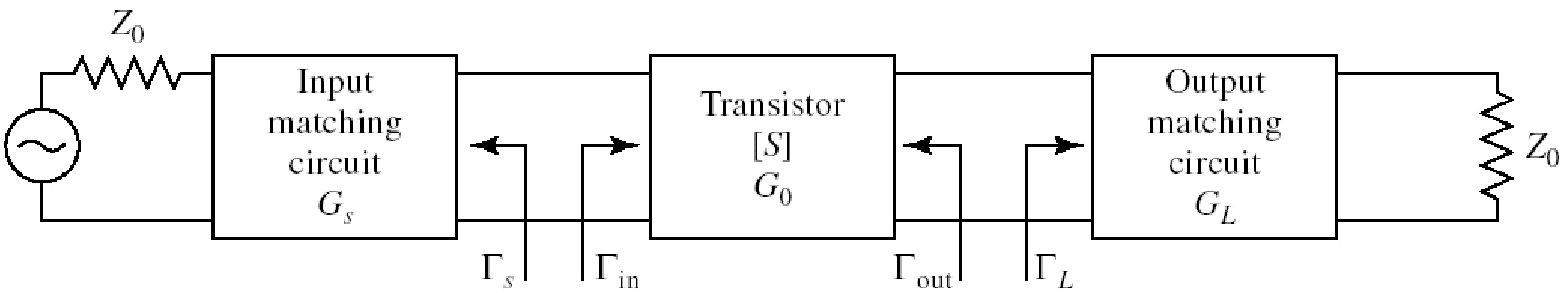
Microwave Amplifiers

Amplifier as two-port



- For an amplifier two-port we are interested in:
  - stability
  - **power gain**
  - noise (sometimes – small signals)
  - linearity (sometimes – large signals)

Design for Maximum Gain



- Maximum power gain (complex conjugate matching):

$$\Gamma_{in} = \Gamma_S^* \qquad \Gamma_{out} = \Gamma_L^*$$

- For lossless matching sections

$$G_{T\max} = \frac{|S_{21}|^2 \cdot \left(1 - |\Gamma_S|^2\right) \cdot \left(1 - |\Gamma_L|^2\right)}{|1 - \Gamma_S \cdot \Gamma_{in}|^2 \cdot |1 - S_{22} \cdot \Gamma_L|^2} \qquad G_{T\max} = \frac{1}{1 - |\Gamma_S|^2} \cdot |S_{21}|^2 \cdot \frac{1 - |\Gamma_L|^2}{|1 - S_{22} \cdot \Gamma_L|^2}$$

- For the general case of the bilateral transistor ($S_{12} \neq 0$) Γ_{in} and Γ_{out} depend on each other so the input and output sections must be matched simultaneously

Simultaneous matching

- Simultaneous matching can be achieved **if and only if** the amplifier is **unconditionally stable** at the operating frequency, and |Γ|<1 solutions are those with "–" sign of quadratic solutions

$$\Gamma_S = \frac{B_1 - \sqrt{B_1^2 - 4 \cdot |C_1|^2}}{2 \cdot C_1} \qquad \Gamma_L = \frac{B_2 - \sqrt{B_2^2 - 4 \cdot |C_2|^2}}{2 \cdot C_2}$$

$$\begin{cases} B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 \\ C_1 = S_{11} - \Delta \cdot S_{22}^* \end{cases} \qquad \begin{cases} B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2 \\ C_2 = S_{22} - \Delta \cdot S_{11}^* \end{cases}$$

Example

- ATF-34143 **at Vds=4V Id=40mA**.
- @5GHz
  - $S_{11} = 0.64\angle 111°$
  - $S_{12} = 0.117\angle -27°$
  - $S_{21} = 2.923\angle -6°$
  - $S_{22} = 0.21\angle 111°$

Computations

$$\begin{cases} B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 \\ C_1 = S_{11} - \Delta \cdot S_{22}^* \end{cases}$$

$$\begin{cases} B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2 \\ C_2 = S_{22} - \Delta \cdot S_{11}^* \end{cases}$$

$$\begin{cases} B_1 = 1.207 \\ C_1 = -0.277 + j \cdot 0.529 \end{cases}$$

$$\begin{cases} B_2 = 0.476 \\ C_2 = -0.222 - j \cdot 0.013 \end{cases}$$

$$\Gamma_S = \frac{B_1 - \sqrt{B_1^2 - 4 \cdot |C_1|^2}}{2 \cdot C_1}$$

$$\Gamma_L = \frac{B_2 - \sqrt{B_2^2 - 4 \cdot |C_2|^2}}{2 \cdot C_2}$$

$$\Gamma_S = -0.403 - j \cdot 0.768$$

$$|\Gamma_S| = 0.867 < 1$$

$$\Gamma_S = 0.867 \angle -117.7°$$

$$\Gamma_L = -0.685 + j \cdot 0.04$$

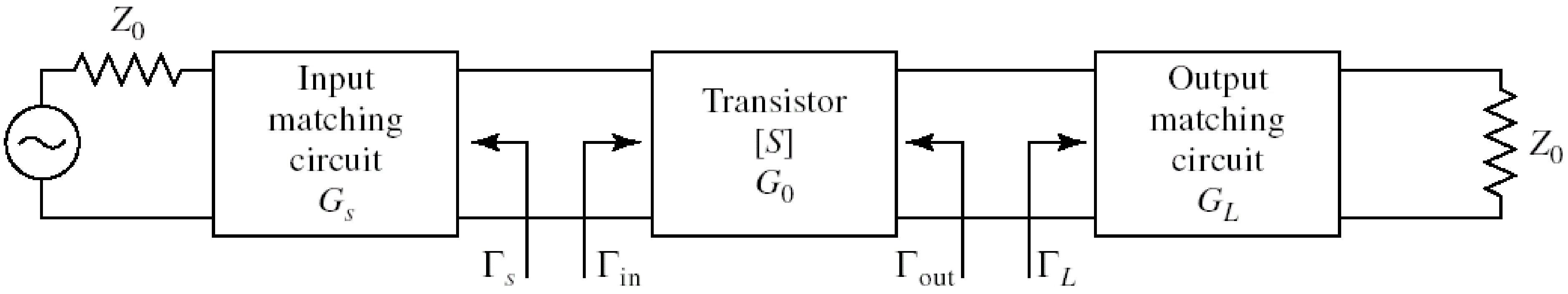
$$|\Gamma_L| = 0.686 < 1$$

$$\Gamma_L = 0.686 \angle 176.7°$$

Design for Specified Gain

- In many cases we need an approach other than "brute force" when we prefer to design for **less than the maximum obtainable gain**, in order to:
  - improve noise behavior (L3 + C9)
  - improve stability
  - improve VSWR
  - control performance at multiple frequencies
  - improve amplifier's bandwidth

Design for Specified Gain



- In the unilateral assumption:

$$G_{TU} = \frac{1-|\Gamma_S|^2}{|1-S_{11}\cdot\Gamma_S|^2}\cdot|S_{21}|^2\cdot\frac{1-|\Gamma_L|^2}{|1-S_{22}\cdot\Gamma_L|^2}$$

$$G_S = \frac{1-|\Gamma_S|^2}{|1-S_{11}\cdot\Gamma_S|^2}$$

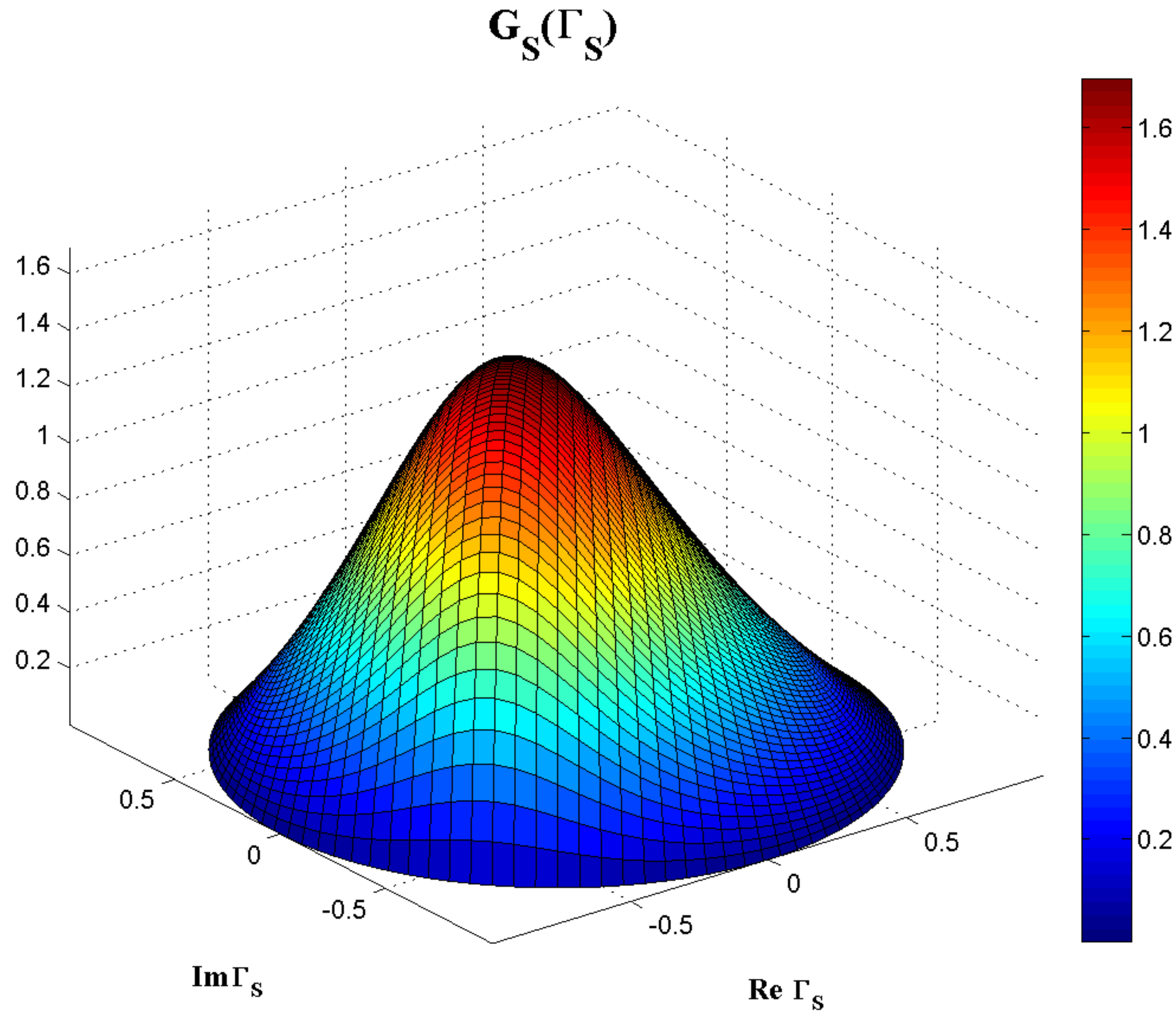
$$G_S = G_S(\Gamma_S)$$

$$G_0 = |S_{21}|^2$$

$$G_L = \frac{1-|\Gamma_L|^2}{|1-S_{22}\cdot\Gamma_L|^2}$$

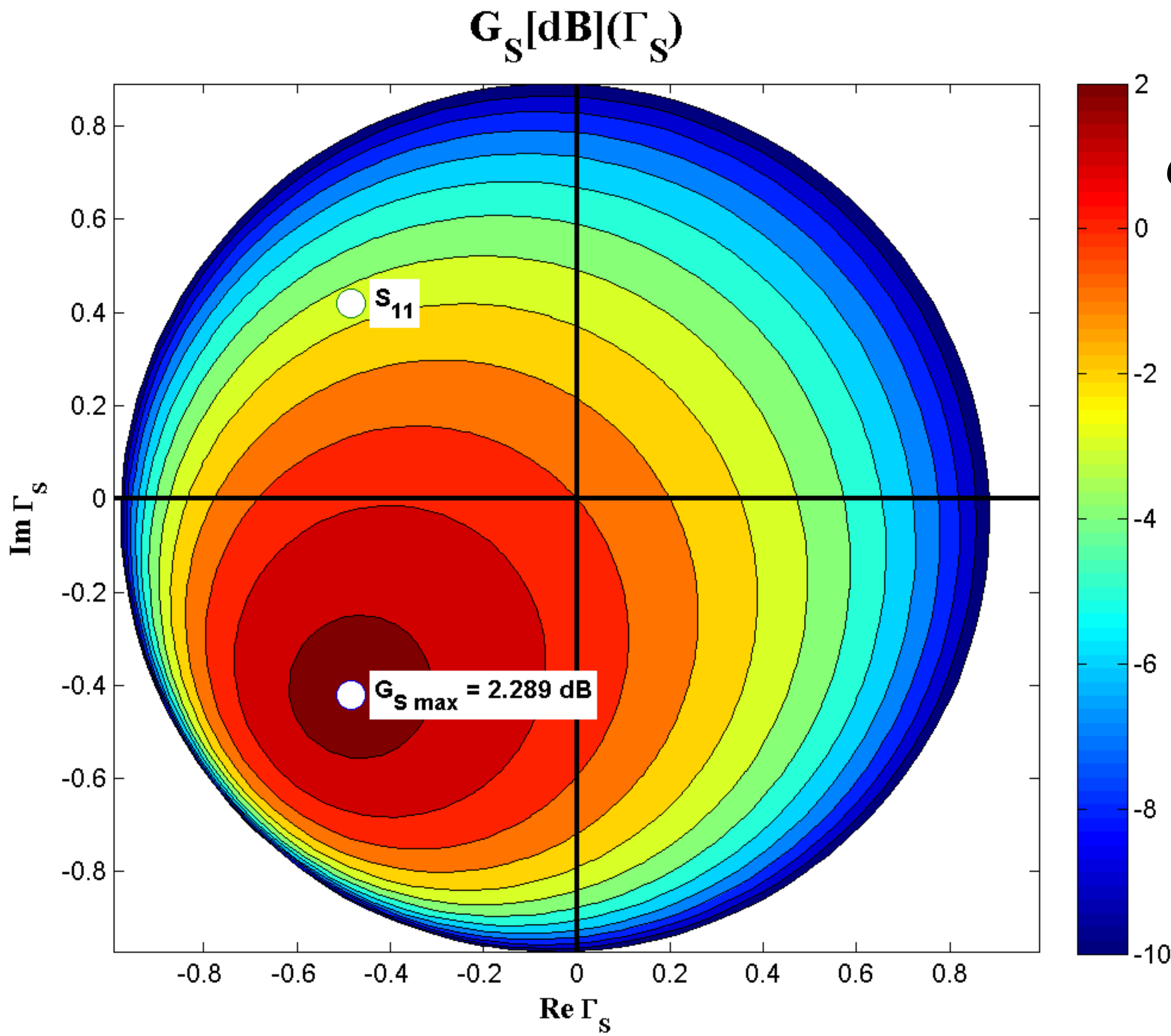
$$G_L = G_L(\Gamma_L)$$

$G_S(\Gamma_S)$



$$G_S = \frac{1-|\Gamma_S|^2}{|1-S_{11}\cdot\Gamma_S|^2}$$

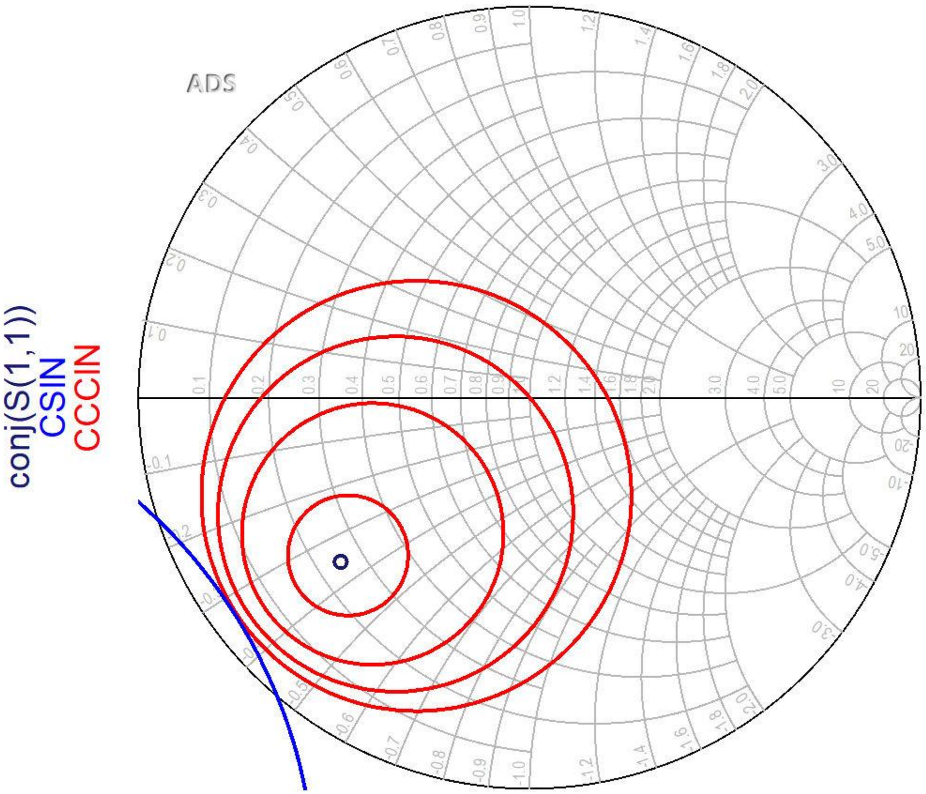
G_S[dB](Γ_S), constant value contours

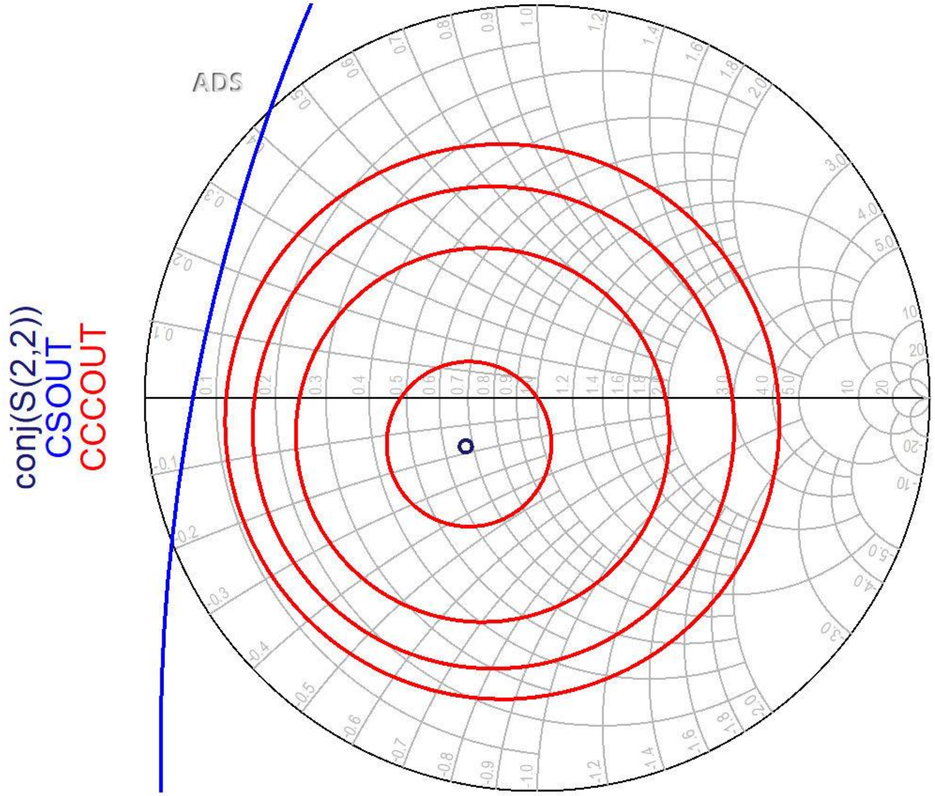


$$G_S[dB] = 10 \cdot \log\left(\frac{1-|\Gamma_S|^2}{|1-S_{11}\cdot\Gamma_S|^2}\right)$$

$$G_{S\max} = G_S\big|_{\Gamma_S = S_{11}^*}$$

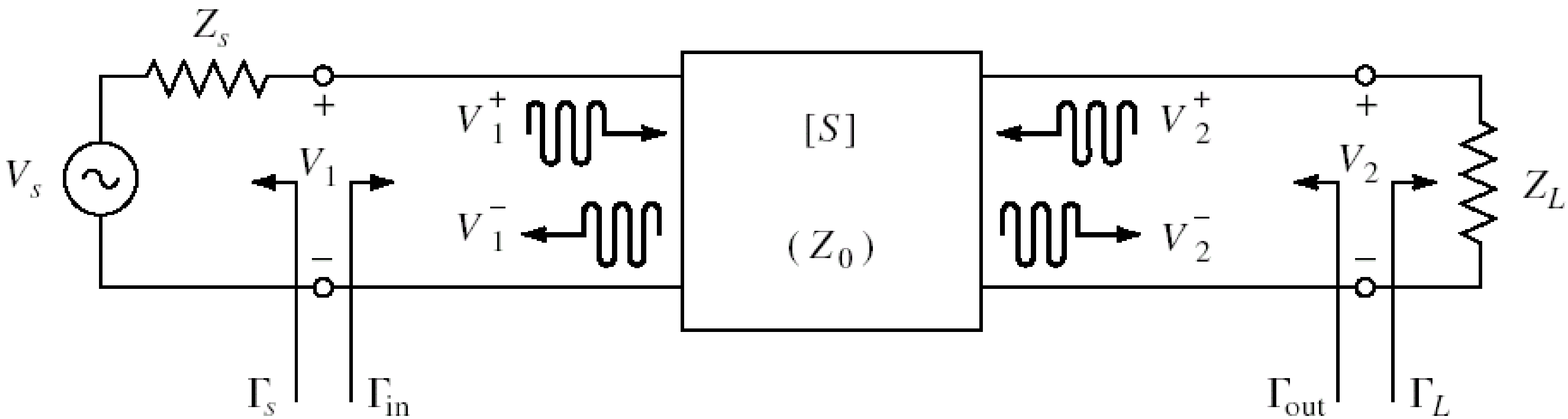
ADS





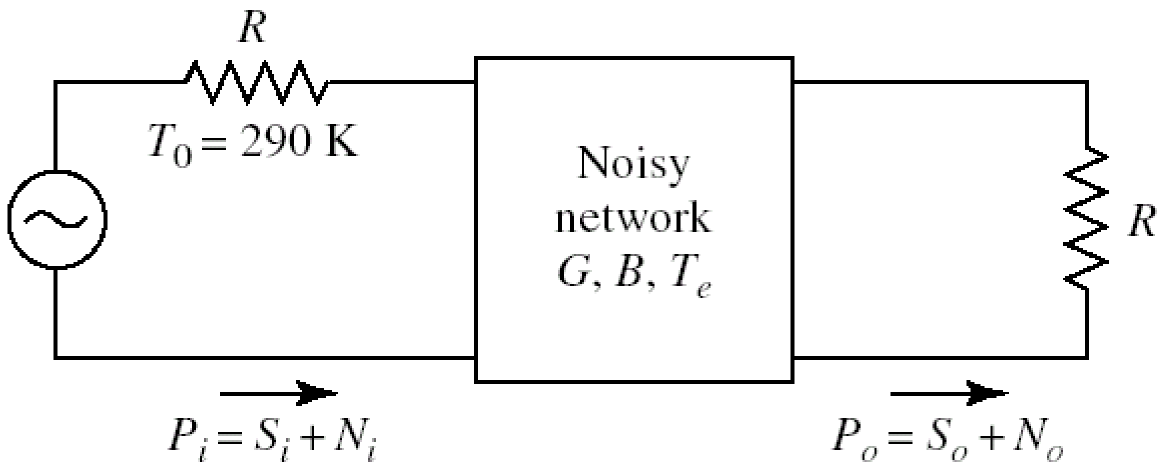
- Circles are plotted for requested values (**in dB!**)
- It is usefull to compute G_{Smax} and G_{Lmax} before
  - in order to request relevant circles

Amplifier as two-port



- For an amplifier two-port we are interested in:
  - stability
  - power gain
  - **noise** (sometimes – small signals)
  - linearity (sometimes – large signals)

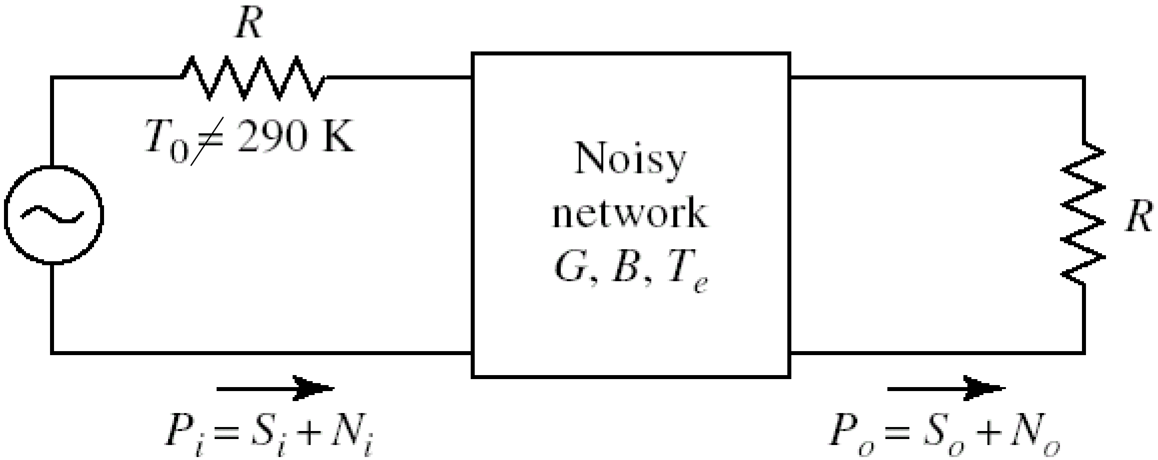
Noise Figure F



- The noise figure F, is a measure of the reduction in signal-to-noise ratio between the input and output of a device, when (by definition) the input noise power is assumed to be the noise power resulting from a matched resistor at To = 290 K (reference noise conditions)

$$F = \left. \frac{S_i / N_i}{S_o / N_o} \right|_{T_0 = 290K}$$

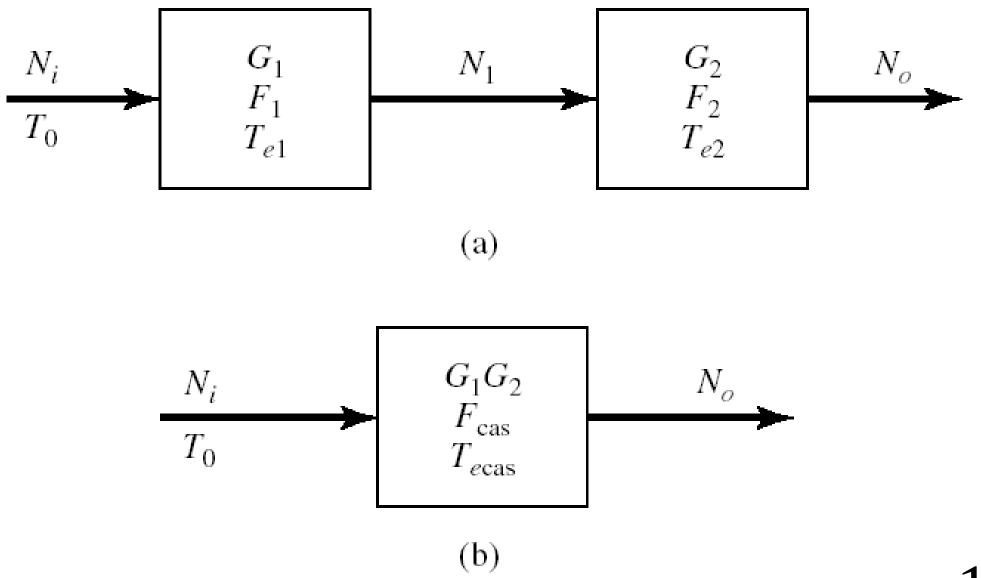
Noise Figure F



- The noise figure F, **is not** directly a measure of the reduction in signal-to-noise ratio between the input and output of a device, when the input noise power is different from that of the reference noise conditions

$$F \neq \left. \frac{S_i / N_i}{S_o / N_o} \right|_{T_0 \neq 290K}$$

Noise figure of a cascaded system

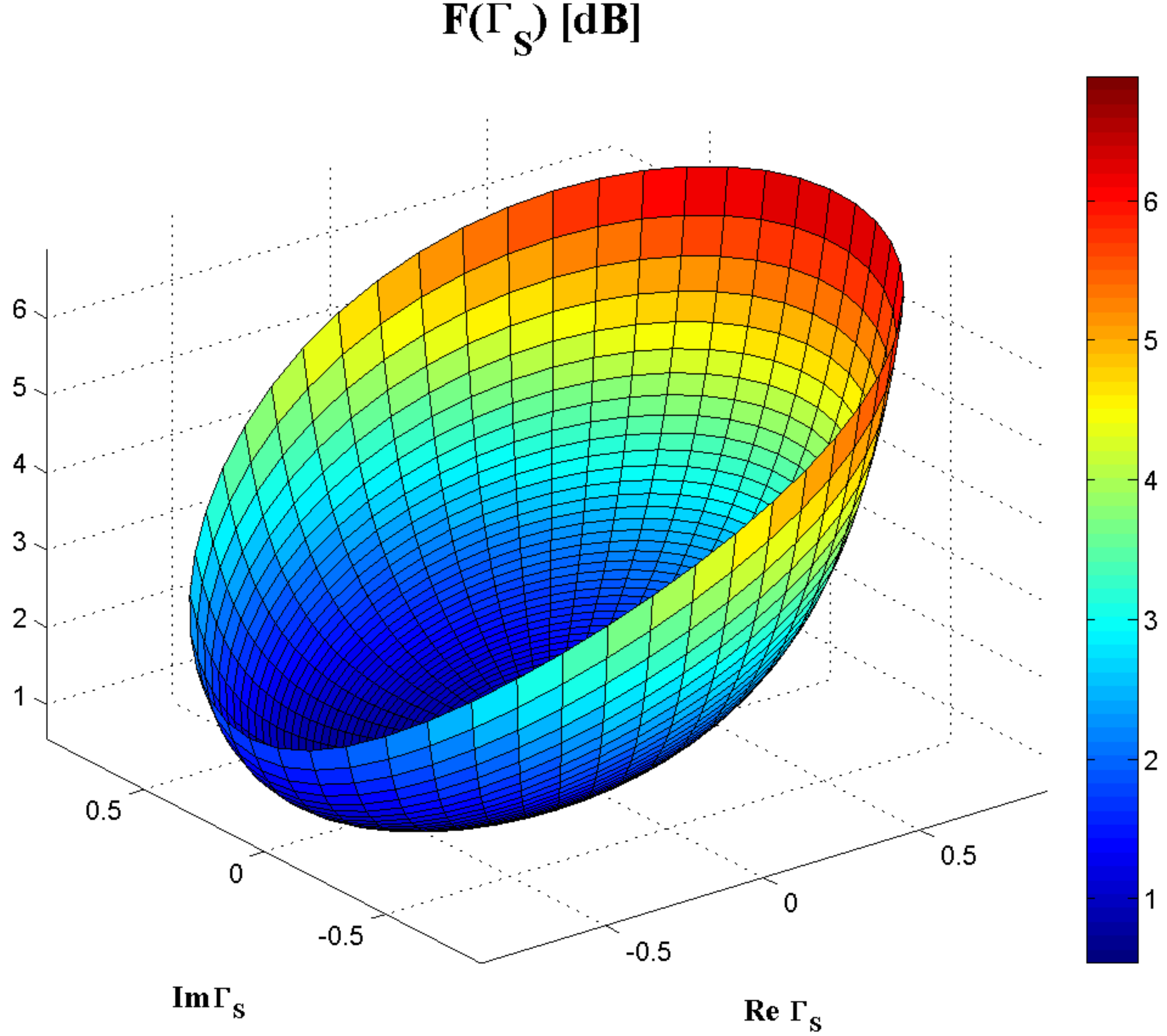


$$G_{cas} = G_1 \cdot G_2 \qquad F_{cas} = F_1 + \frac{1}{G_1}(F_2 - 1)$$

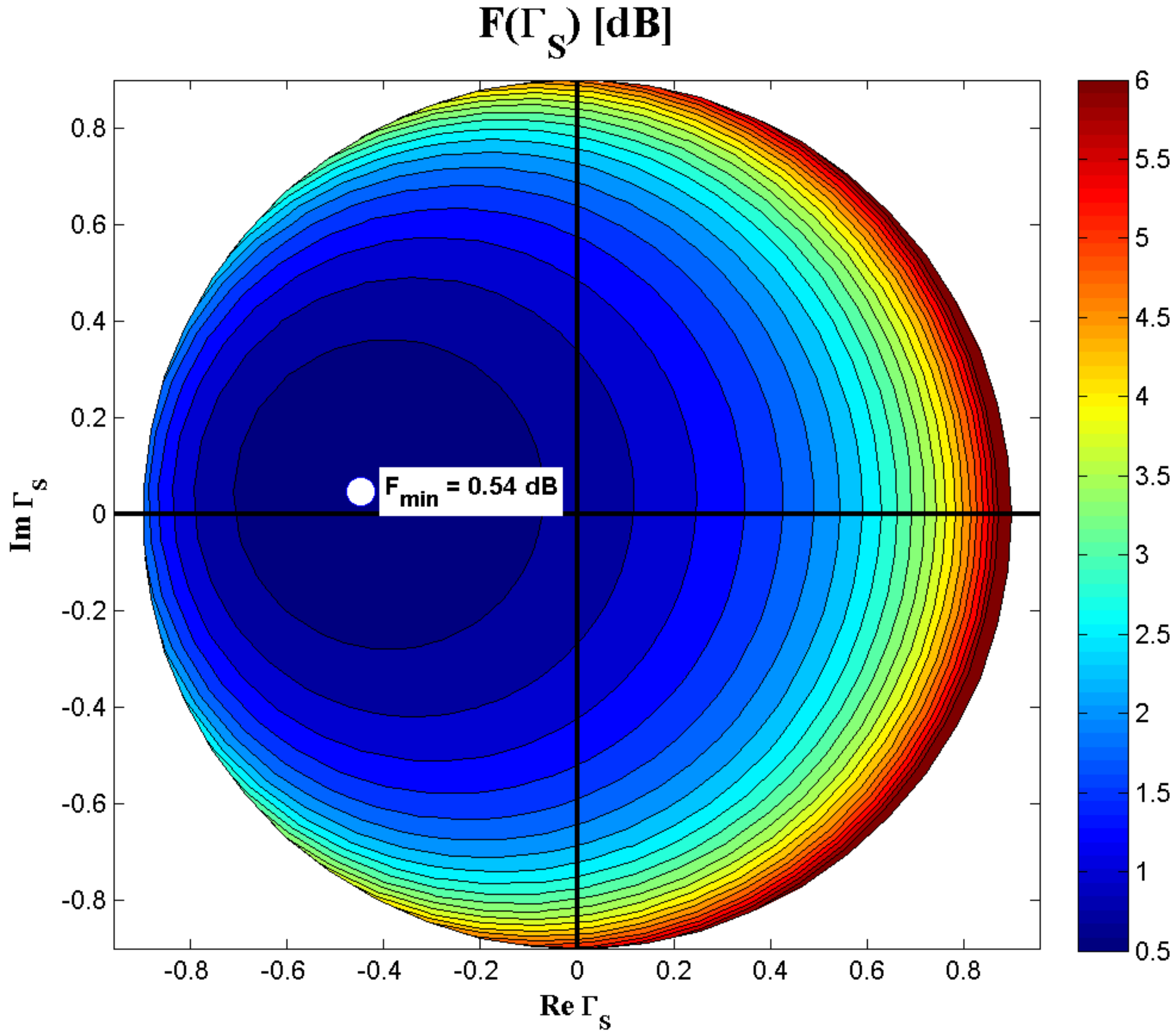
- Friis Formula (**!linear scale**)

$$F_{cas} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 \cdot G_2} + \frac{F_4 - 1}{G_1 \cdot G_2 \cdot G_3} + \cdots$$

F[dB](Γ_S)

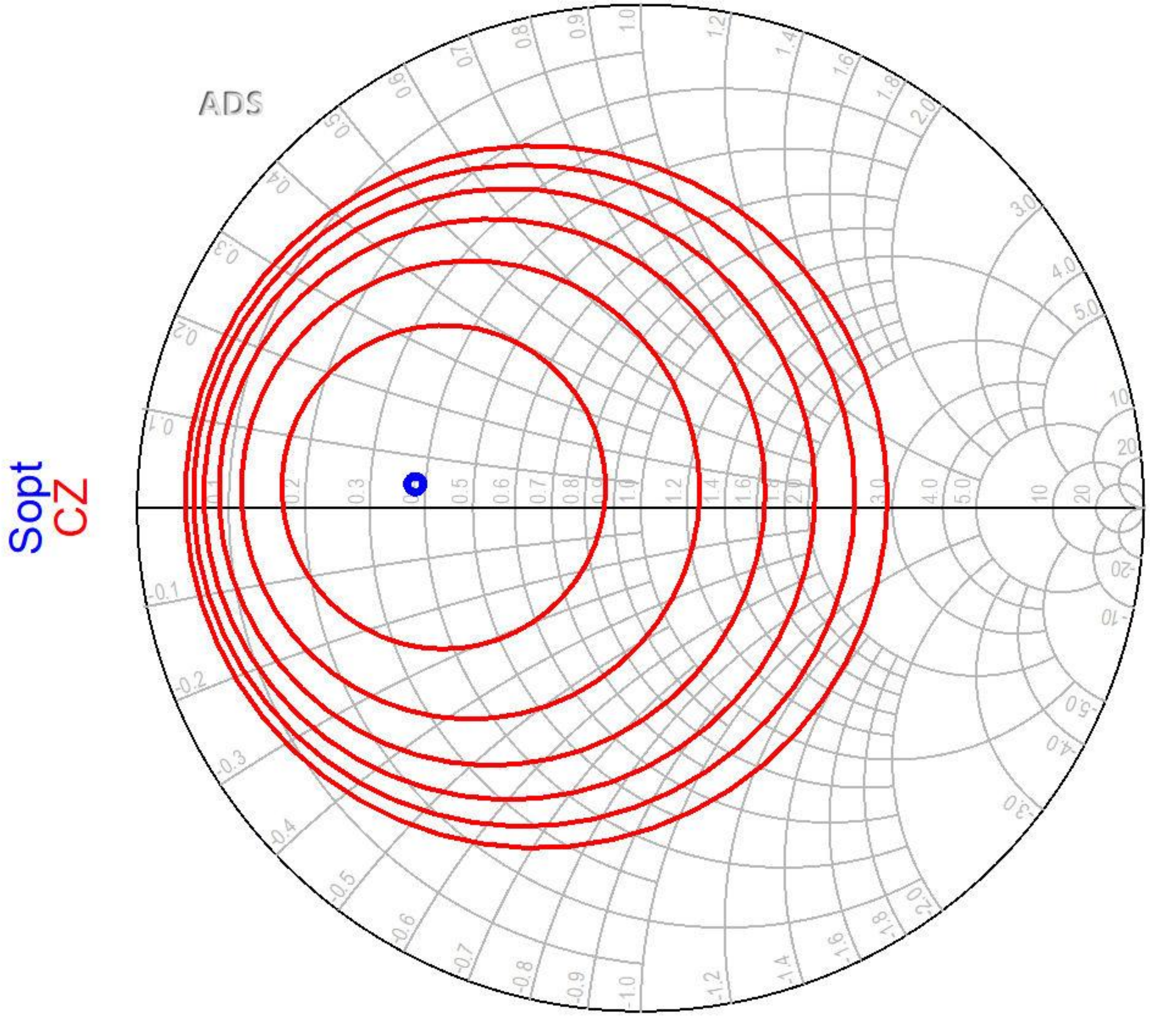


$G_S[dB](\Gamma_S)$, constant value contours



$\Gamma opt = 0.45 \angle 174°$

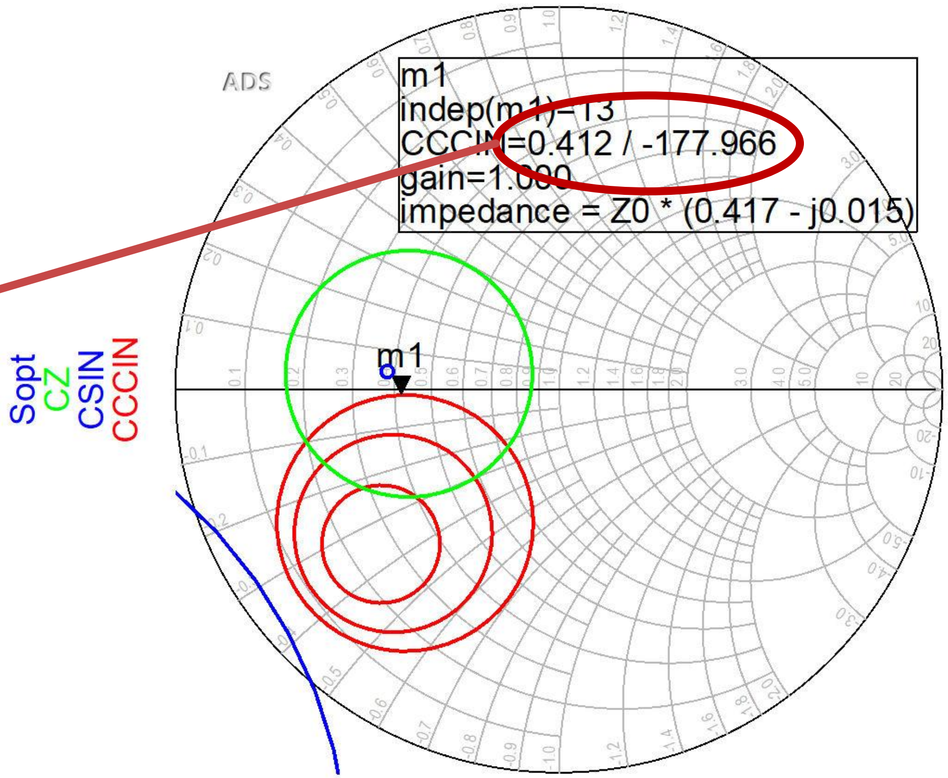
ADS



Matching – 2

- We plot on the complex plane (Smith Chart) the stability/gain/noise circles (depending on the particular application)
- We choose a point with a suitable position relative to these circles (also application dependent)
- We determine the input reflection coefficient corresponding to this point, Γ_S

$$\Gamma_S = 0.412\angle -177.966°$$

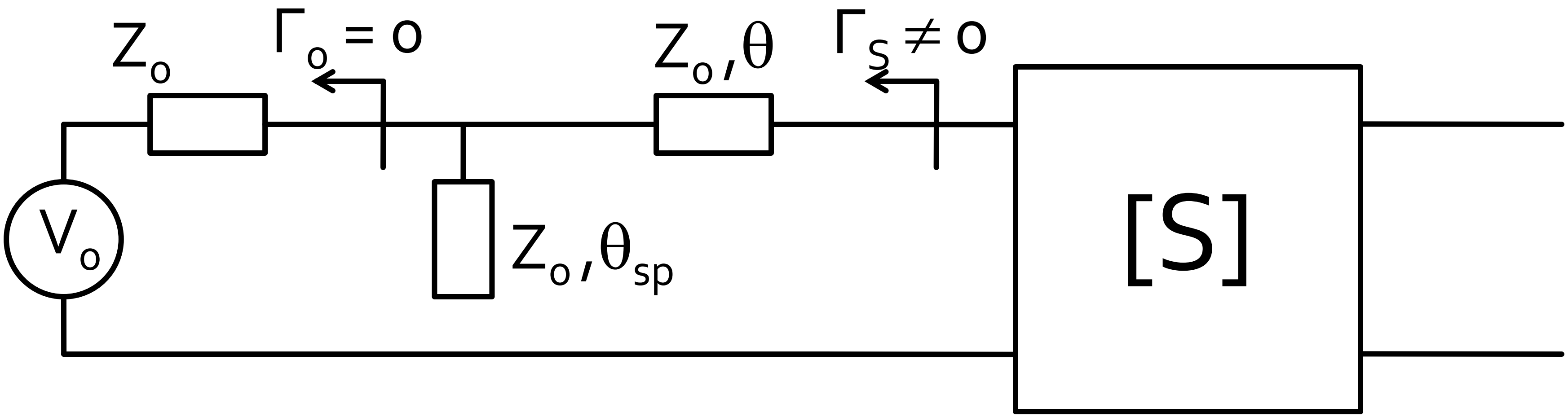


Matching – 5

- Computation depends solely on Γ_S (magnitude and phase)

$$\cos(\varphi_S + 2\theta) = -|\Gamma_S| \qquad \tan\theta_{sp} = \frac{\mp 2 \cdot |\Gamma_S|}{\sqrt{1-|\Gamma_S|^2}}$$

- The sign (+/-) chosen for the series line equation imposes the sign used for the shunt stub equation

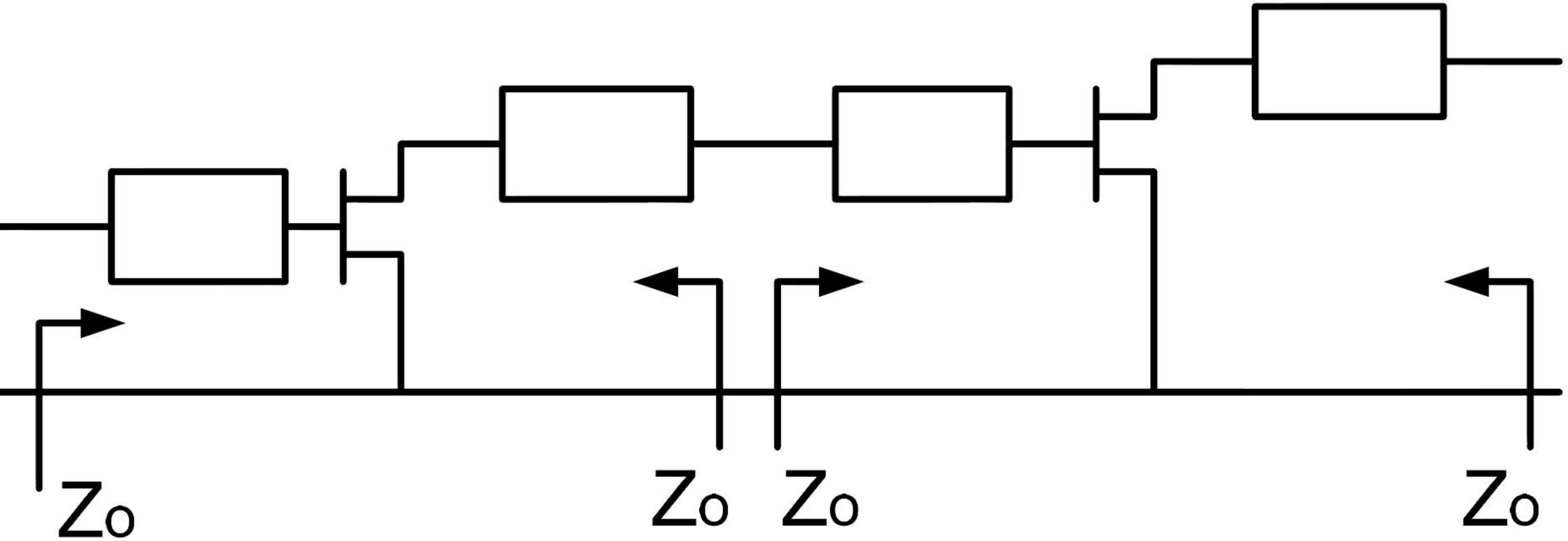


Multistage Amplifier Design

Microwave Amplifiers

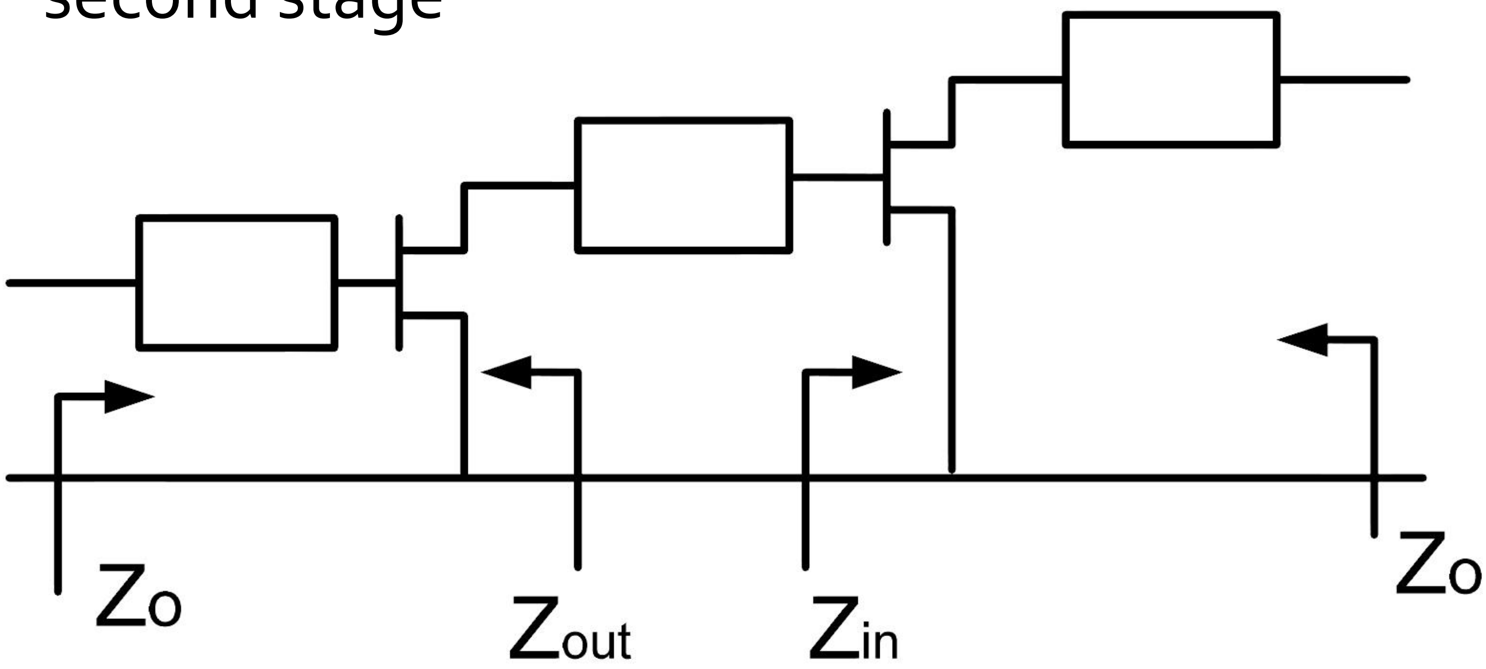
Multistage amplifiers

- Interstage matching can be designed in two modes:
  - Each stage is matched to a virtual $\Gamma = 0$



Multistage amplifiers

- Interstage matching can be designed in two modes:
  - One stage is matched to offer necessary Γ for the second stage

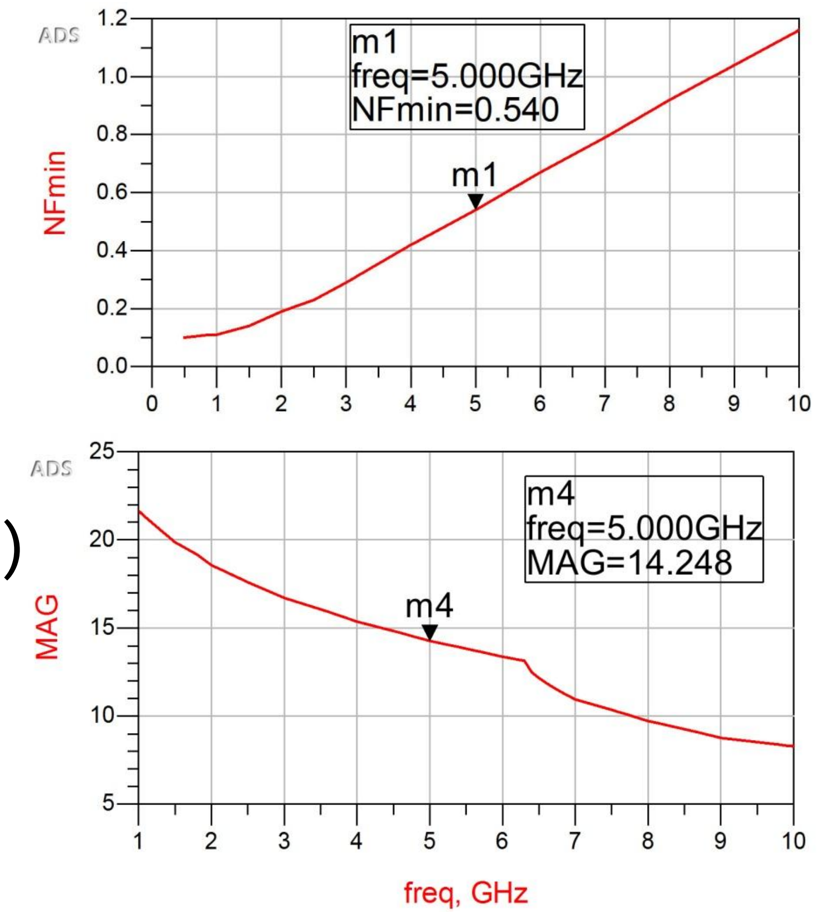


Example multistage LNA

- Similar to the homework
- LNA using ATF-34143 providing:
  - G = 20dB
  - F = 1dB
  - @f = 5GHz

Example

- ATF-34143 **at Vds=3V Id=20mA**.
- @5GHz
  - $S_{11} = 0.64\angle 139°$
  - $S_{12} = 0.119\angle -21°$
  - $S_{21} = 3.165\angle 16°$
  - $S_{22} = 0.22\angle 146°$
  - Fmin = 0.54 (**typically[dB] !**)
  - $\Gamma_{opt} = 0.45\angle 174°$
  - $r_n = 0.03$



Multistage amplifiers

- If we need more power gain than only one transistor can supply
  - necessary 20dB
  - MAG @5GHz = 14.248 dB < 20dB
- We use Friis formula to divide the necessary:
  - Power gain
  - Noise
- between two amplifier stages

Friis Formula (noise)

$$F_{cas} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 \cdot G_2} + \frac{F_4 - 1}{G_1 \cdot G_2 \cdot G_3} + \cdots$$

- Effects of Friis Formula:
  - it's essential that the first stage is as **noiseless** as possible even if that means sacrificing power
  - the second stage can be optimized for power **gain**
- Friis Formula **must** be used in **linear scale!**
- **Avago/Broadcom AppCAD**
  - AppCAD Free Design Assistant Tool for Microsoft Windows → Google

Friis Formula (noise)

$$G_{cas} = G_1 \cdot G_2 \qquad\qquad F_{cas} = F_1 + \frac{1}{G_1}(F_2 - 1)$$

- Friis formula
  - first stage: low noise factor, probably resulting in a smaller gain
  - second stage: high gain, probably resulting in higher noise factor
- It’s essential to introduce a design margin (reserve: ΔF, ΔG)
  - $G = G_{design} + \Delta G$
  - $F = F_{design} - \Delta F$
- Interpretation of the design target
  - $G > G_{design}$, better, but it’s not required to sacrifice other parameters to maximize the gain
  - $F < F_{design}$, better, the smaller the better, we must target **the smallest possible noise** factor as long as the other design parameters **are met**

Friis Formula (noise)

- Friis formula
  - first stage: low noise factor, probably resulting in a smaller gain
  - second stage: high gain, probably resulting in higher noise factor
- Division between the two stages (Estimated!)
  - input stage: F1 = 0.7 dB, G1 = 9 dB
  - output stage: F2 = 1.2 dB, G2 = 13 dB
- To verify the result apply Friis formula
- First transform to **linear scale !**

$$F_1 = 10^{\frac{F_1[dB]}{10}} = 10^{0.07} = 1.175 \qquad G_1 = 10^{\frac{G_1[dB]}{10}} = 10^{0.9} = 7.943$$

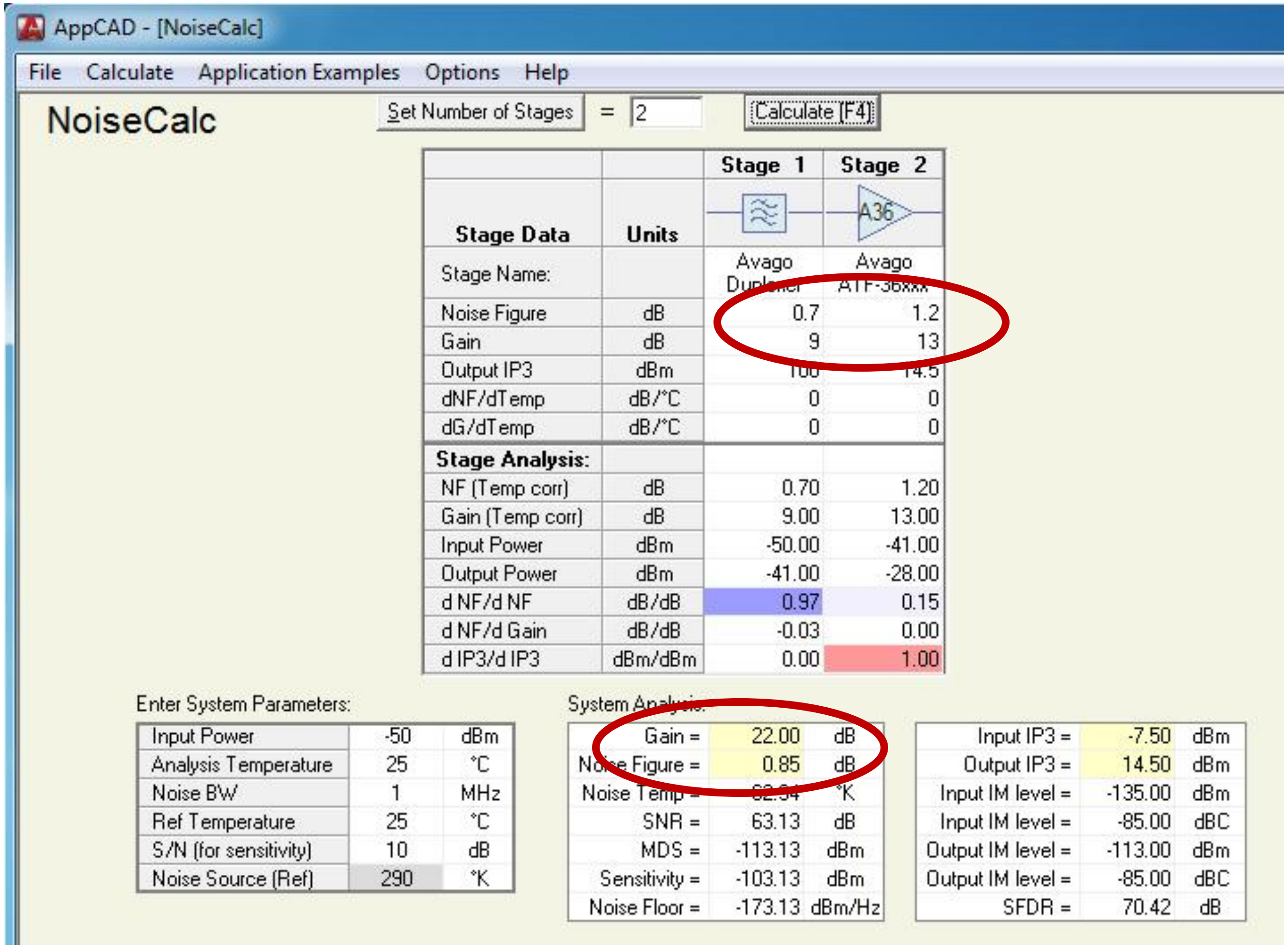
$$F_2 = 10^{\frac{F_2[dB]}{10}} = 10^{0.12} = 1.318 \qquad G_2 = 10^{\frac{G_2[dB]}{10}} = 10^{1.3} = 19.953$$

$$F_{cas} = F_1 + \frac{1}{G_1}(F_2 - 1) = 1.215 \qquad G_{cas} = G_1 \cdot G_2 = 158.49$$

$$F_{cas} = 10 \cdot \log(1.215) = 0.846\,dB \qquad G_{cas} = 10 \cdot \log(158.49) = 22\,dB$$

Friis Formula (noise)

- ## Avago/Broadcom AppCAD



AppCAD - [NoiseCalc]

File Calculate Application Examples Options Help

NoiseCalc

Set Number of Stages = 2 Calculate [F4]

Stage Data	Units	Stage 1	Stage 2
Stage Name:		Avago Duplexer	Avago ATF-36xxx
Noise Figure	dB	0.7	1.2
Gain	dB	9	13
Output IP3	dBm	100	14.5
dNF/dTemp	dB/°C	0	0
dG/dTemp	dB/°C	0	0
Stage Analysis:			
NF (Temp corr)	dB	0.70	1.20
Gain (Temp corr)	dB	9.00	13.00
Input Power	dBm	-50.00	-41.00
Output Power	dBm	-41.00	-28.00
d NF/d NF	dB/dB	0.97	0.15
d NF/d Gain	dB/dB	-0.03	0.00
d IP3/d IP3	dBm/dBm	0.00	1.00

Enter System Parameters:

Input Power	-50	dBm
Analysis Temperature	25	°C
Noise BW	1	MHz
Ref Temperature	25	°C
S/N (for sensitivity)	10	dB
Noise Source (Ref)	290	°K

System Analysis:

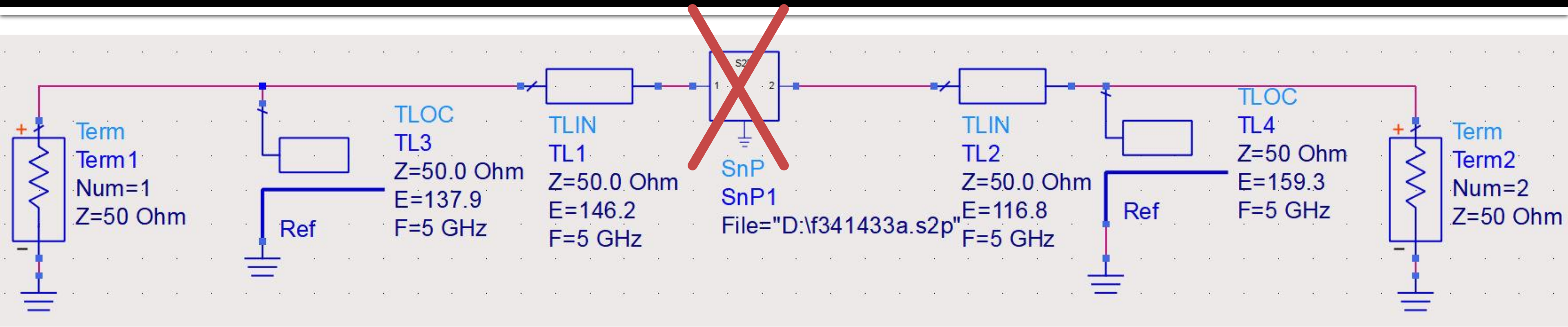
Gain =	22.00	dB
Noise Figure =	0.85	dB
Noise Temp =	[illegible]	°K
SNR =	63.13	dB
MDS =	-113.13	dBm
Sensitivity =	-103.13	dBm
Noise Floor =	-173.13	dBm/Hz

Input IP3 =	-7.50	dBm
Output IP3 =	14.50	dBm
Input IM level =	-135.00	dBm
Input IM level =	-85.00	dBC
Output IM level =	-113.00	dBm
Output IM level =	-85.00	dBC
SFDR =	70.42	dB

Multistage amplifier design

- Division between the two stages (Estimated!)
  - input stage: $F_1 = 0.7$ dB, $G_1 = 9$ dB
  - output stage: $F_2 = 1.2$ dB, $G_2 = 13$ dB
  - total: $F = 0.85$ dB, $G = 22$ dB
- Meets design specifications (with design margin)
- We can reuse the single stage LNA design (Lecture 9)
  - input matching (LNA L9) can be used for the input of the first stage – very low noise, good enough power gain
  - output matching (LNA L9) was designed for maximum gain, can be used for the output of the second stage
  - input and output matching in L9 were designed for 50Ω source and load, similar to current conditions

Input and output matching, stubs



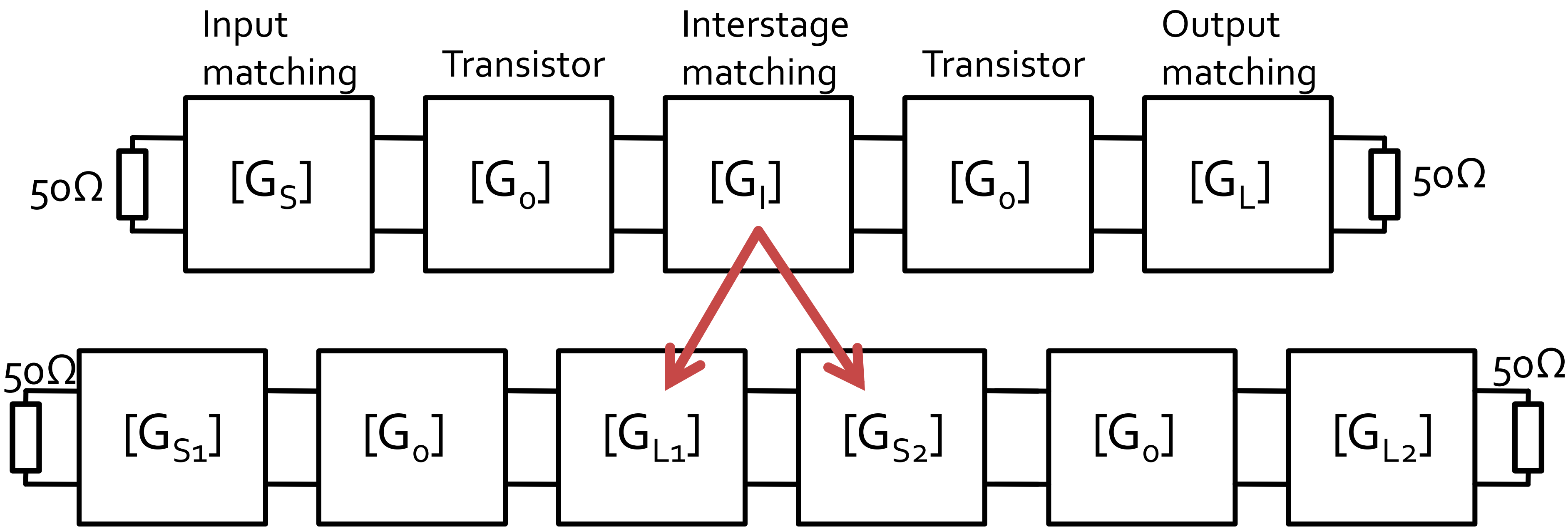
$$\cos(\varphi+2\theta) = -|\Gamma_S| \qquad \mathrm{Im}[y_S(\theta)] = \frac{\mp 2\cdot|\Gamma_S|}{\sqrt{1-|\Gamma_S|^2}}$$

$$(\varphi+2\theta) = \begin{cases} +114.33^\circ \\ -114.33^\circ \end{cases} \quad \theta = \begin{cases} 146.2^\circ \\ 31.8^\circ \end{cases} \quad \mathrm{Im}[y_S(\theta)] = \begin{cases} -0.904 \\ +0.904 \end{cases} \quad \theta_{sp} = \begin{cases} 137.9^\circ \\ 42.1^\circ \end{cases}$$

$$\cos(\varphi+2\theta) = -|\Gamma_L| \qquad \mathrm{Im}[y_L(\theta)] = \frac{-2\cdot|\Gamma_L|}{\sqrt{1-|\Gamma_L|^2}} = -0.379$$

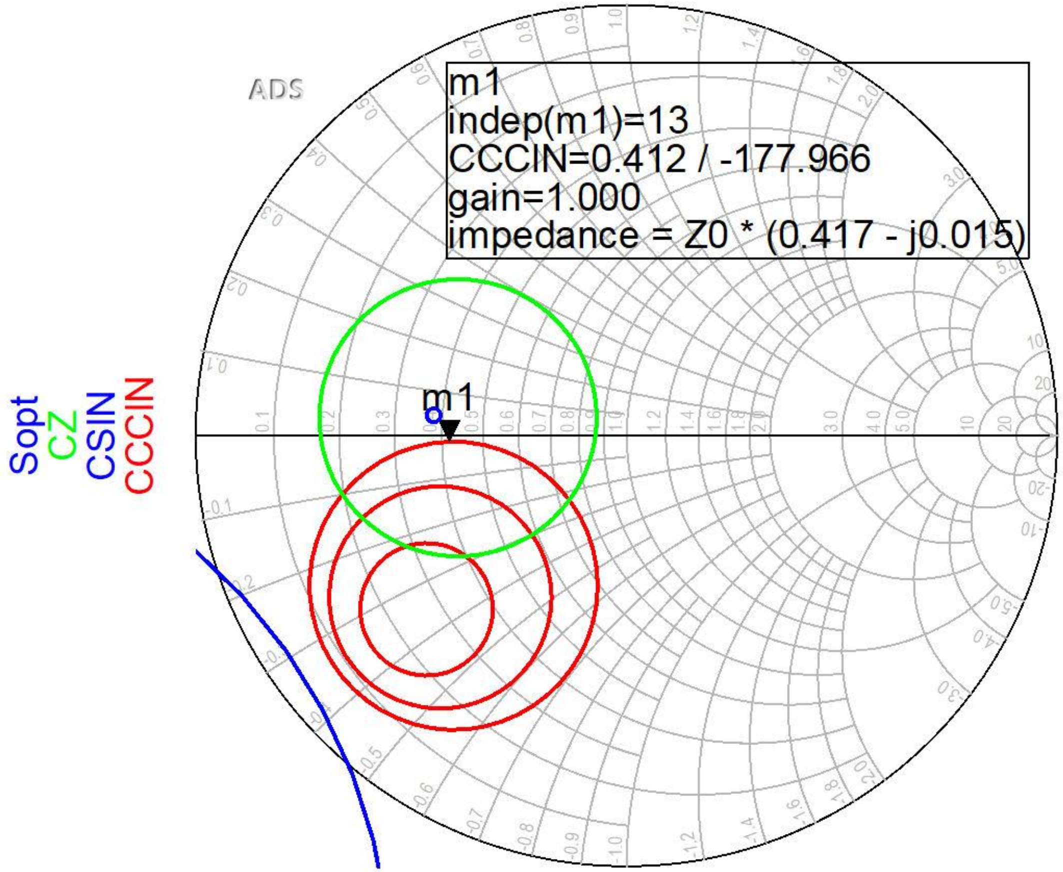
$$(\varphi+2\theta) = \begin{cases} +100.72^\circ \\ -100.72^\circ \end{cases} \quad \theta = \begin{cases} 116.8^\circ \\ 16.1^\circ \end{cases} \quad \mathrm{Im}[y_L(\theta)] = \begin{cases} -0.379 \\ +0.379 \end{cases} \quad \theta_{sp} = \begin{cases} 159.3^\circ \\ 20.7^\circ \end{cases}$$

Multistage amplifier design



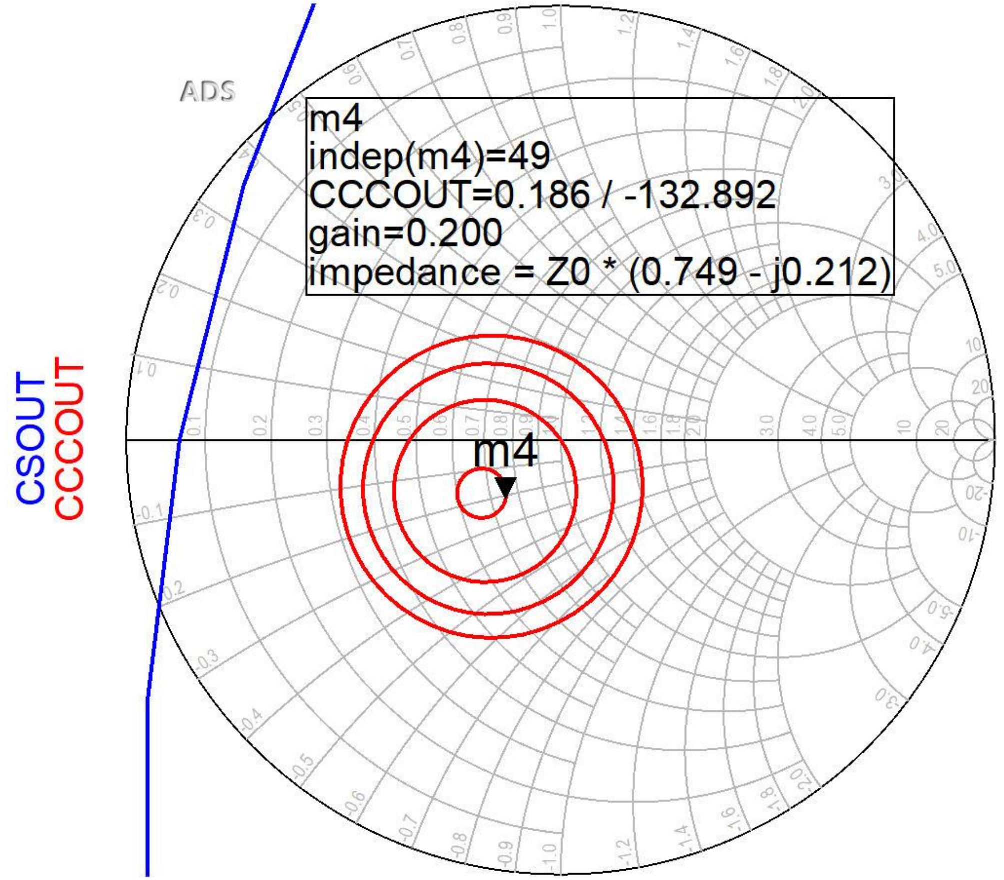
- Gain computation
  - Interstage matching can supplement the gain for both amplifier stages
  - The design for input and output matching must be achieved on a single transistor schematic (recommended: easier)

Input matching circuit (L9)



- If we can afford a 1.2dB decrease of the input gain for better NF,Q (Gs = 1 dB), position m1 above is better
- We obtain better (smaller) NF

Output matching circuit (L9)



- output constant gain circles CCCOUT: -0.4dB, -0.2dB, 0dB, +0.2dB
- The lack of noise restrictions allows optimization for better gain (close to maximum – position m4)

Proiectare etaje cascadate

- Power gain

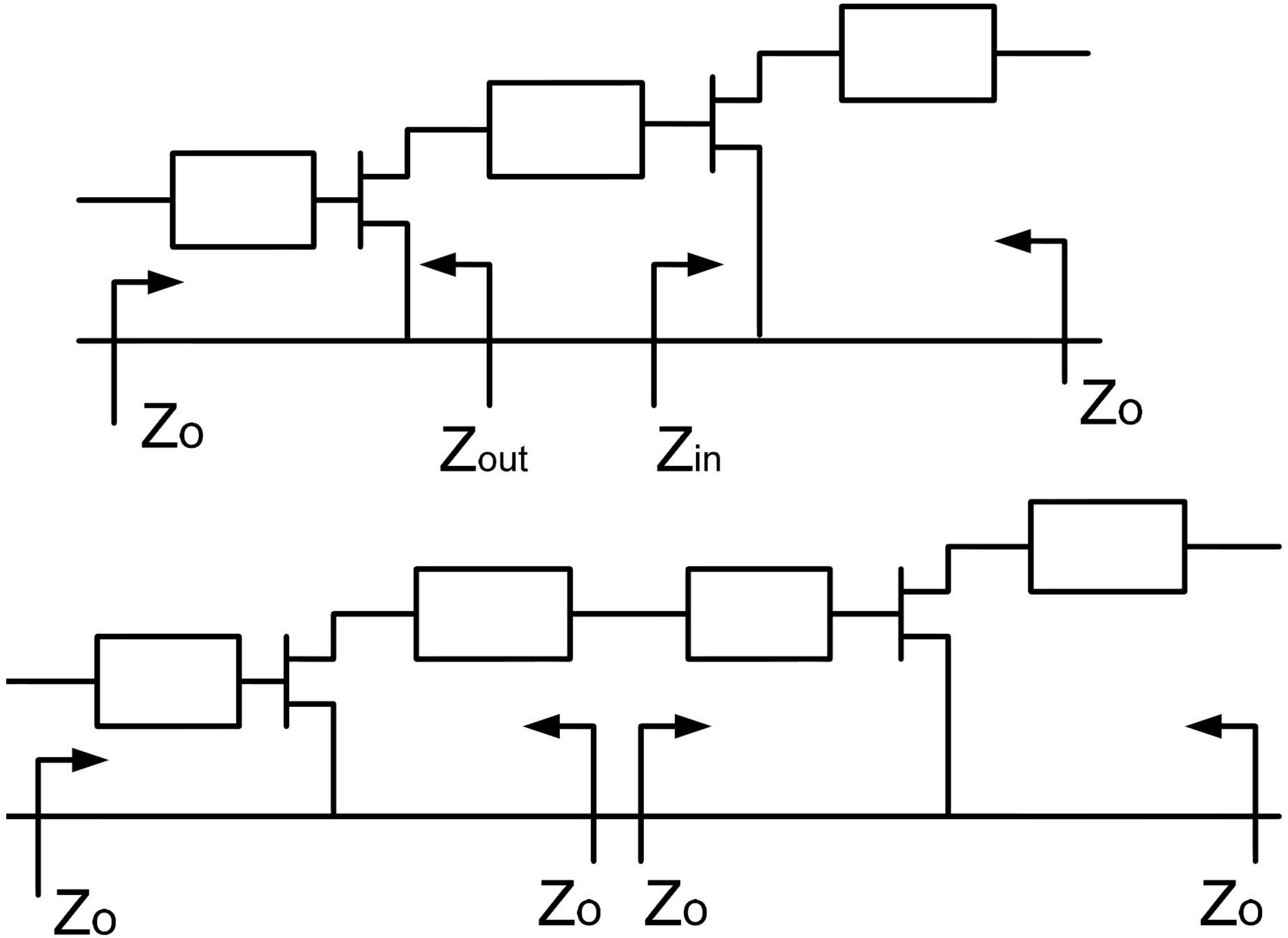
$$G_T[dB] = G_{S1}[dB] + G_0[dB] + G_{L1}[dB] + G_{S2}[dB] + G_0[dB] + G_{L2}[dB]$$

$$G_T[dB] = 1\,dB + 10\,dB + G_{L1}[dB] + G_{S2}[dB] + 10\,dB + 0.2\,dB$$

$$G_T[dB] = 21.2\,dB + G_{L1}[dB] + G_{S2}[dB]$$

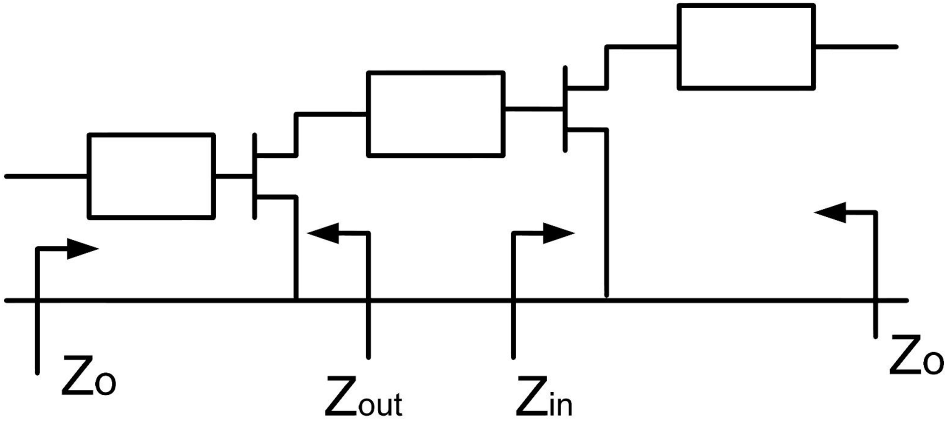
- Interstage design must provide at least 0.8dB gain to meet specifications, by better match for the output of the first transistor and for the input of the second transistor

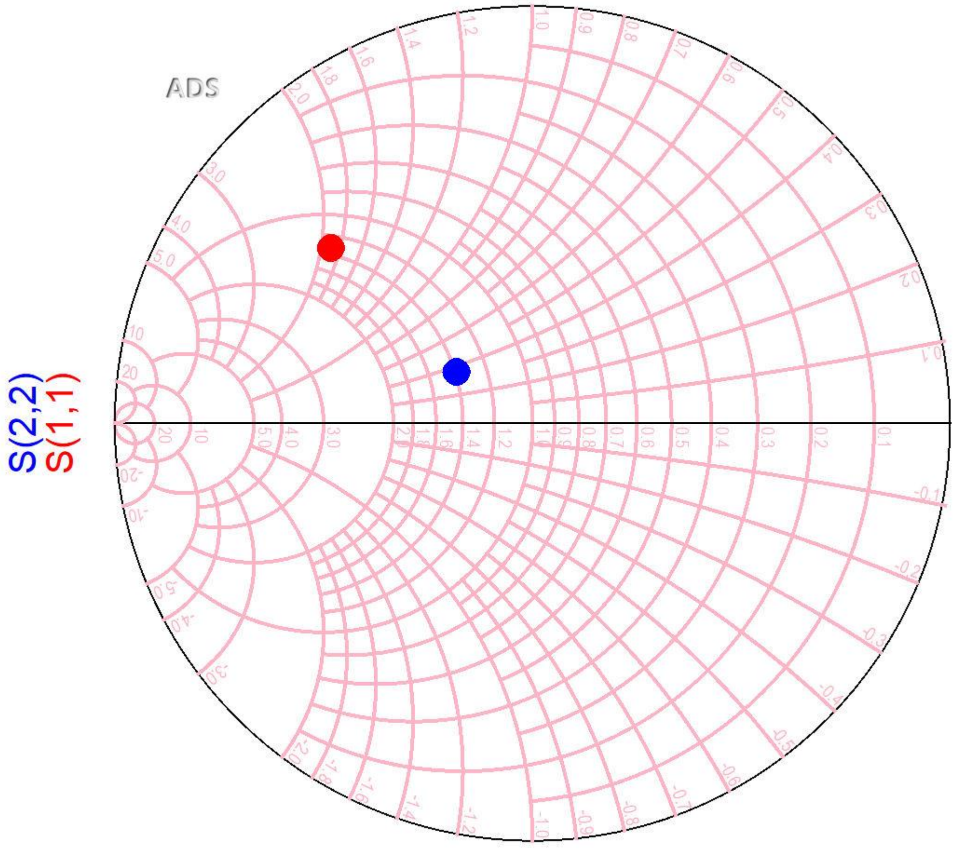
Interstage matching 1/2



Interstage matching 1

- A single transmission line keeps constant the magnitude of the reflection coefficient
  - a circle around the Smith Chart center



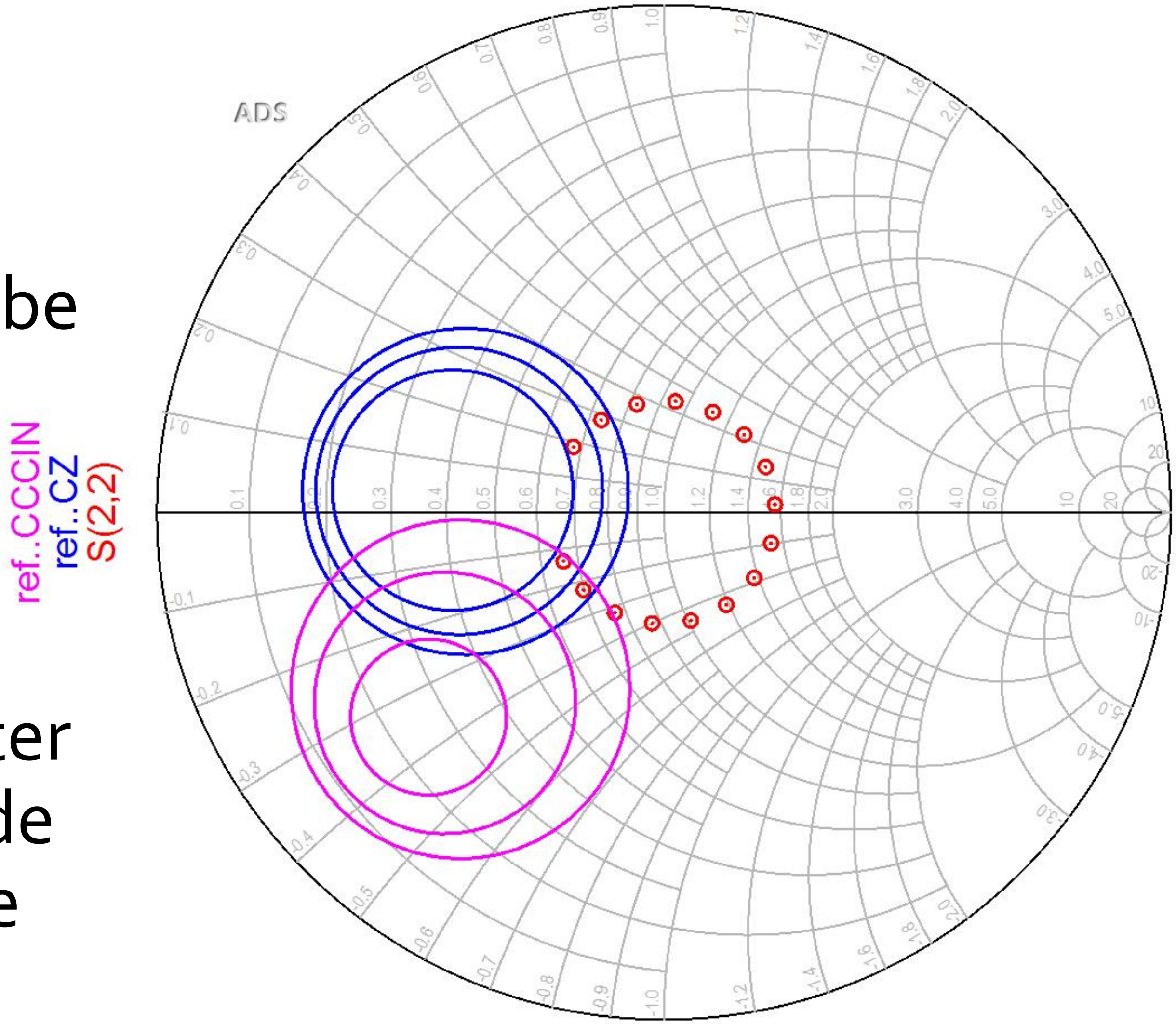


Interstage matching 1

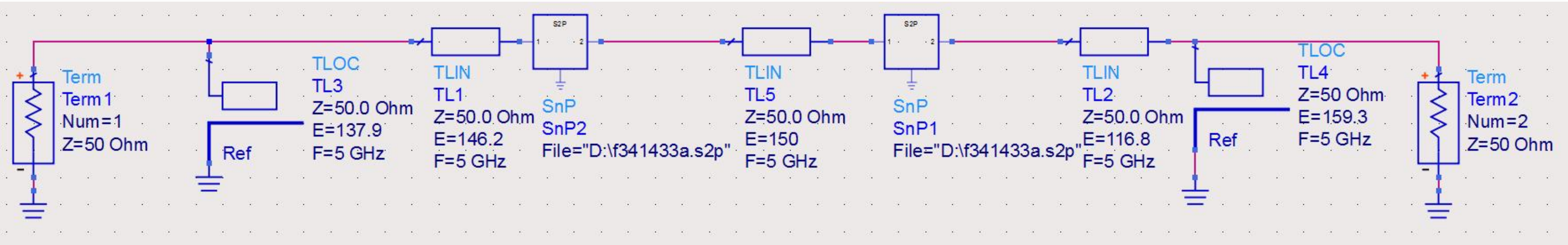
- Dual design:
  - starting from the output of the first stage (reflection coefficient S22) towards the circles (drawn for the second stage):
    - stability
    - gain
    - noise
  - starting from the input of the second stage (reflection coefficient S11) towards the circles (drawn for the first stage):
    - stability
    - gain
- First design direction has the advantage to control the noise introduced by the second stage

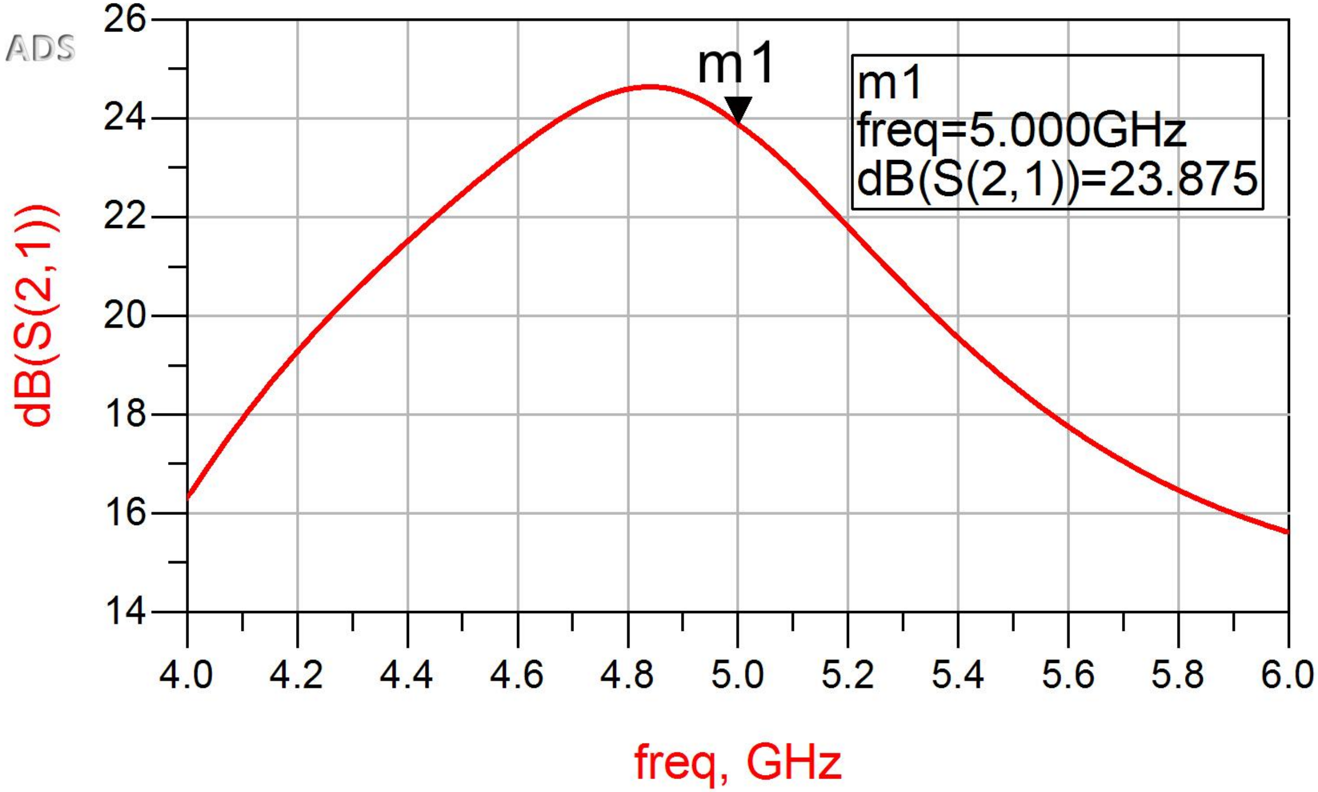
Interstage matching 1

- A **single** transmission line allows reaching a point that cannot be optimized
  - G_{L1} = 0.2dB (max)
  - G_{S2} = ~1 dB
  - F_2 = ~0.65 dB
- Only one parameter is available for wide band performance tuning

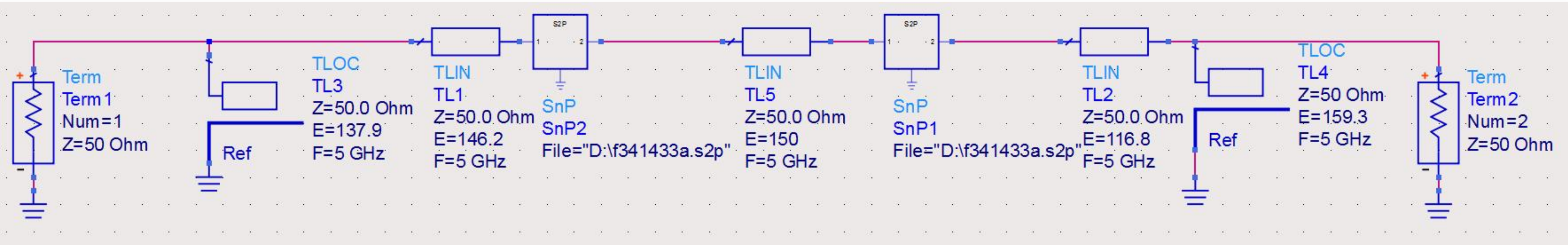


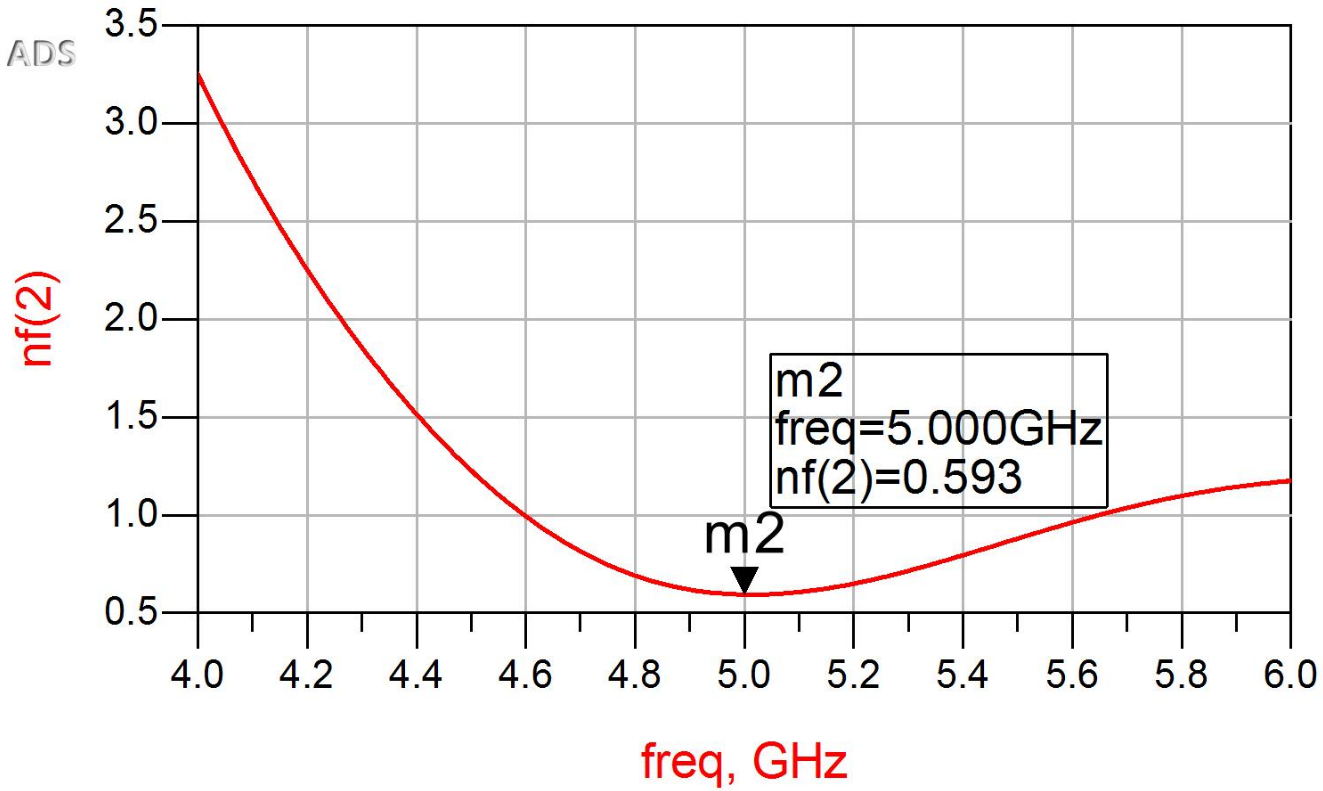
ADS


Term
Term1
Num=1
Z=50 Ohm
Ref
TLOC
TL3
Z=50.0 Ohm
E=137.9
F=5 GHz
TLIN
TL1
Z=50.0 Ohm
E=146.2
F=5 GHz
SnP
SnP2
File="D:\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=150
F=5 GHz
SnP
SnP1
File="D:\f341433a.s2p"
TLIN
TL2
Z=50.0 Ohm
E=116.8
F=5 GHz
Ref
TLOC
TL4
Z=50 Ohm
E=159.3
F=5 GHz
Term
Term2
Num=2
Z=50 Ohm


ADS
m1
m1
freq=5.000GHz
dB(S(2,1))=23.875
dB(S(2,1))
freq, GHz
26
24
22
20
18
16
14
4.0
4.2
4.4
4.6
4.8
5.0
5.2
5.4
5.6
5.8
6.0

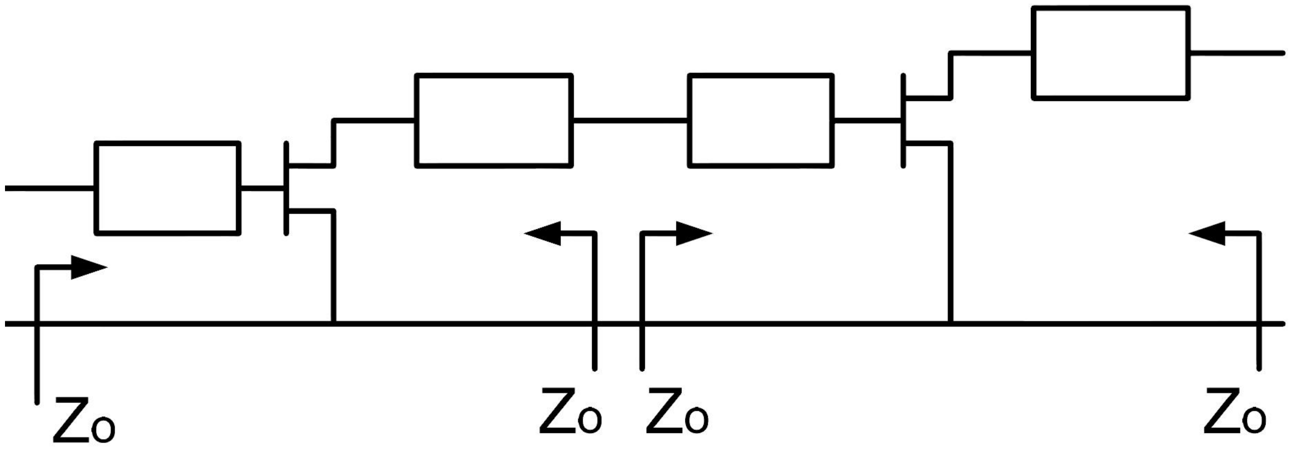
ADS


Term
Term1
Num=1
Z=50 Ohm
Ref
TLOC
TL3
Z=50.0 Ohm
E=137.9
F=5 GHz
TLIN
TL1
Z=50.0 Ohm
E=146.2
F=5 GHz
SnP
SnP2
File="D:\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=150
F=5 GHz
SnP
SnP1
File="D:\f341433a.s2p"
TLIN
TL2
Z=50.0 Ohm
E=116.8
F=5 GHz
Ref
TLOC
TL4
Z=50 Ohm
E=159.3
F=5 GHz
Term
Term2
Num=2
Z=50 Ohm


ADS
nf(2)
3.5
3.0
2.5
2.0
1.5
1.0
0.5
m2
freq=5.000GHz
nf(2)=0.593
m2
4.0
4.2
4.4
4.6
4.8
5.0
5.2
5.4
5.6
5.8
6.0
freq, GHz

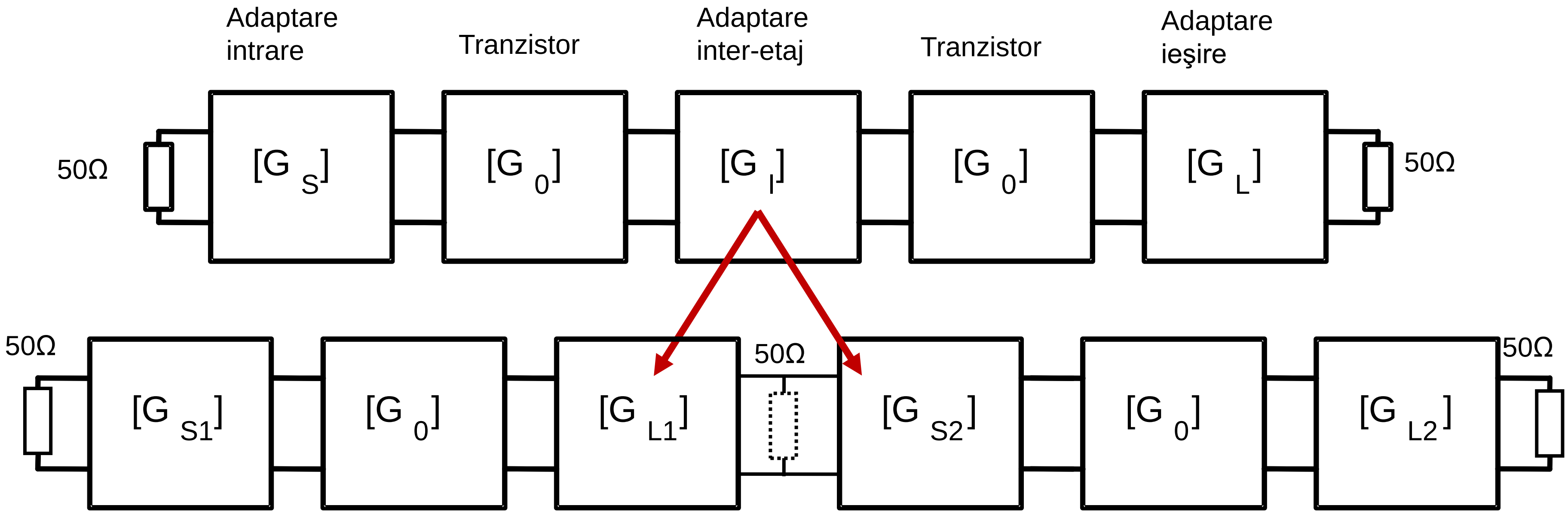
Interstage matching 2

- Using multiple transmission lines for matching each stage to a intermediate Γ=0 (virtual) allows detailed control over final reflection coefficient



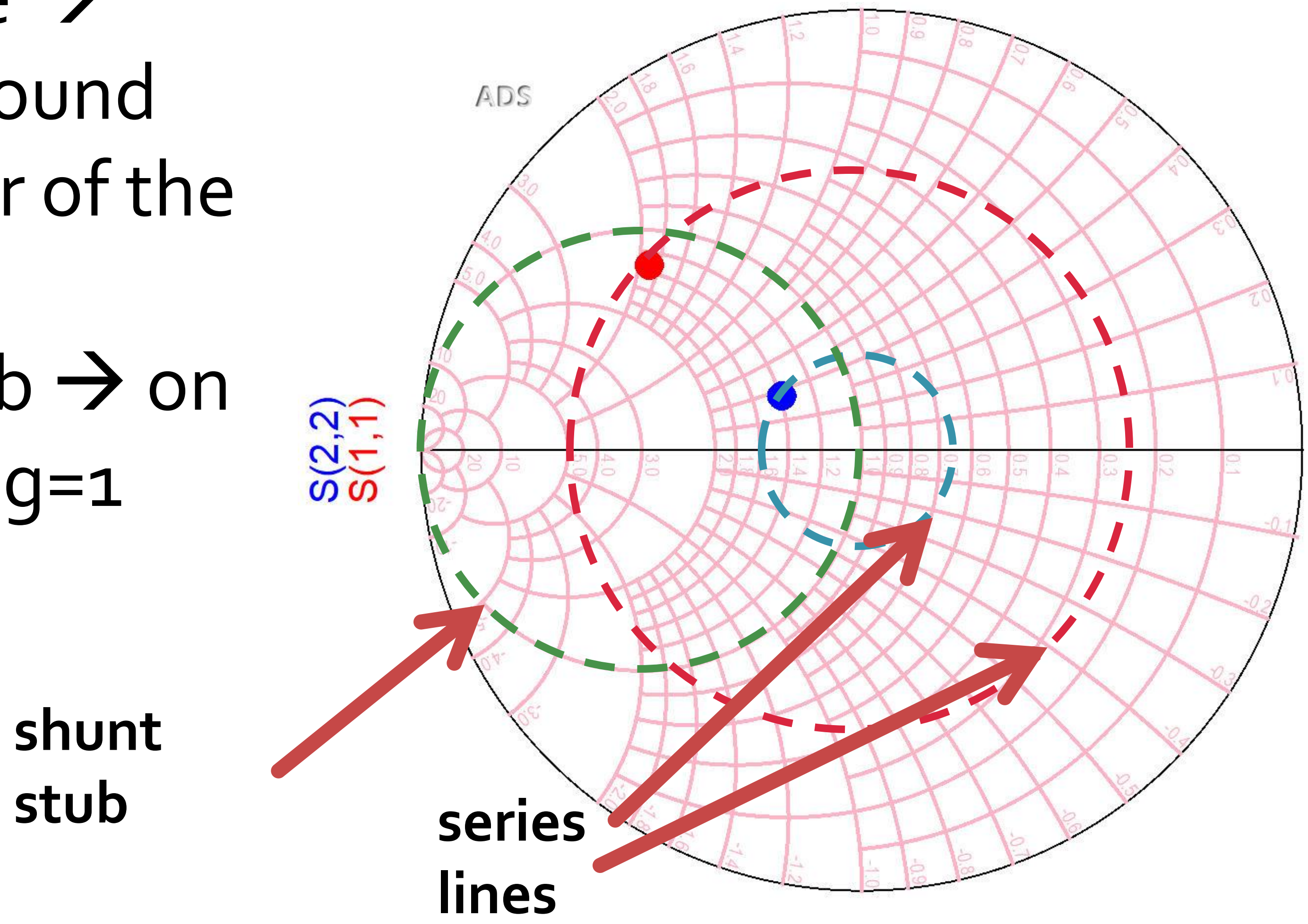
Interstage matching 2

- Using multiple transmission lines for matching each stage to a intermediate Γ=0 (virtual) allows detailed control over final reflection coefficient



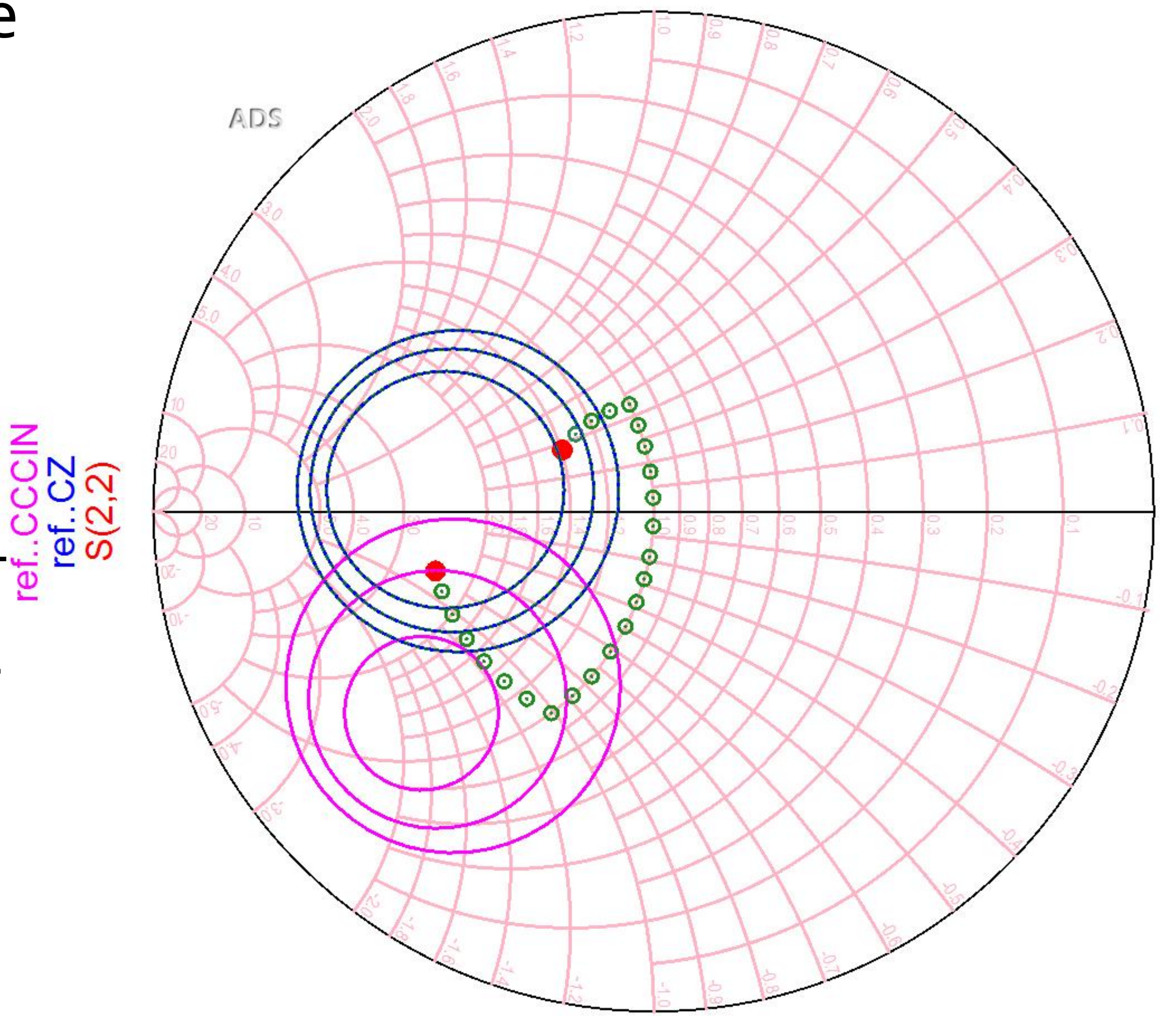
Smith Chart

- series line → moves around the center of the SC
- shunt stub → on the circle g=1



Adaptare inter-etaje

- For every stage we use a series line and a shunt stub
  - the series line moves the reflection coefficient from the desired starting point on the unity conductance circle g=1
  - the shunt stub moves the point to the center of the Smith Chart (Zo match)
- The two shunt stubs combine into one



Calcul analitic

- G_{L1} (plecare din S22 spre origine)

$$S_{22} = 0.22\angle 146^\circ \qquad |S_{22}| = 0.22; \quad \varphi = 146^\circ$$

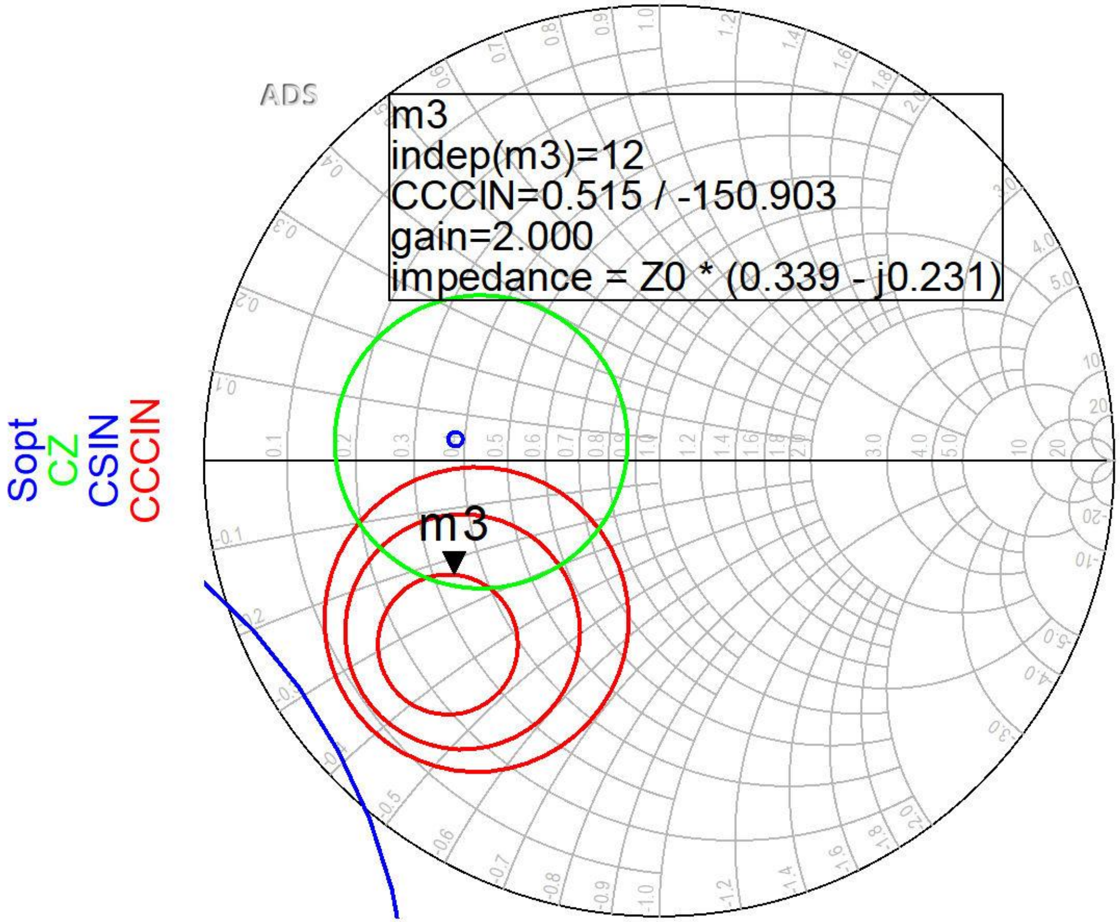
$$\cos(\varphi + 2\theta) = -|S_{22}| \qquad \mathrm{Im}[y_{L1}(\theta)] = \frac{\mp 2 \cdot |S_{22}|}{\sqrt{1 - |S_{22}|^2}}$$

$$\cos(\varphi + 2\theta) = -0.22 \Rightarrow \quad (\varphi + 2\theta) = \pm 102.71^\circ$$

$$(\varphi + 2\theta) = \begin{cases} +102.71^\circ \\ -102.71^\circ \end{cases} \quad \theta = \begin{cases} 158.4^\circ \\ 55.6^\circ \end{cases} \quad \mathrm{Im}[y_{L1}(\theta)] = \begin{cases} -0.451 \\ +0.451 \end{cases} \quad \theta_{sp} = \begin{cases} 155.7^\circ \\ 24.3^\circ \end{cases}$$

Calcul analitic

- G_{S2} (plecare din Γ_{S2} ales spre origine – castig 2dB)



Calcul analitic

- G_{S_2} (plecare din 2 spre origine)

$$\Gamma_{S2} = 0.515\angle -150.9^\circ \qquad |\Gamma_{S2}| = 0.515; \quad \varphi = -150.9^\circ$$

$$\cos(\varphi + 2\theta) = -|\Gamma_{S2}| \qquad \mathrm{Im}[y_{S2}(\theta)] = \frac{\mp 2\cdot|\Gamma_{S2}|}{\sqrt{1-|\Gamma_{S2}|^2}}$$

$$\cos(\varphi + 2\theta) = -0.515 \Rightarrow \qquad (\varphi + 2\theta) = \pm 121^\circ$$

$$(\varphi + 2\theta) = \begin{cases} +121^\circ \\ -121^\circ \end{cases} \quad \theta = \begin{cases} 135.9^\circ \\ 15^\circ \end{cases} \quad \mathrm{Im}[y_{S2}(\theta)] = \begin{cases} -1.202 \\ +1.202 \end{cases} \quad \theta_{sp} = \begin{cases} 129.8^\circ \\ 50.2^\circ \end{cases}$$

Calcul analitic

- **Cele doua stub-uri in gol se combina intr-unul singur**
- Exista **4 combinatii posibile** in functie de cum se combina lungimile electrice ale celor doua linii serie
  - pentru fiecare lungime electrica aleasa (θ) se foloseste obligatoriu Im[y(θ)] corespunzator
- Ex:

$$\theta_{L1} = 158.4° \qquad \theta_{S2} = 135.9° \qquad \mathrm{Im}[y_{sp}] = \mathrm{Im}[y_{L1}(\theta)] + \mathrm{Im}[y_{S2}(\theta)] = -1.653$$

$$\theta_{sp} = \tan^{-1}(\mathrm{Im}[y_{sp}]) \qquad \theta_{sp} = 121.2°$$

Diagrama Smith

- linie serie → pe cercul cu centrul in originea DS
- stub paralel → pe cercul g=1

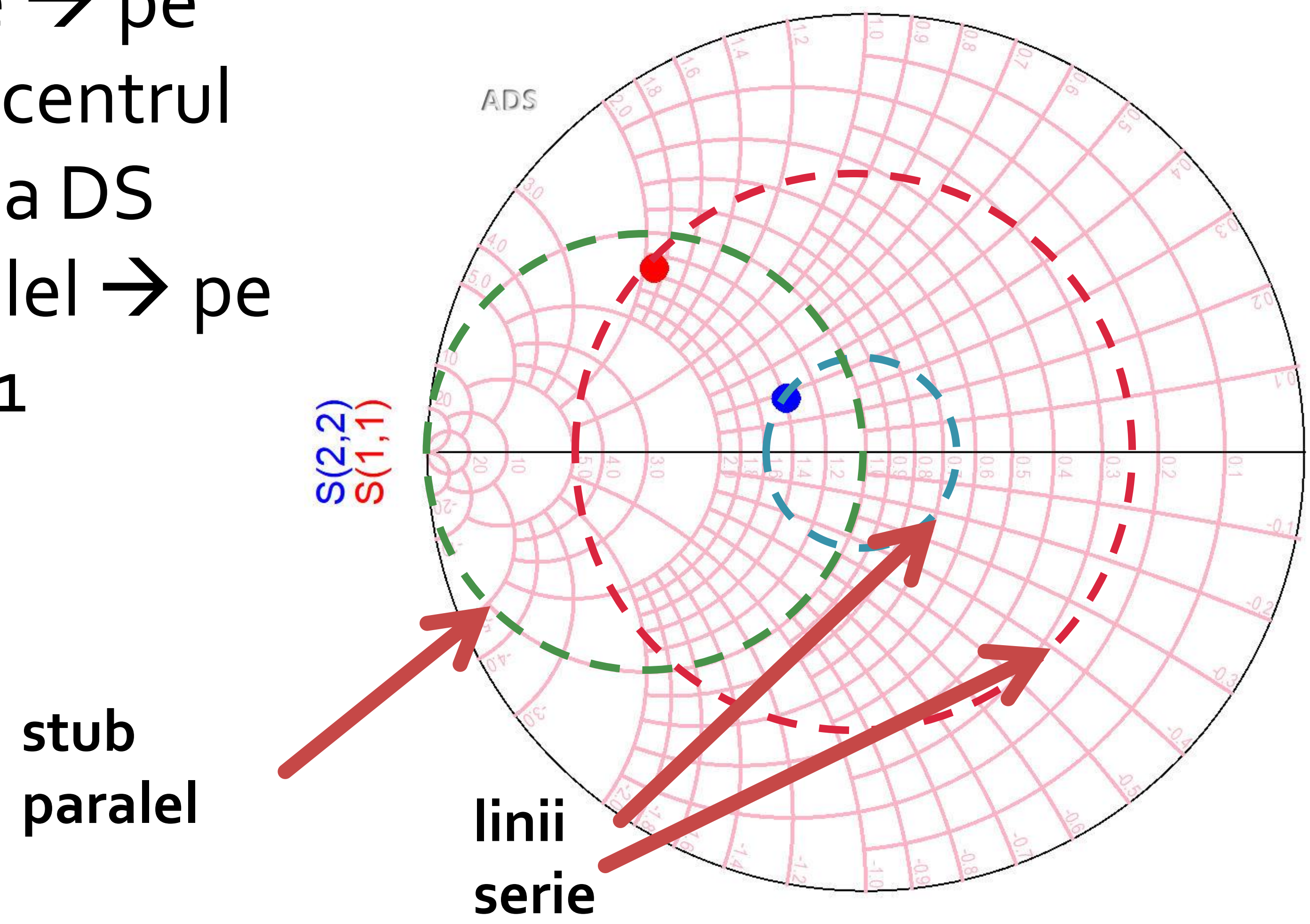


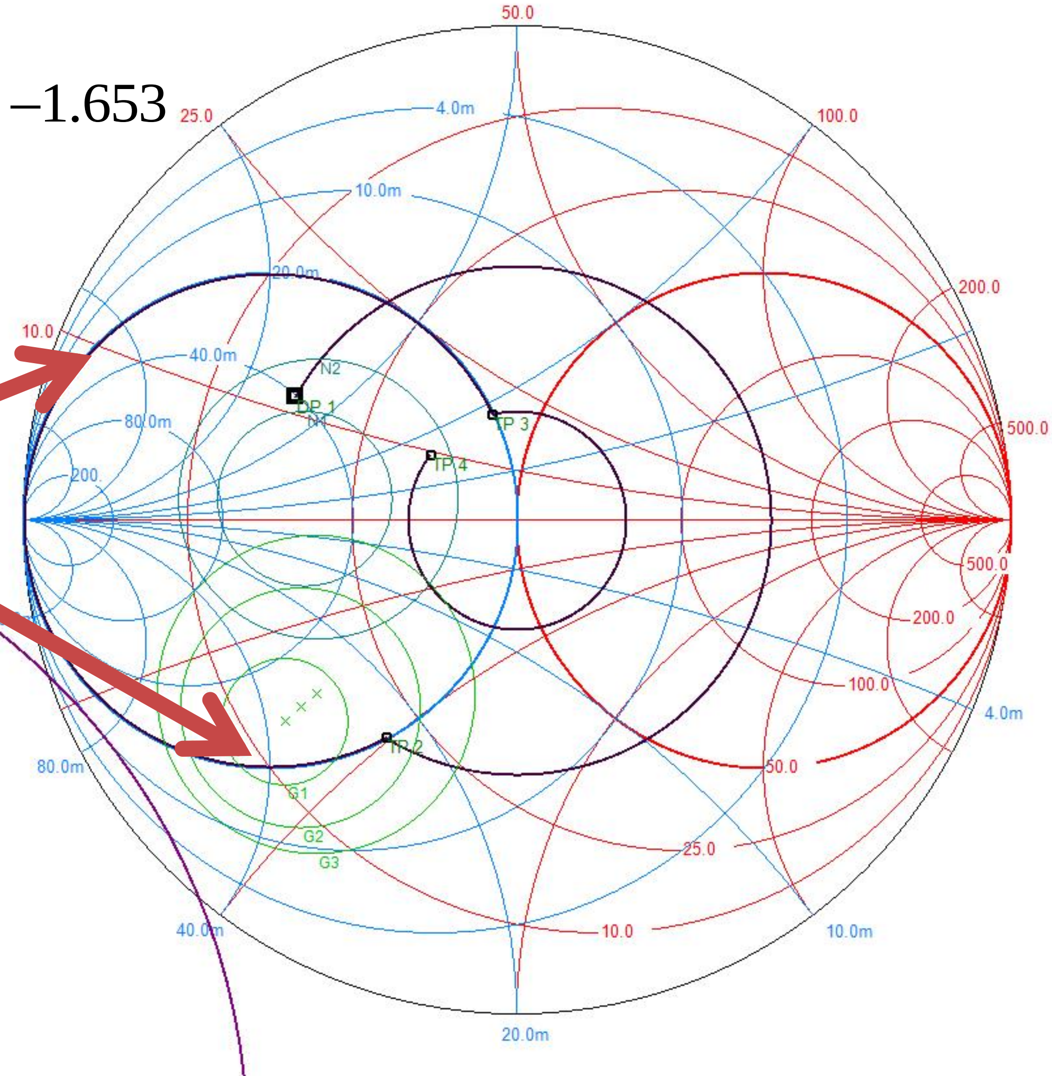
Diagrama Smith 1

$\theta_{L1} = 158.4°$ $\theta_{S2} = 135.9°$

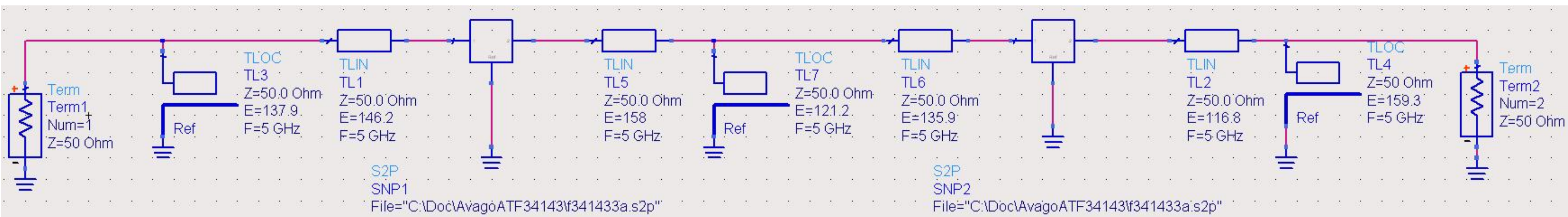
$$\mathrm{Im}[y_{sp}] = \mathrm{Im}[y_{L1}(\theta)] + \mathrm{Im}[y_{S2}(\theta)] = -1.653$$

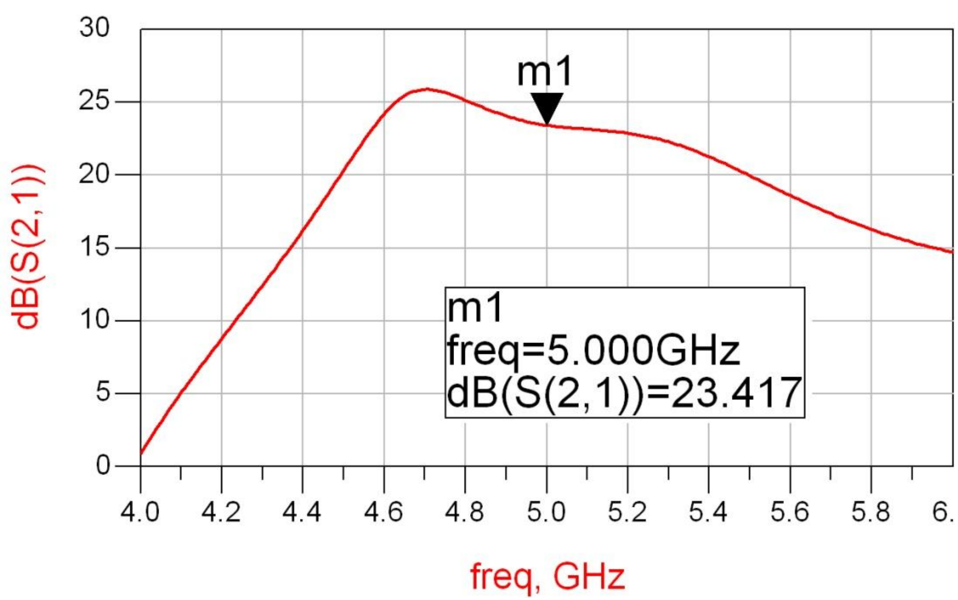
$\theta_{sp} = 121.2°$

stub "combinat"



ADS 1


Term
Term1
Num=1
Z=50 Ohm
Ref
TLOC
TL3
Z=50.0 Ohm
E=137.9
F=5 GHz
TLIN
TL1
Z=50.0 Ohm
E=146.2
F=5 GHz
S2P
SNP1
File="C:\Doc\AvagoATF34143\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=158
F=5 GHz
Ref
TLOC
TL7
Z=50.0 Ohm
E=121.2
F=5 GHz
TLIN
TL6
Z=50.0 Ohm
E=135.9
F=5 GHz
S2P
SNP2
File="C:\Doc\AvagoATF34143\f341433a.s2p"
TLIN
TL2
Z=50.0 Ohm
E=116.8
F=5 GHz
Ref
TLOC
TL4
Z=50 Ohm
E=159.3
F=5 GHz
Term
Term2
Num=2
Z=50 Ohm


m1
m1
freq=5.000GHz
dB(S(2,1))=23.417
dB(S(2,1))
30
25
20
15
10
5
0
4.0
4.2
4.4
4.6
4.8
5.0
5.2
5.4
5.6
5.8
6.
freq, GHz

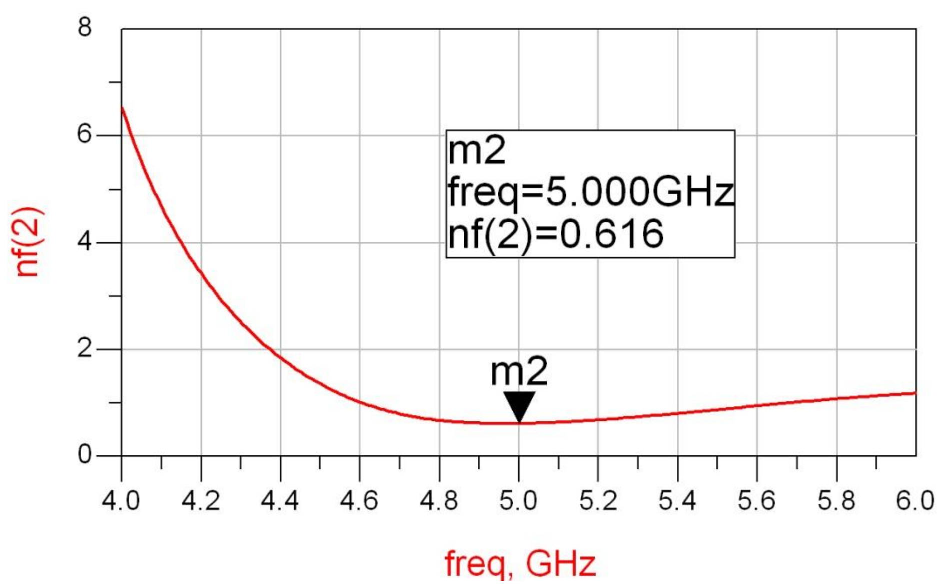

m2
freq=5.000GHz
nf(2)=0.616
m2
nf(2)
8
6
4
2
0
4.0
4.2
4.4
4.6
4.8
5.0
5.2
5.4
5.6
5.8
6.0
freq, GHz

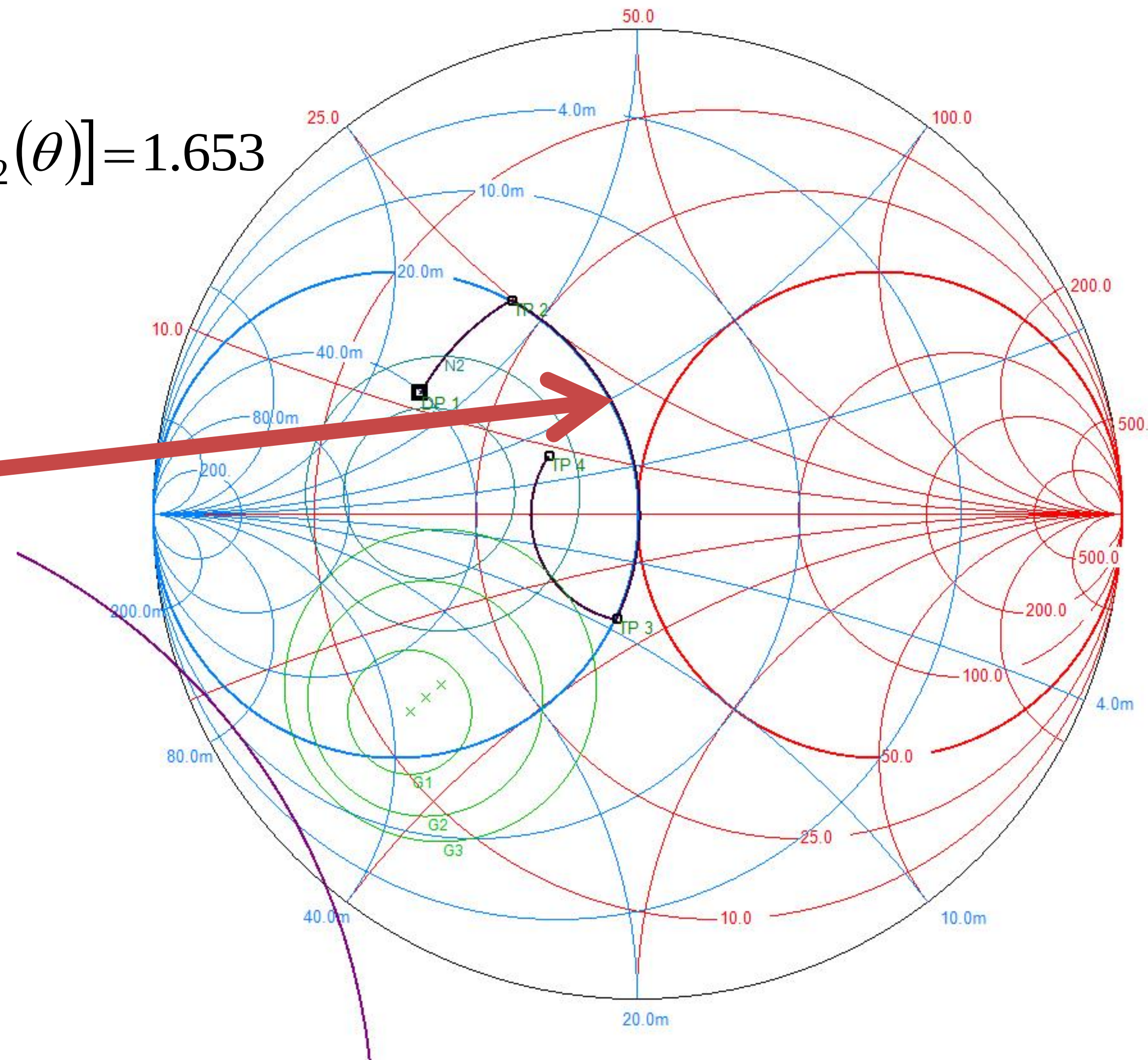
Diagrama Smith 2

$$\theta_{L1} = 55.6^\circ \qquad \theta_{S2} = 15^\circ$$

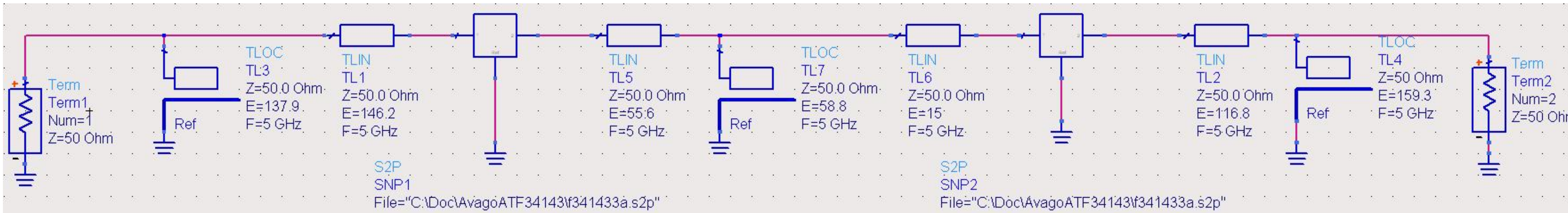
$$\mathrm{Im}[y_{sp}] = \mathrm{Im}[y_{L1}(\theta)] + \mathrm{Im}[y_{S2}(\theta)] = 1.653$$

$$\theta_{sp} = 58.8^\circ$$

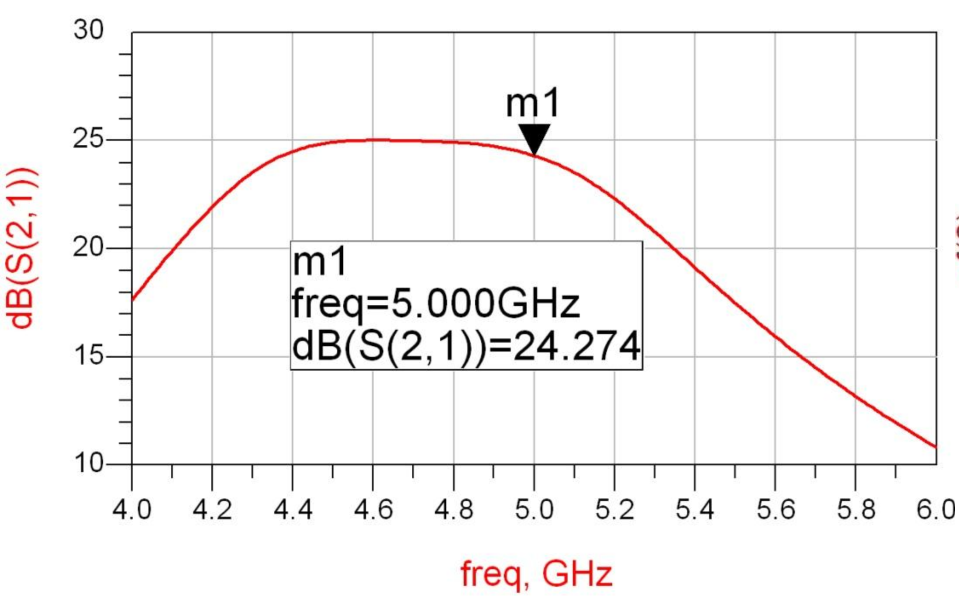
stub “combinat”



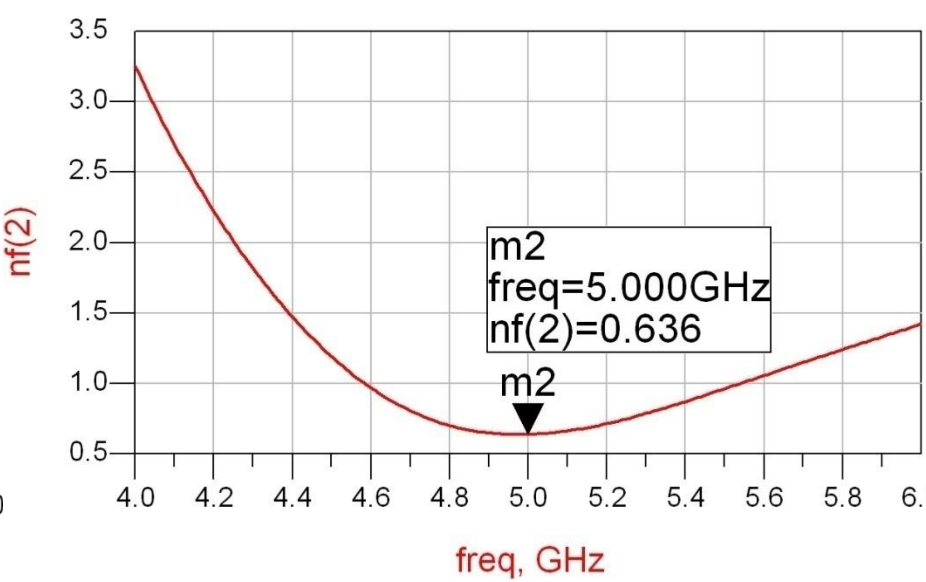
ADS 2



Term
Term1
Num=1
Z=50 Ohm
TLOC
TL3
Z=50.0 Ohm
E=137.9
F=5 GHz
Ref
TLIN
TL1
Z=50.0 Ohm
E=146.2
F=5 GHz
S2P
SNP1
File="C:\Doc\AvagoATF34143\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=55.6
F=5 GHz
TLOC
TL7
Z=50.0 Ohm
E=58.8
F=5 GHz
Ref
TLIN
TL6
Z=50.0 Ohm
E=15
F=5 GHz
S2P
SNP2
File="C:\Doc\AvagoATF34143\f341433a.s2p"
TLIN
TL2
Z=50.0 Ohm
E=116.8
F=5 GHz
TLOC
TL4
Z=50 Ohm
E=159.3
F=5 GHz
Ref
Term
Term2
Num=2
Z=50 Ohm



m1
m1
freq=5.000GHz
dB(S(2,1))=24.274
dB(S(2,1))
freq, GHz
30
25
20
15
10
4.0
4.2
4.4
4.6
4.8
5.0
5.2
5.4
5.6
5.8
6.0



m2
freq=5.000GHz
nf(2)=0.636
m2
nf(2)
freq, GHz
3.5
3.0
2.5
2.0
1.5
1.0
0.5
4.0
4.2
4.4
4.6
4.8
5.0
5.2
5.4
5.6
5.8
6.0

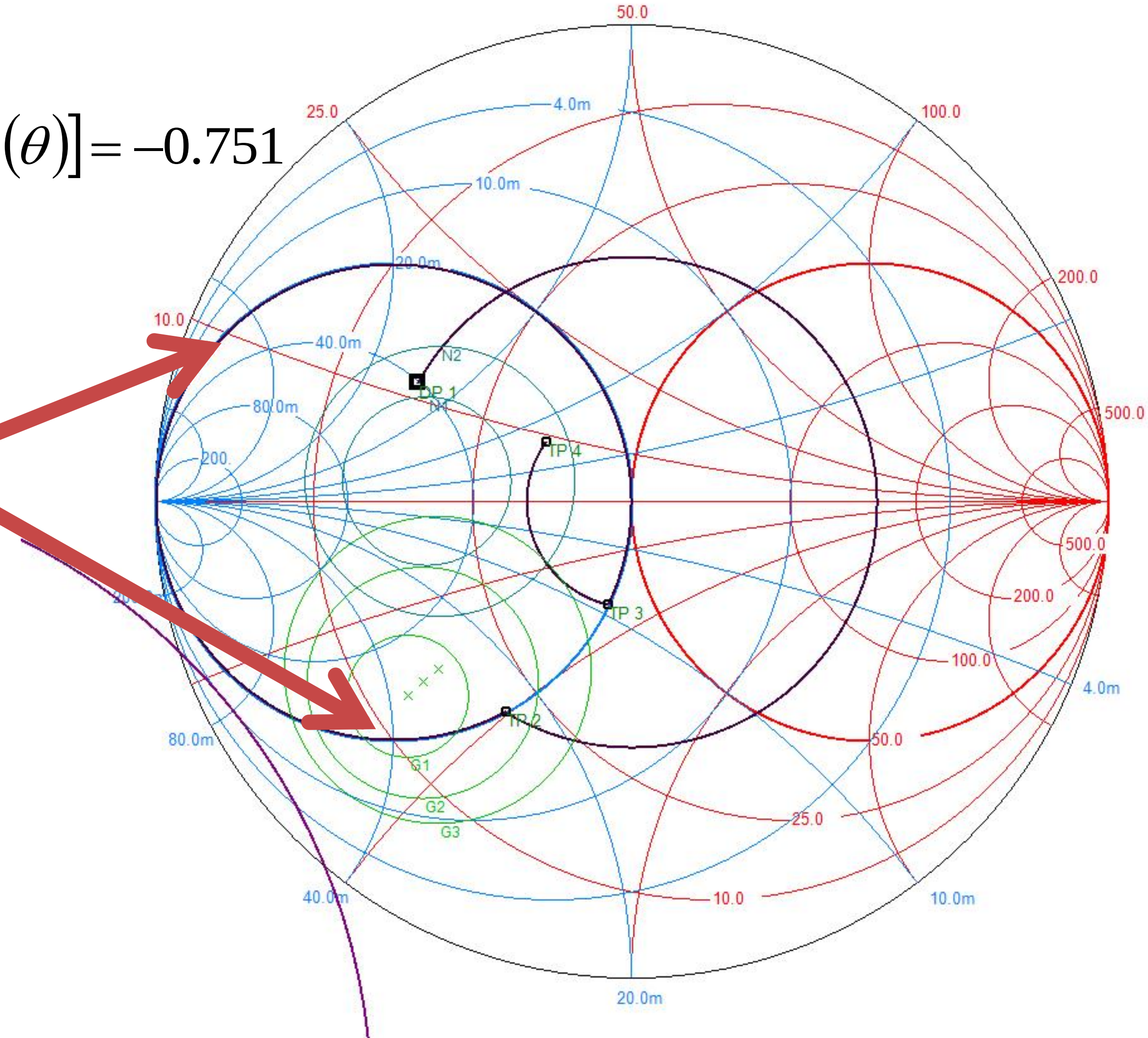
Diagrama Smith 3

$\theta_{L1} = 55.6°$ $\theta_{S2} = 135.9°$

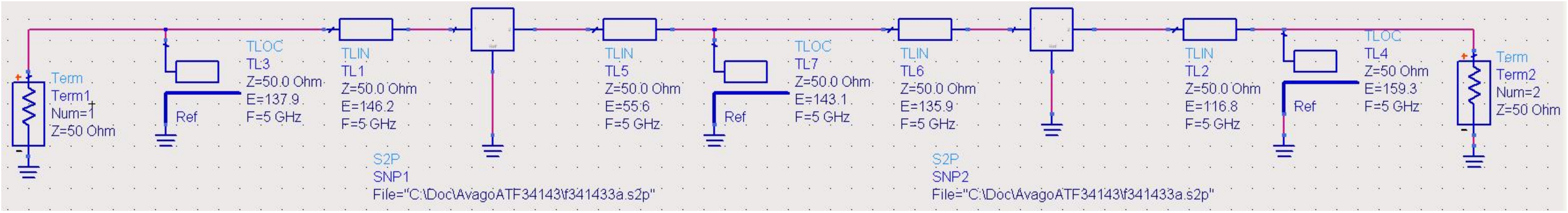
$$\mathrm{Im}[y_{sp}] = \mathrm{Im}[y_{L1}(\theta)] + \mathrm{Im}[y_{S2}(\theta)] = -0.751$$

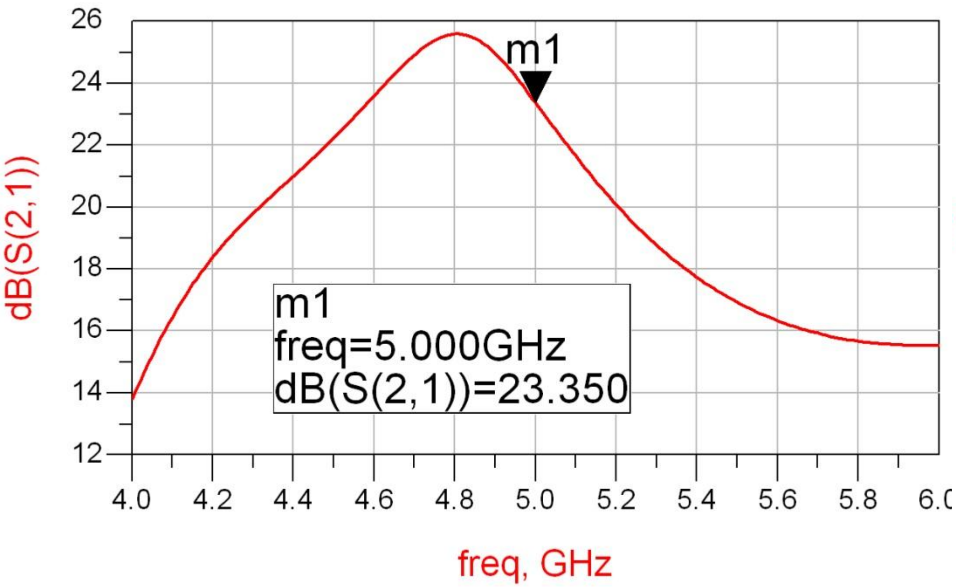
$\theta_{sp} = 143.1°$

stub "combinat"



ADS 3


Term
Term1
Num=1
Z=50 Ohm
Ref
TLOC
TL3
Z=50.0 Ohm
E=137.9
F=5 GHz
TLIN
TL1
Z=50.0 Ohm
E=146.2
F=5 GHz
S2P
SNP1
File="C:\Doc\AvagoATF34143\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=55.6
F=5 GHz
Ref
TLOC
TL7
Z=50.0 Ohm
E=143.1
F=5 GHz
TLIN
TL6
Z=50.0 Ohm
E=135.9
F=5 GHz
S2P
SNP2
File="C:\Doc\AvagoATF34143\f341433a.s2p"
TLIN
TL2
Z=50.0 Ohm
E=116.8
F=5 GHz
Ref
TLOC
TL4
Z=50 Ohm
E=159.3
F=5 GHz
Term
Term2
Num=2
Z=50 Ohm


m1
m1
freq=5.000GHz
dB(S(2,1))=23.350
dB(S(2,1))
freq, GHz

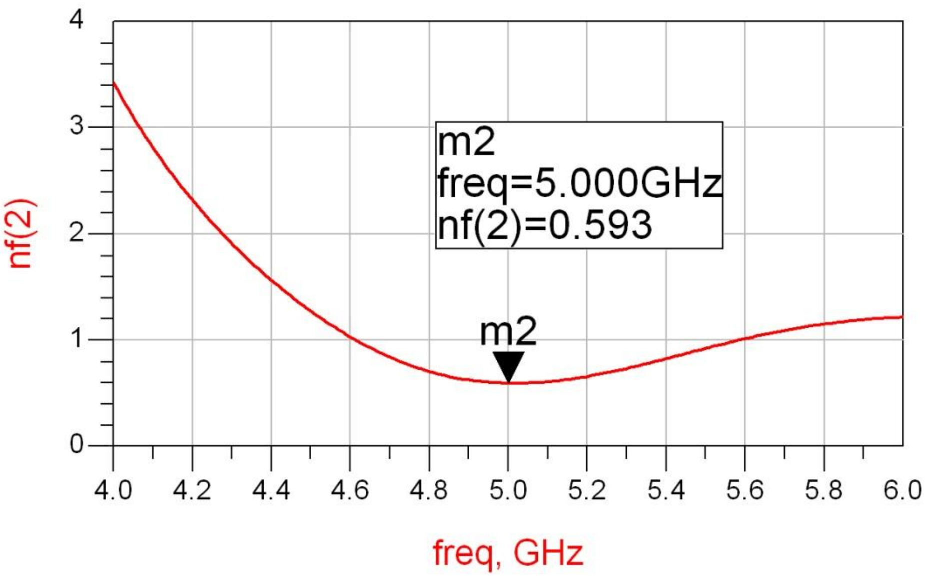

m2
freq=5.000GHz
nf(2)=0.593
m2
nf(2)
freq, GHz

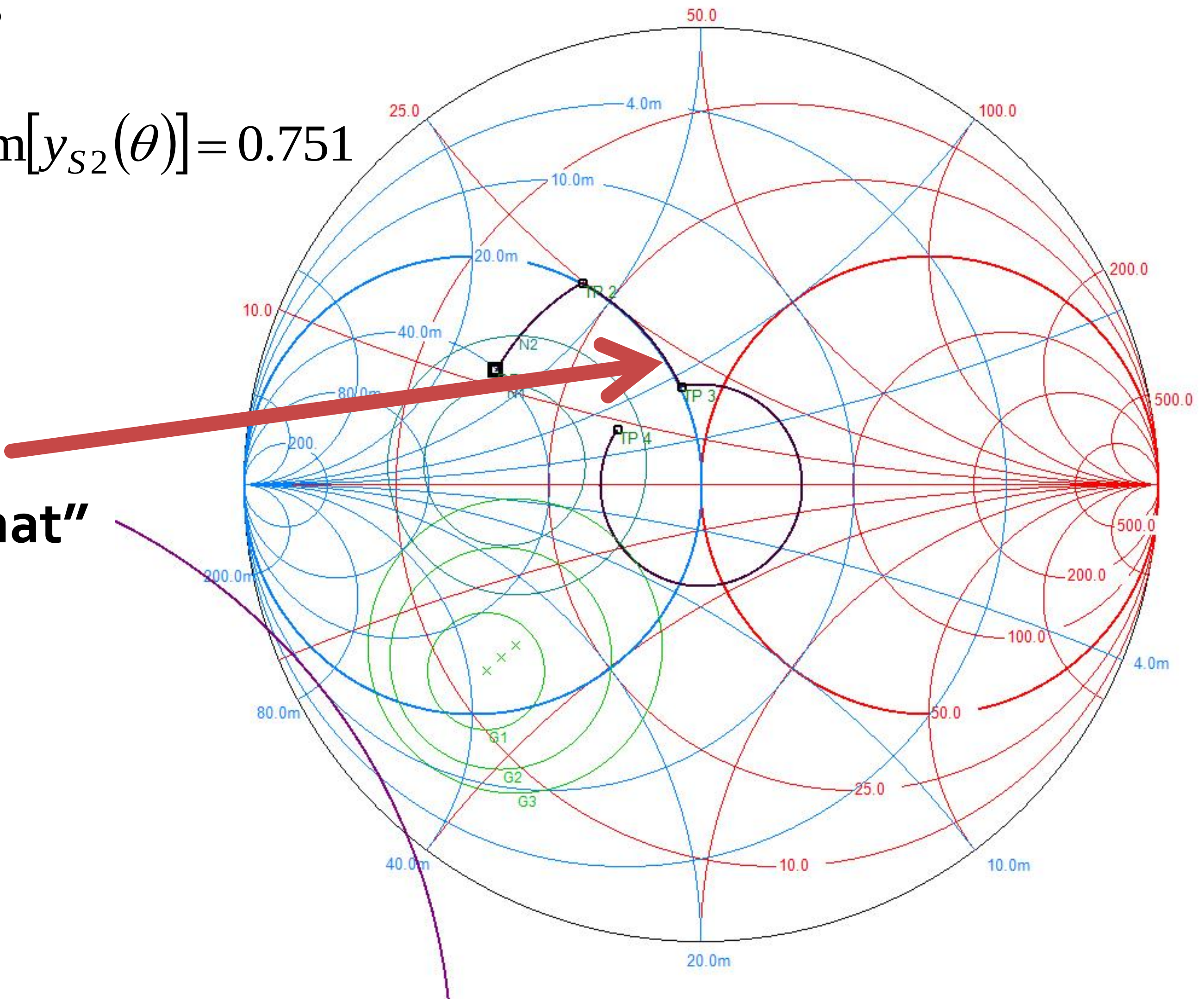
Diagrama Smith 4

$$\theta_{L1} = 158.4° \qquad \theta_{S2} = 15°$$

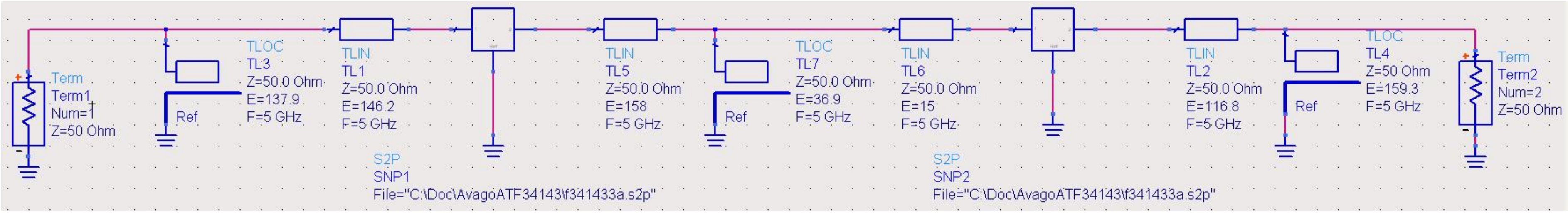
$$\mathrm{Im}[y_{sp}] = \mathrm{Im}[y_{L1}(\theta)] + \mathrm{Im}[y_{S2}(\theta)] = 0.751$$

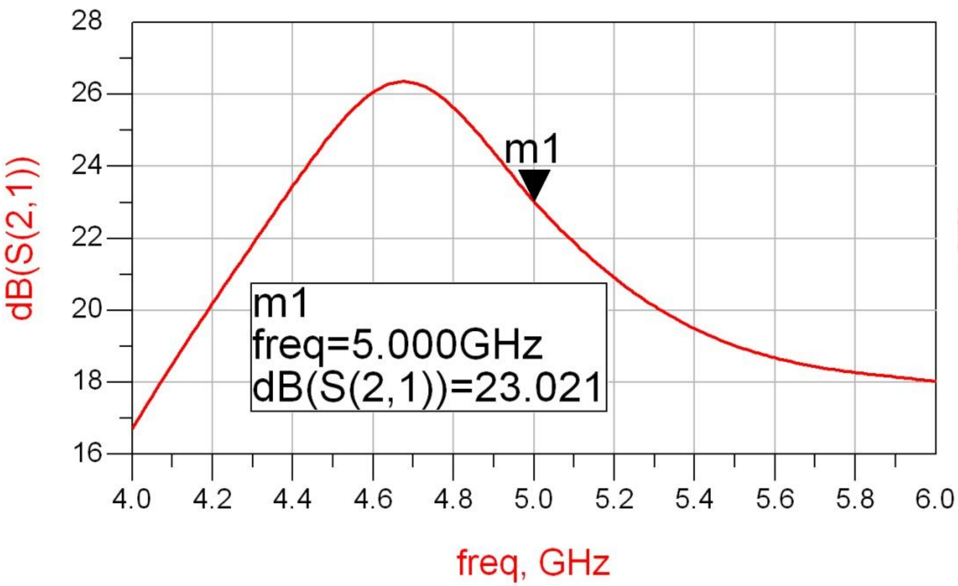
$$\theta_{sp} = 36.9°$$

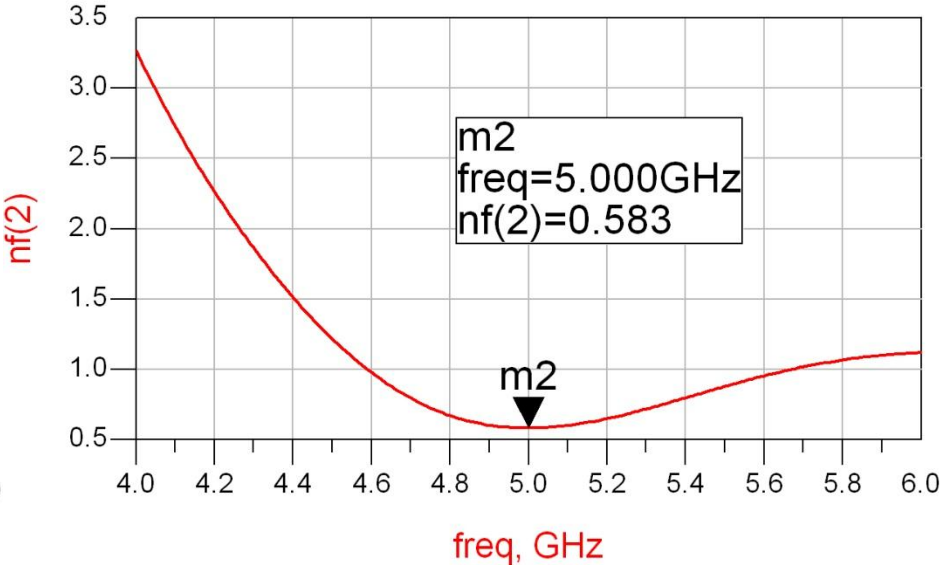
stub "combinat"



ADS 4


Term
Term1
Num=1
Z=50 Ohm
Ref
TLOC
TL3
Z=50.0 Ohm
E=137.9
F=5 GHz
TLIN
TL1
Z=50.0 Ohm
E=146.2
F=5 GHz
S2P
SNP1
File="C:\Doc\AvagoATF34143\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=158
F=5 GHz
Ref
TLOC
TL7
Z=50.0 Ohm
E=36.9
F=5 GHz
TLIN
TL6
Z=50.0 Ohm
E=15
F=5 GHz
S2P
SNP2
File="C:\Doc\AvagoATF34143\f341433a.s2p"
TLIN
TL2
Z=50.0 Ohm
E=116.8
F=5 GHz
Ref
TLOC
TL4
Z=50 Ohm
E=159.3
F=5 GHz
Term
Term2
Num=2
Z=50 Ohm


m1
m1
freq=5.000GHz
dB(S(2,1))=23.021
dB(S(2,1))
freq, GHz
28
26
24
22
20
18
16
4.0
4.2
4.4
4.6
4.8
5.0
5.2
5.4
5.6
5.8
6.0


m2
freq=5.000GHz
nf(2)=0.583
m2
nf(2)
freq, GHz
3.5
3.0
2.5
2.0
1.5
1.0
0.5
4.0
4.2
4.4
4.6
4.8
5.0
5.2
5.4
5.6
5.8
6.0

Adaptare inter-etaje

- Toate variantele obtinute indeplinesc conditiile de castig si zgomot
- Se alege una convenabila in functie de:
  - dimensiunile fizice ale liniilor $l = \frac{\theta}{360^\circ} \cdot \lambda$
  - comportare in frecventa
  - stabilitate
  - performanta (zgomot/castig)
  - reflexie intrare iesire
  - etc.

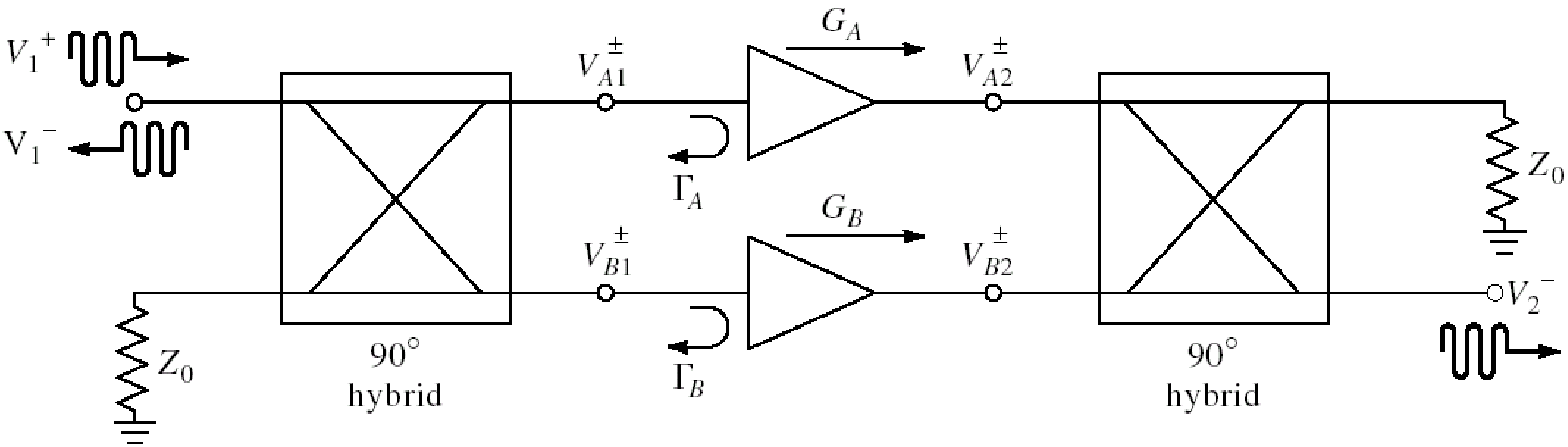
Amplificatoare de banda larga

Amplificatoare de microunde

Amplificatoare de banda larga

- Se pot obtine prin un numar de tehnici de proiectare
  1. Retele de adaptare care sa compenseze scaderea castigului cu frecventa
  2. Retele de adaptare rezistive
  3. Reactie negativa
  4. Amplificatoare echilibrate
  5. Amplificatoare distribuite
  6. Amplificatoare diferentiale

Amplificatoare echilibrate



- 2 Amplificatoare (identice) cu doua cuploare hibride 3 dB / 90° la intrare si iesire

$$S_{21} = \frac{-j}{2} \cdot (G_A + G_B)$$

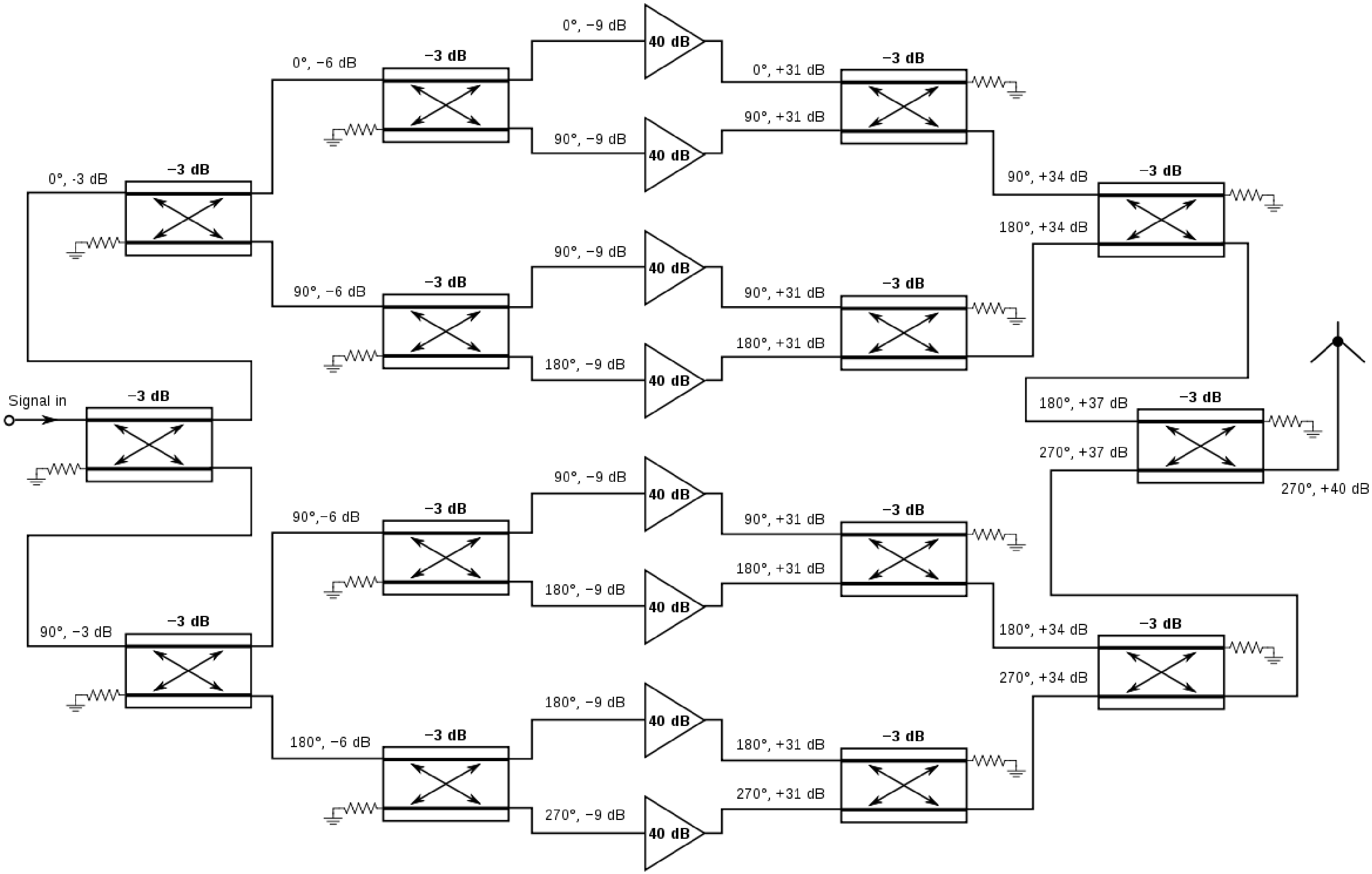
$$S_{11} = \frac{1}{2} \cdot (\Gamma_A - \Gamma_B)$$

$$F = \frac{1}{2} \cdot (F_A + F_B)$$

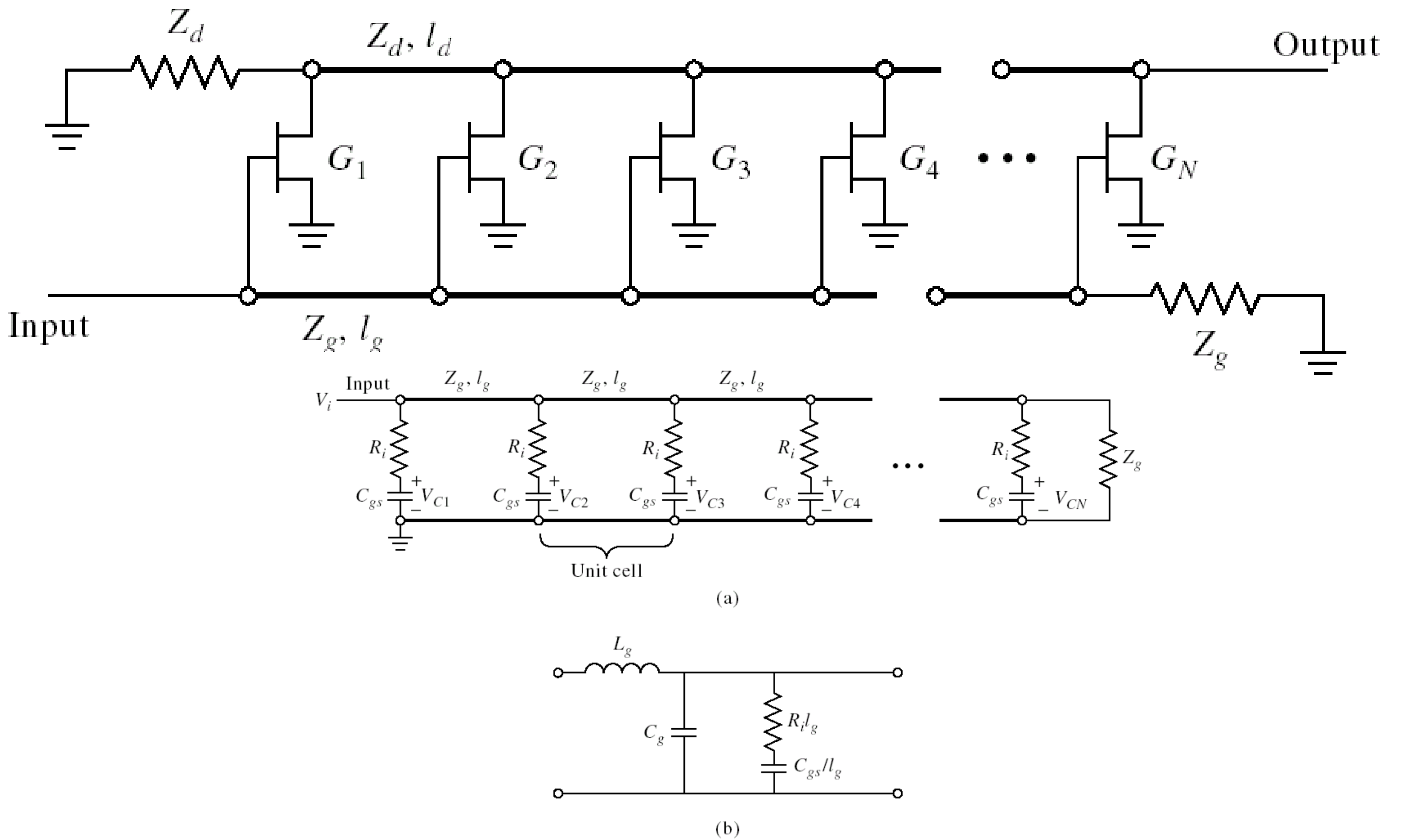
$$S_{21}\big|_{A=B} = -j \cdot G$$

$$S_{11}\big|_{A=B} = 0$$

Amplificatoare echilibrate



Amplificatoare distribuite



Amplificatoare distribuite

- Conditia de sincronizare
  - intarzierea pe liniile de intrare (grila) egala cu cea de pe liniile de iesire (drena)

$$\gamma_g = \alpha_g + j \cdot \beta_g \qquad \gamma_d = \alpha_d + j \cdot \beta_d \qquad \beta_g \cdot l_g = \beta_d \cdot l_d$$

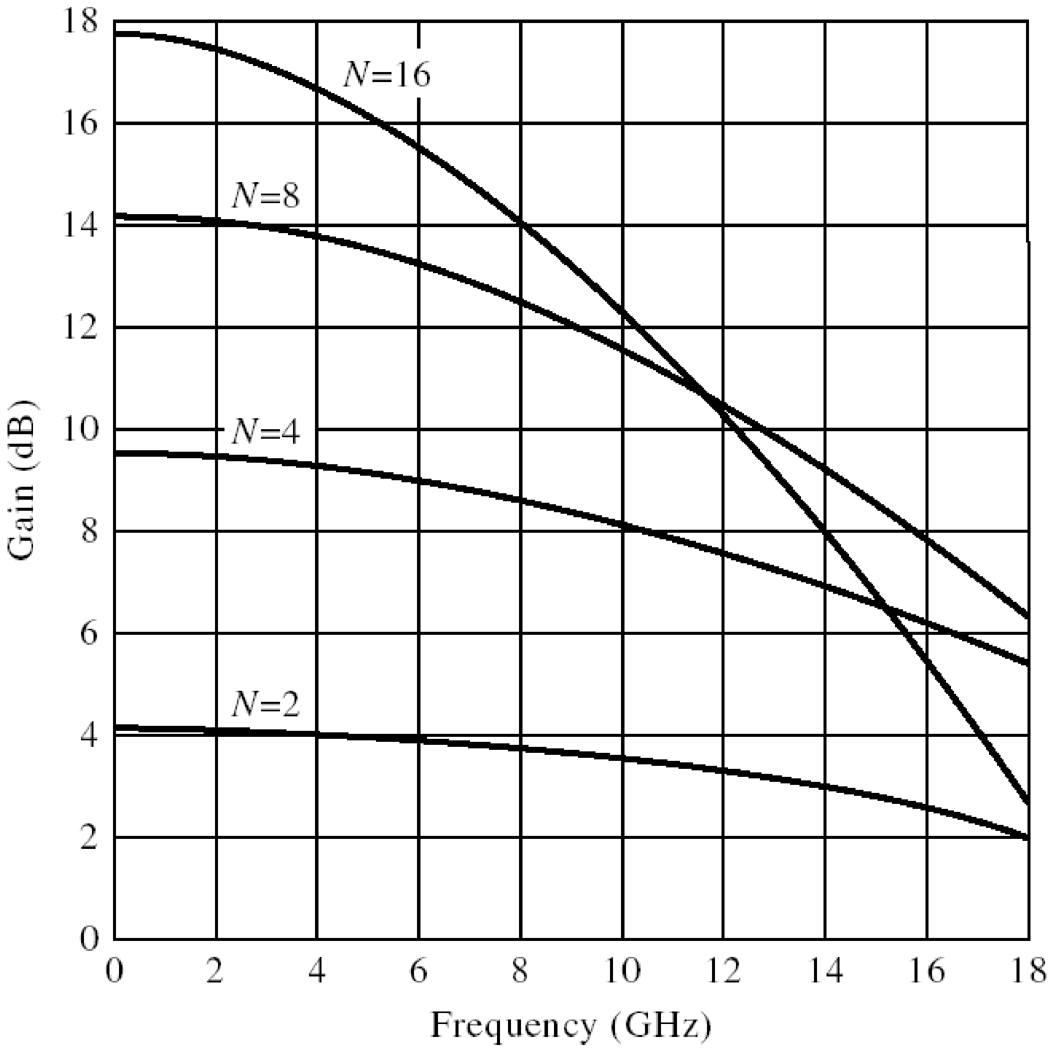
- Castigul de putere

$$G = \frac{g_m^2 \cdot Z_d \cdot Z_g}{4} \cdot \frac{\left(e^{-N \cdot \alpha_g \cdot l_g} - e^{-N \cdot \alpha_d \cdot l_d}\right)^2}{\left(e^{-\alpha_g \cdot l_g} - e^{-\alpha_d \cdot l_d}\right)^2}$$

- Castigul de putere fara pierderi

$$G = \frac{g_m^2 \cdot Z_d \cdot Z_g \cdot N^2}{4}$$

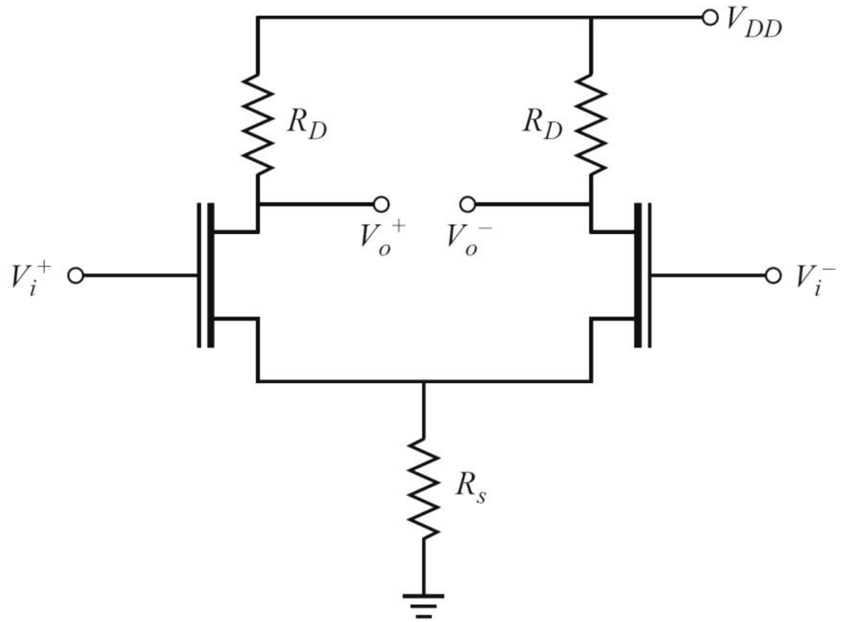
Amplificatoare distribuite



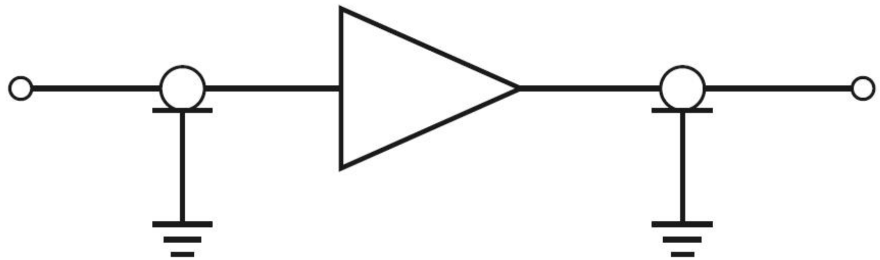
$$N_{opt} = \frac{\ln(\alpha_g \cdot l_g) - \ln(\alpha_d \cdot l_d)}{\alpha_g \cdot l_g - \alpha_d \cdot l_d}$$

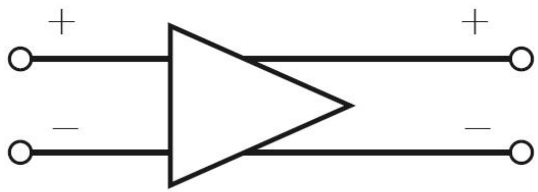
Amplificatoare diferentiale

- Capacitatile de intrare in cele doua tranzistoare in conexiune diferentiala apar conectate in serie
- Se dubleaza astfel frecventa unitara

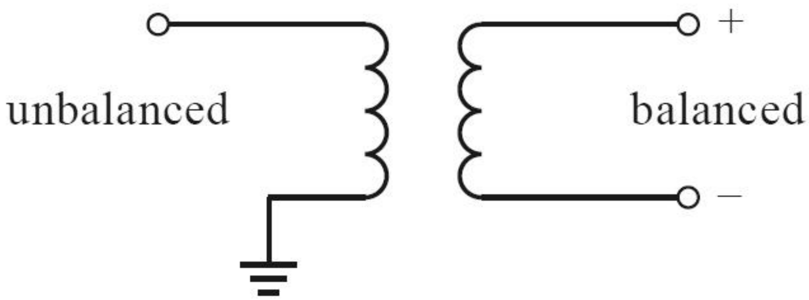


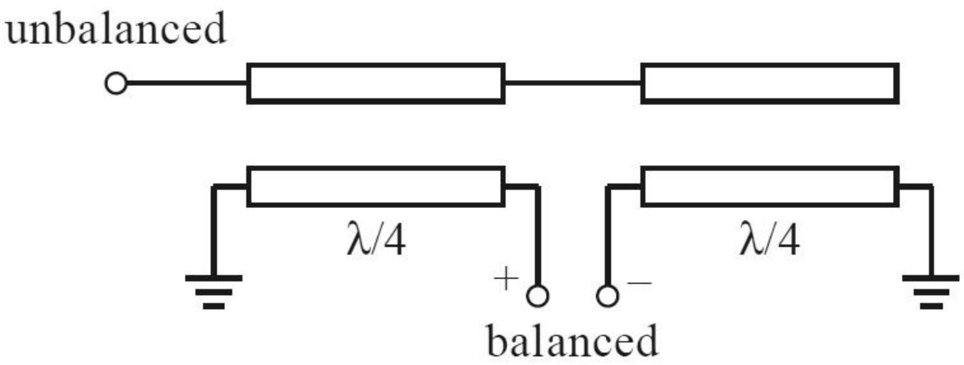
Amplificatoare diferentiale





- Se utilizeaza structuri de circuit care sa faca conversia de la dispozitivele unipolare la cele diferentiale
  - cuploare hibride 3dB / 180°
  - "balun" (balanced - unbalanced)





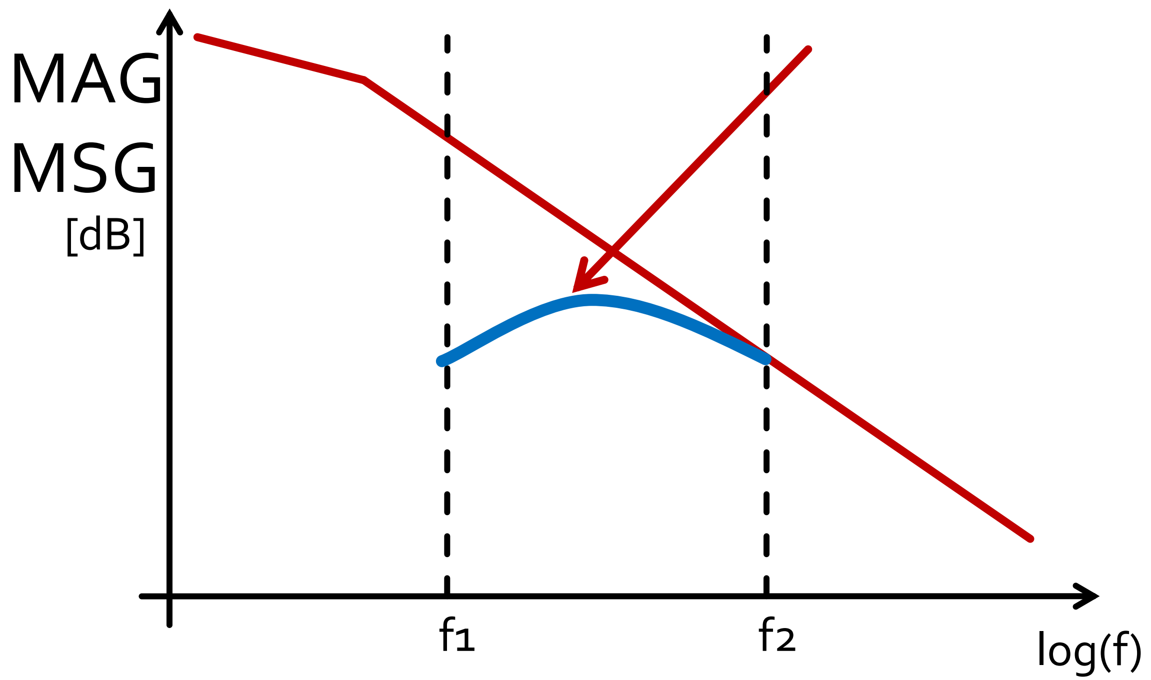
Amplificatoare de banda larga

- Se pot obtine prin un numar de tehnici de proiectare
  1. **Retele de adaptare care sa compenseze scaderea castigului cu frecventa**
  2. Retele de adaptare rezistive
  3. Reactie negativa
  4. Amplificatoare echilibrate
  5. Amplificatoare distribuite
  6. Amplificatoare diferentiale

Amplificatoare de banda larga

1. **Retele de adaptare care sa compenseze scaderea castigului cu frecventa**

- Metoda utilizata este de a repeta proiectarea la mai multe (macar 2) frecvente si impunerea unui castig egal la acestea



Filtre pentru microunde

Filtre pentru microunde

- In domeniul microundelor se utilizeaza doua strategii de implementare a filtrelor
  - structuri specifice microundelor (linii cuplate, rezonatori dielectrici, structuri periodice)
  - sinteza de filtre cu elemente concentrate urmate de implementare cu linii de transmisie
- prima strategie duce la obtinerea unor filtre mai eficiente dar e caracterizata de
  - generalitate mai mica
  - proiectare deseori dificila (lipsa relatiilor analitice)

Sinteza filtrelor

- Sinteza filtrelor cu elemente concentrate, urmata de implementarea acestora cu elemente distribuite (linii)
  - generala
  - relatii analitice usor de implementat pe calculator
  - eficienta
- Metoda preferata este metoda pierderilor de insertie

Metoda pierderilor de insertie

$$P_{LR} = \frac{P_S}{P_L} = \frac{1}{1-\left|\Gamma(\omega)\right|^2}$$

- $|\Gamma(\omega)|^2$ este o functie para de ω

$$\left|\Gamma(\omega)\right|^2 = \frac{M\left(\omega^2\right)}{M\left(\omega^2\right)+N\left(\omega^2\right)}$$

$$P_{LR} = 1+\frac{M\left(\omega^2\right)}{N\left(\omega^2\right)}$$

- Alegerea corespunzatoare a polinoamelor M si N determina comportarea filtrului

Metoda pierderilor de insertie

- Se aleg polinoamele pentru implementarea unui FTJ (prototip)
- Acest filtru poate fi convertit la alte functii, scalat in frecventa pentru a obtine alte tipuri de functii

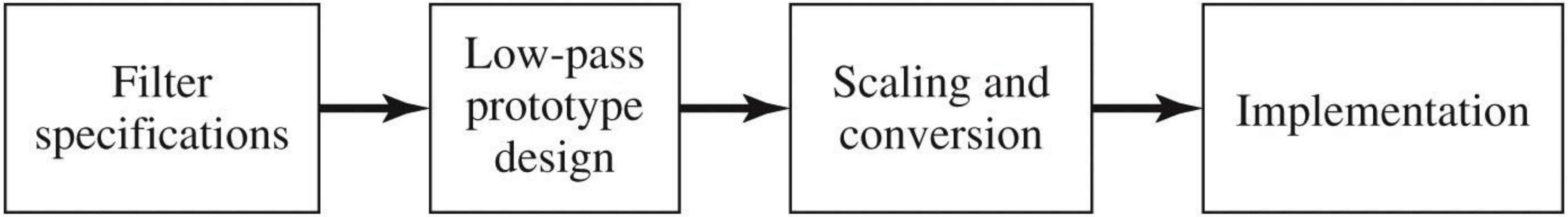


Figure 8.23

Caracteristici de filtru trece jos prototip

- **Maxim plat** (Butterworth, binomial) ofera cea mai plata comportare in banda de trecere
- **Echiriplu** (Cebasev) ofera atenuare mai mare in banda de taiere cu dezavantajul existentei unor variatii (riplu) in banda de trecere
- **Filtre eliptice**, caracterizate de variatii (riplu) si in banda de taiere si in banda de trecere
- **Filtru cu raspuns liniar in faza**, ofera intarziere de grup de maxim plat, cu dezavantajul unei atenuari in putere mai mica, necesar in anumite aplicatii

FTJ prototip Maxim plat/Echiriplu

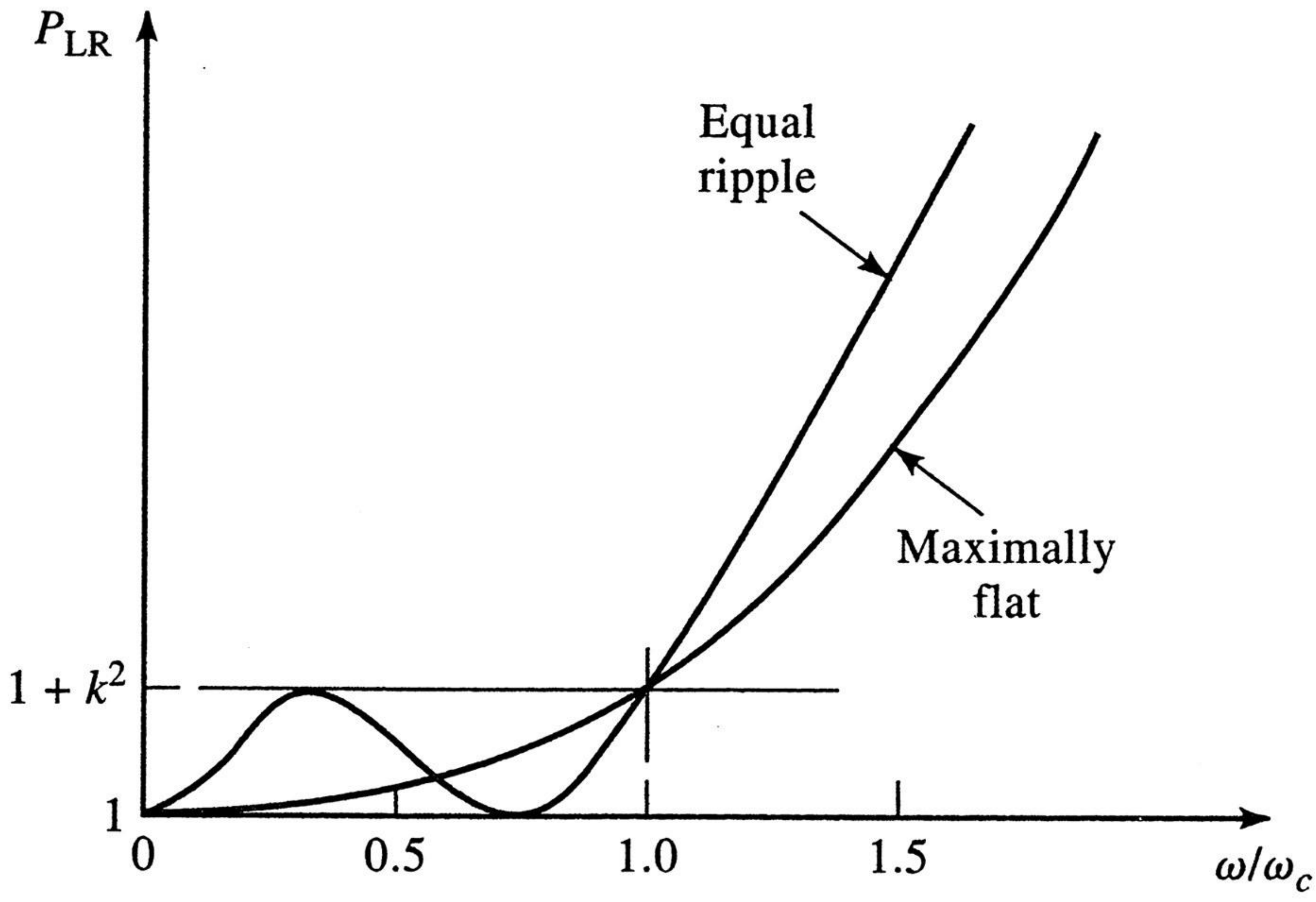


Figure 8.21

FTJ eliptic prototip

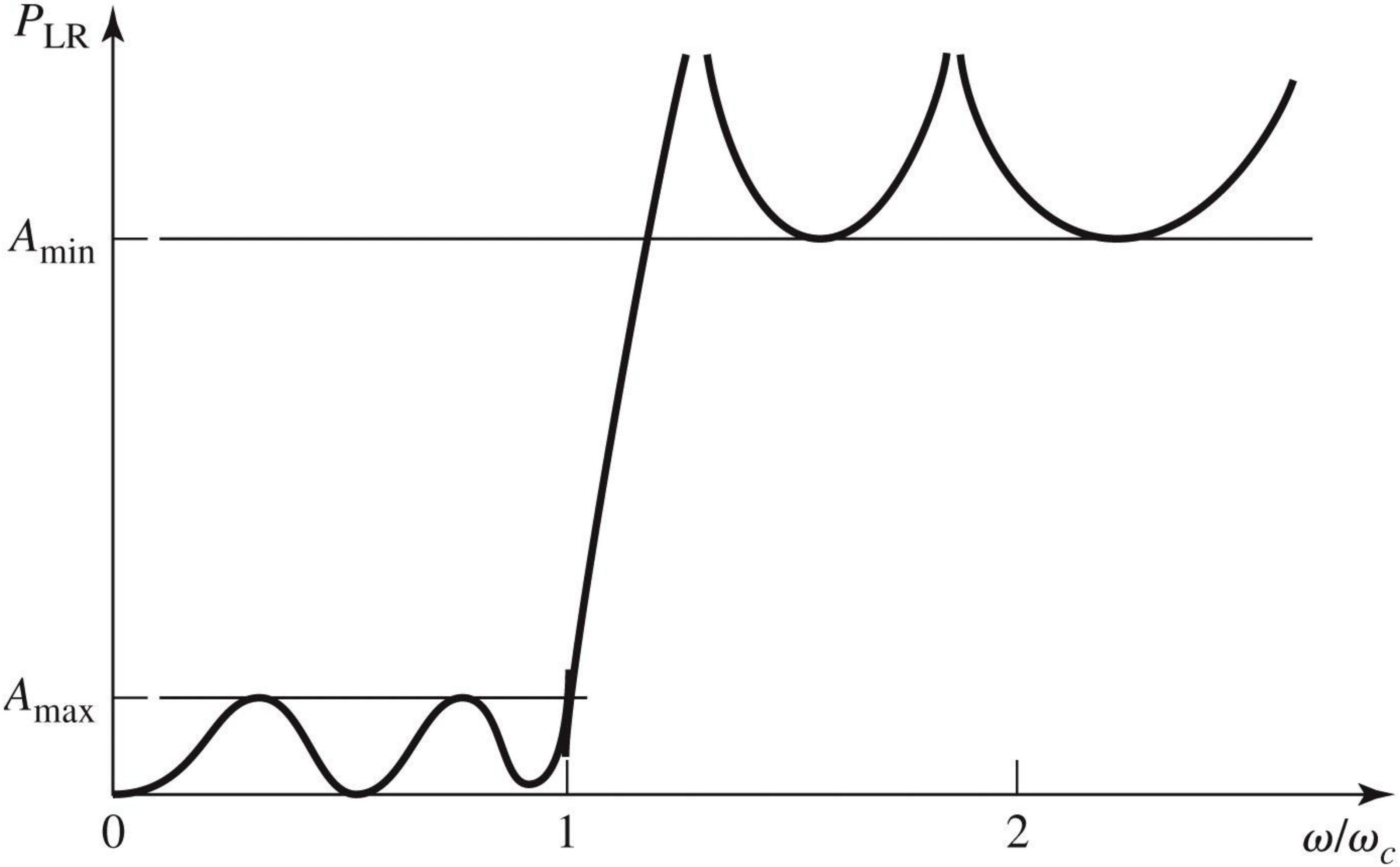


Figure 8.22

FTJ de tip maxim plat

- Polinomul

$$P_{LR} = 1 + k^2 \cdot \left(\frac{\omega}{\omega_c}\right)^{2N}$$

- pentru $\omega >> \omega_c$

$$P_{LR} \approx k^2 \cdot (\omega/\omega_c)^{2N}$$

- atenuarea creste cu 20N dB/decada

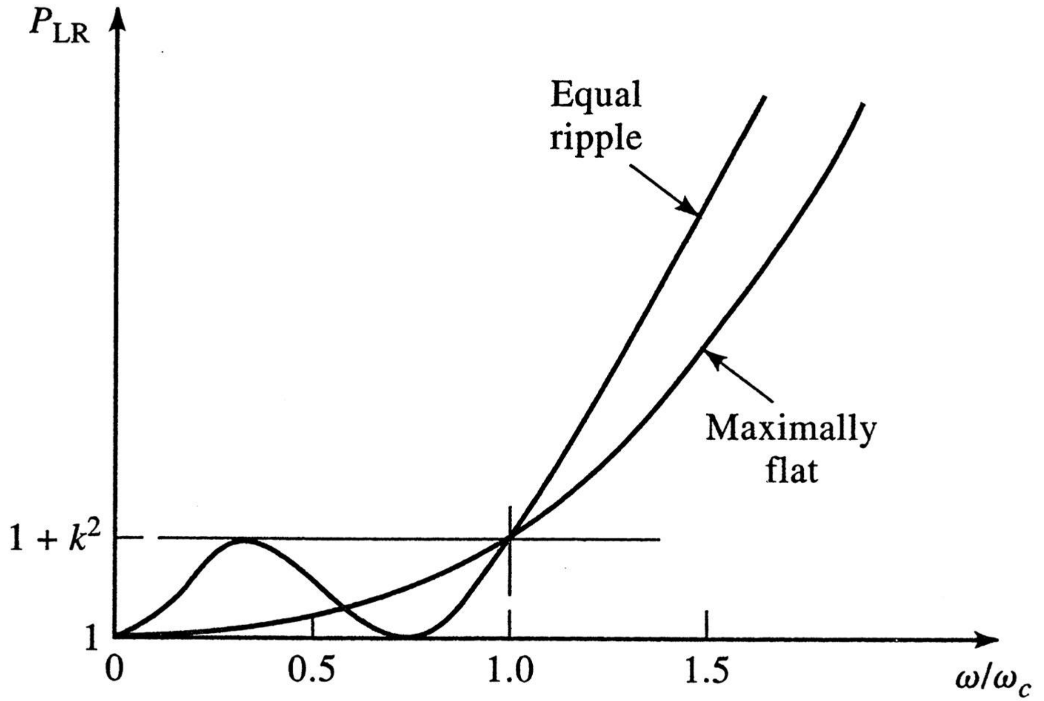


Figure 8.21

- k ofera atenuarea la limita benzii de trecere (3dB implica k = 1)

FTJ de tip echiriplu

- Polinomul

$$P_{LR} = 1 + k^2 \cdot T_N^2\left(\frac{\omega}{\omega_c}\right)$$

- pentru $\omega >> \omega_c$

$$P_{LR} \approx \frac{k^2}{4} \cdot \left(\frac{2 \cdot \omega}{\omega_c}\right)^{2N}$$

- atenuarea creste cu 20N dB/decada

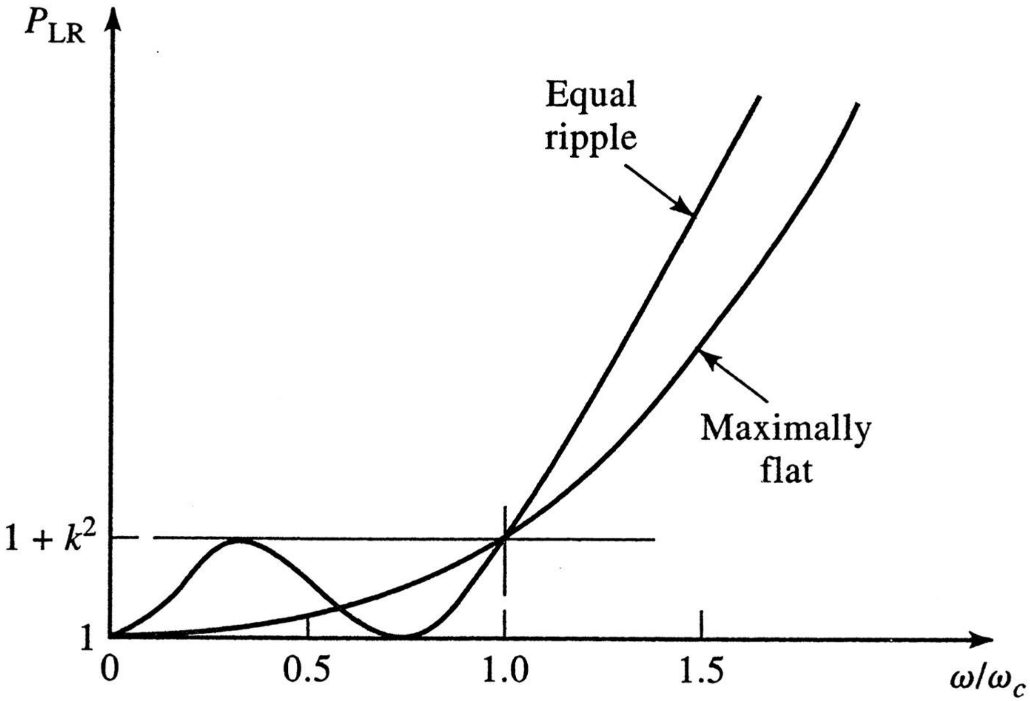


Figure 8.21

- atenuarea este mai mare de $(2^{2N})/4$ decat cea a filtrului binomial la frecventele $\omega >> \omega_c$

Contact

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