Lecture 9
2018/2019

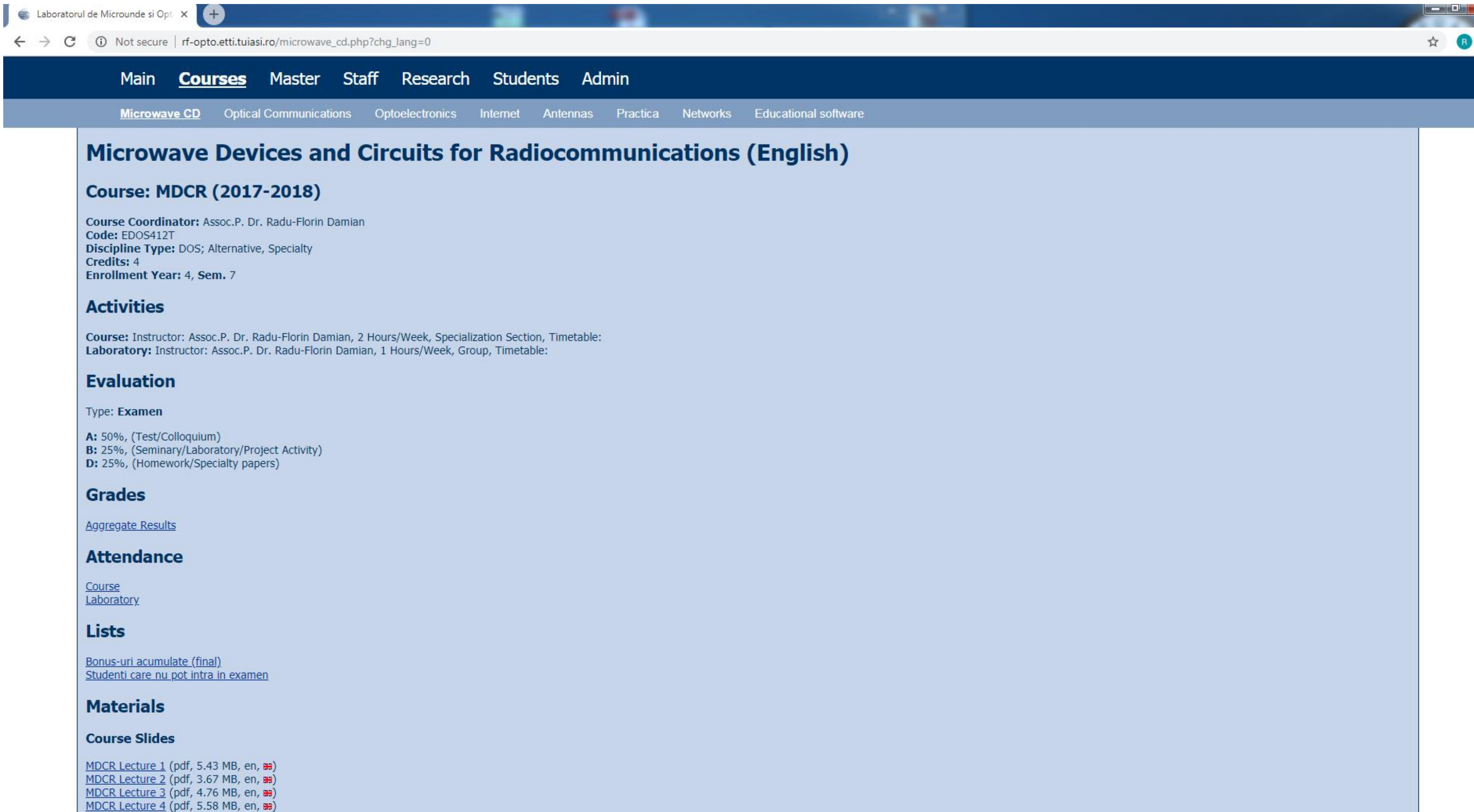
Microwave Devices and Circuits for Radiocommunications

2018/2019

- 2C/1L, **MDCR**
- **Attendance at minimum 7 sessions (course + laboratory)**
- Lectures- **associate professor Radu Damian**
  - Friday 09-11, II.13
  - E – 50% final grade
  - **problems** + (2p atten. lect.) + (**3 tests**) + (bonus activity)
    - 3p=+0.5p
  - **all materials/equipments authorized**
- Laboratory – **associate professor Radu Damian**
  - Wednesday 12-14, II.12 odd weeks
  - L – 25% final grade
  - P – 25% final grade

Materials

- http://rf-opto.etti.tuiasi.ro



Laboratorul de Microunde si Opt

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Main **Courses** Master Staff Research Students Admin

Microwave CD Optical Communications Optoelectronics Internet Antennas Practica Networks Educational software

Microwave Devices and Circuits for Radiocommunications (English)

Course: MDCR (2017-2018)

Course Coordinator: Assoc.P. Dr. Radu-Florin Damian
Code: EDOS412T
Discipline Type: DOS; Alternative, Specialty
Credits: 4
Enrollment Year: 4, **Sem.** 7

Activities

Course: Instructor: Assoc.P. Dr. Radu-Florin Damian, 2 Hours/Week, Specialization Section, Timetable:
Laboratory: Instructor: Assoc.P. Dr. Radu-Florin Damian, 1 Hours/Week, Group, Timetable:

Evaluation

Type: **Examen**

A: 50%, (Test/Colloquium)
B: 25%, (Seminary/Laboratory/Project Activity)
D: 25%, (Homework/Specialty papers)

Grades

Aggregate Results

Attendance

Course
Laboratory

Lists

Bonus-uri acumulate (final)
Studenti care nu pot intra in examen

Materials

Course Slides

MDCR Lecture 1 (pdf, 5.43 MB, en,)
MDCR Lecture 2 (pdf, 3.67 MB, en,)
MDCR Lecture 3 (pdf, 4.76 MB, en,)
MDCR Lecture 4 (pdf, 5.58 MB, en,)

Examen: Logarithmic scales

dB	$= 10 \cdot \log_{10}(P_2 / P_1)$
0 dB	= 1
+ 0.1 dB	= 1.023 (+2.3%)
+ 3 dB	= 2
+ 5 dB	= 3
+ 10 dB	= 10
-3 dB	= 0.5
-10 dB	= 0.1
-20 dB	= 0.01
-30 dB	= 0.001

dBm	$= 10 \cdot \log_{10}(P / 1\ mW)$
0 dBm	= 1 mW
3 dBm	= 2 mW
5 dBm	= 3 mW
10 dBm	= 10 mW
20 dBm	= 100 mW
-3 dBm	= 0.5 mW
-10 dBm	= 100 μW
-30 dBm	= 1 μW
-60 dBm	= 1 nW

[dBm] + [dB] = [dBm]

[dBm/Hz] + [dB] = [dBm/Hz]

[x] + [dB] = [x]

Examen

- Complex numbers arithmetic**!!!!**
- $z = a + j \cdot b$; $\mathbf{j^2 = -1}$

Impedance Matching

Matching , from the point of view of power transmission

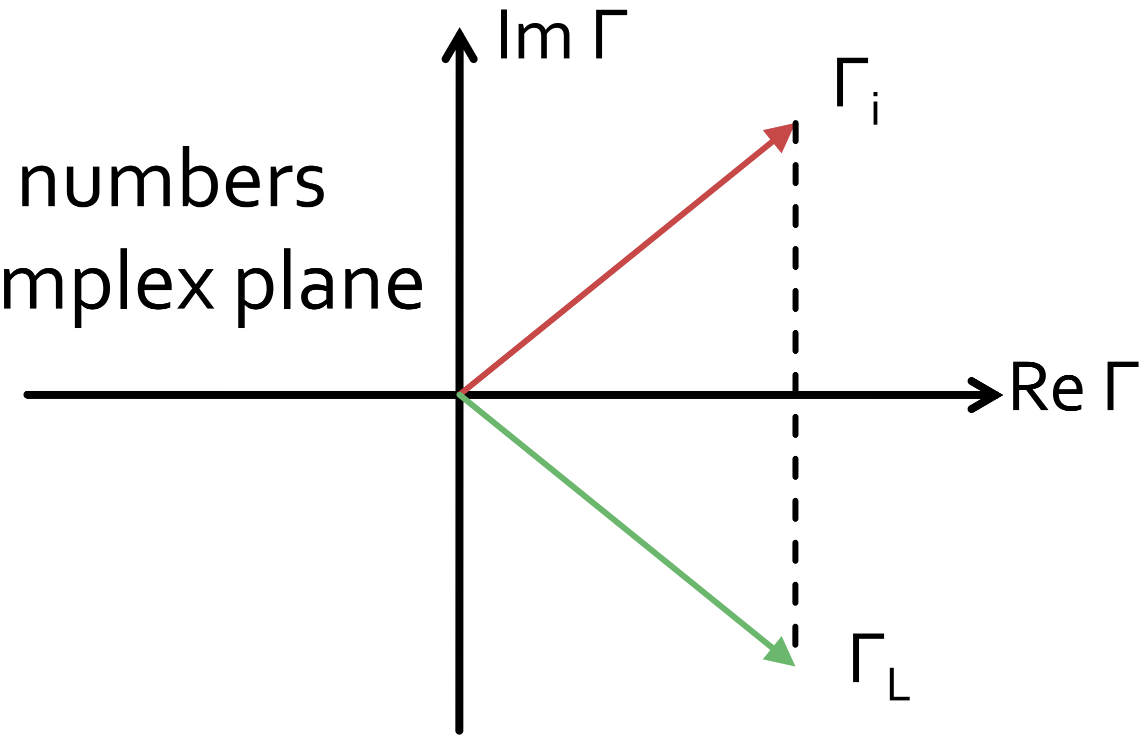
$$Z_L = Z_i^*$$

If we choose a real Zo

$$\Gamma = \frac{Z - Z_0}{Z + Z_0}$$

$$\Gamma_L = \Gamma_i^*$$

- complex numbers
- in the complex plane

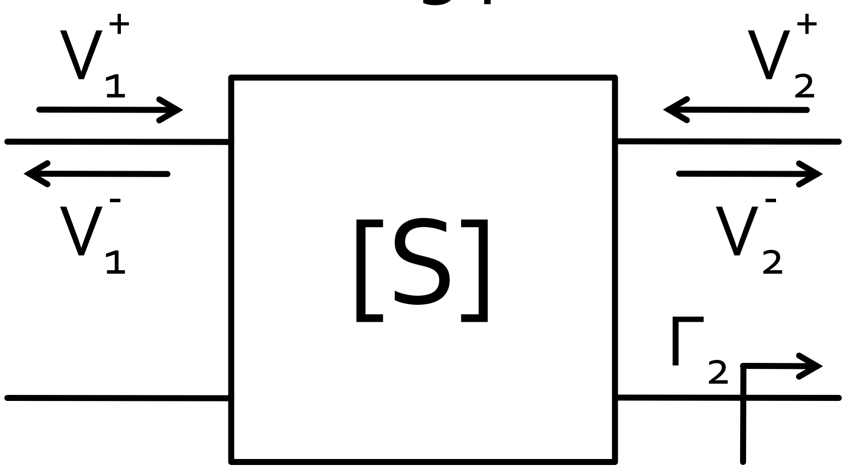


Lecture 3-4

Microwave Network Analysis

Scattering matrix – S

- Scattering parameters



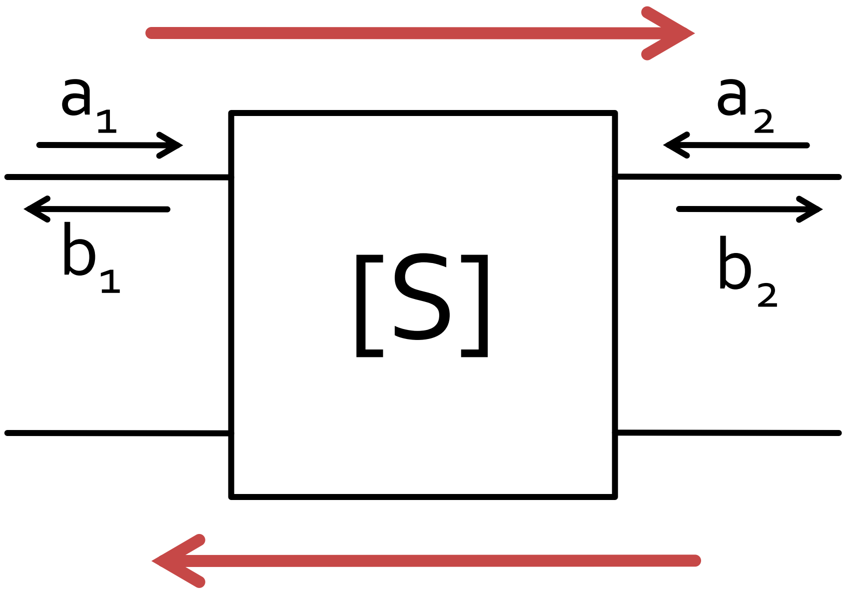
$$\begin{bmatrix} V_1^- \\ V_2^- \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \cdot \begin{bmatrix} V_1^+ \\ V_2^+ \end{bmatrix}$$

$$S_{11} = \left. \frac{V_1^-}{V_1^+} \right|_{V_2^+ = 0} \qquad S_{21} = \left. \frac{V_2^-}{V_1^+} \right|_{V_2^+ = 0}$$

- $V_2^+ = 0$ meaning: port 2 is terminated in matched load to avoid reflections towards the port

$$\Gamma_2 = 0 \rightarrow V_2^+ = 0$$

Scattering matrix – S



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

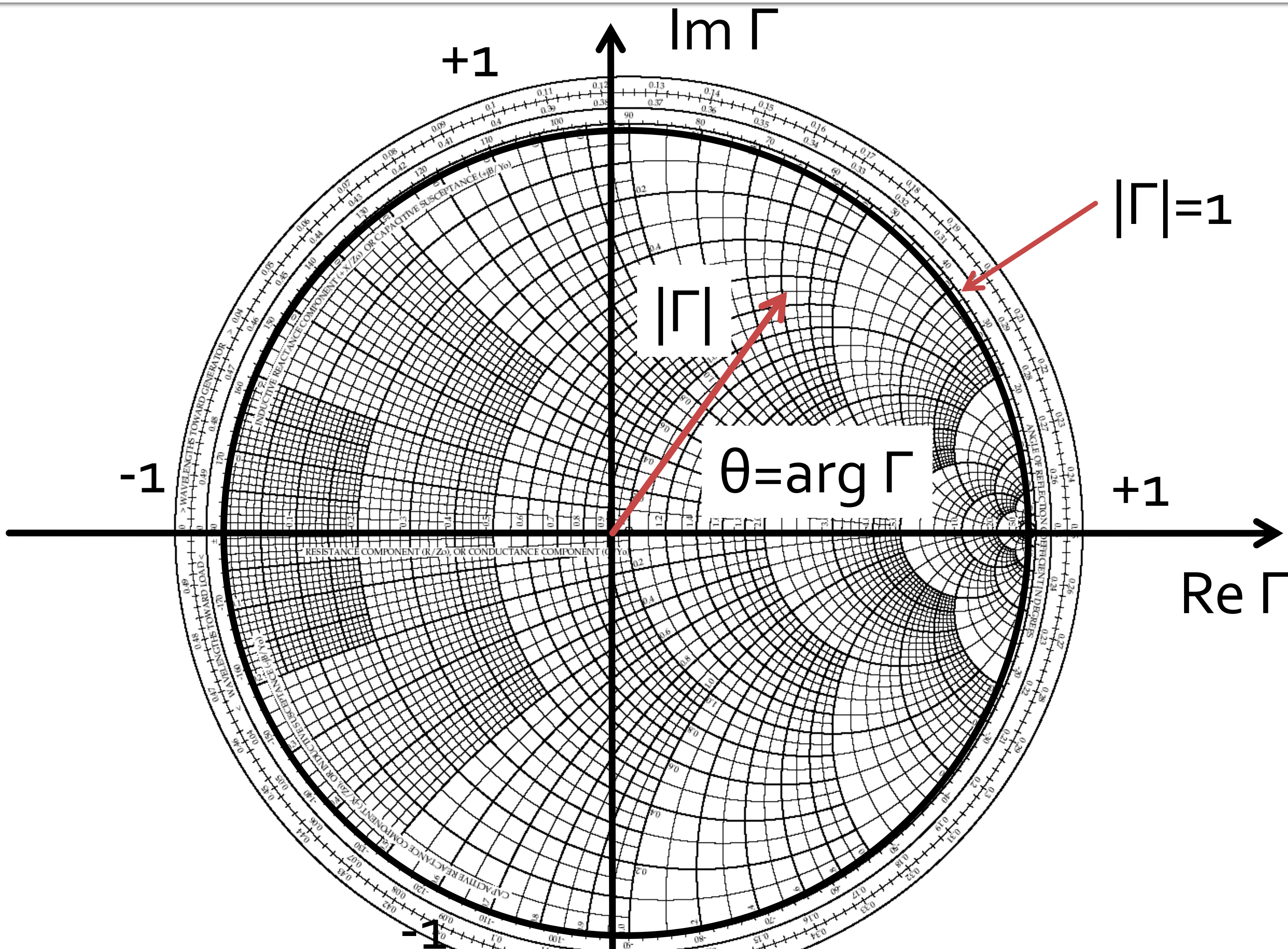
$$|S_{21}|^2 = \frac{Power\ in\ Z_0\ load}{Power\ \ from\ Z_0\ source}$$

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

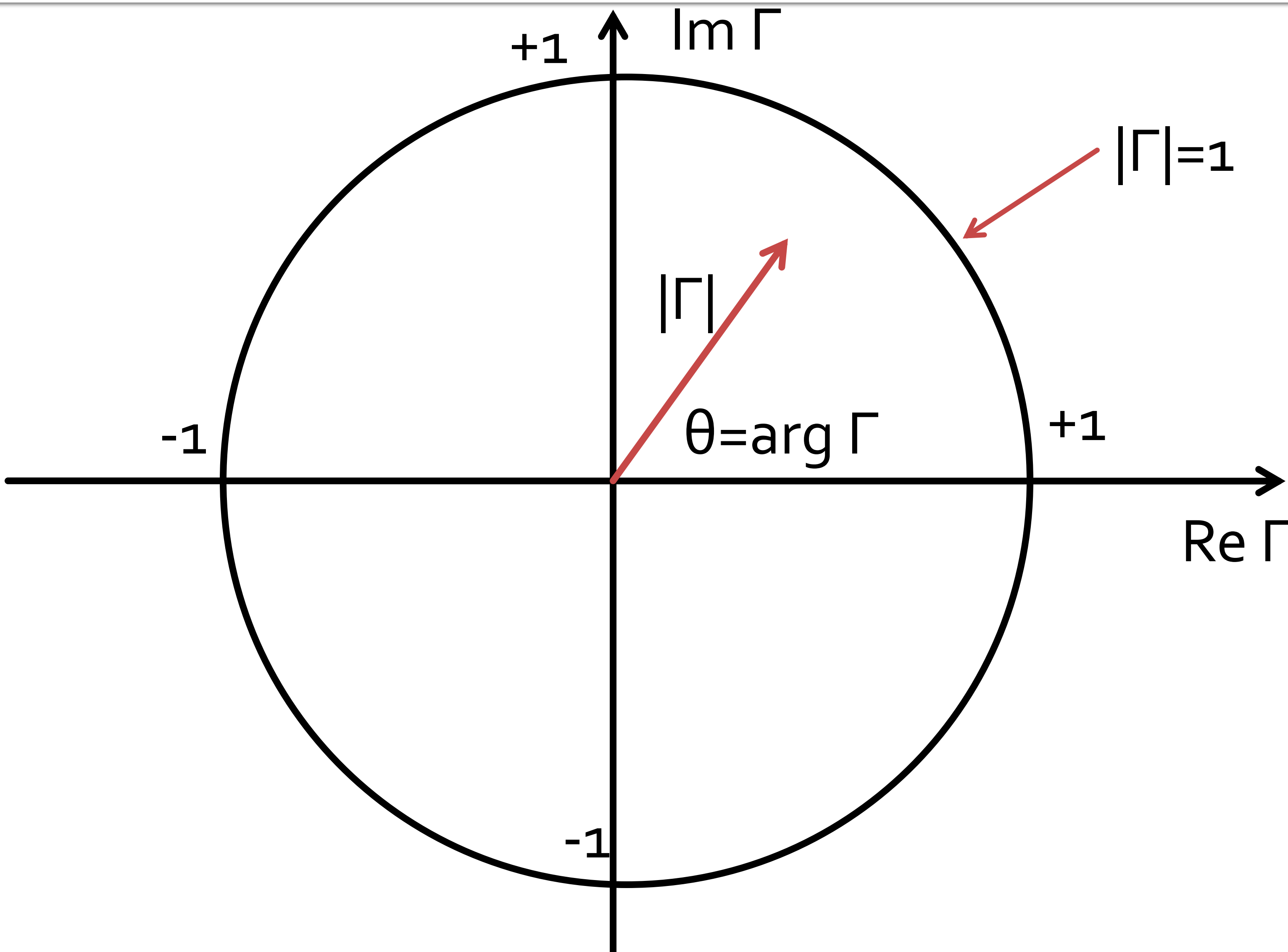
Impedance Matching

The Smith Chart

The Smith Chart



The Smith Chart

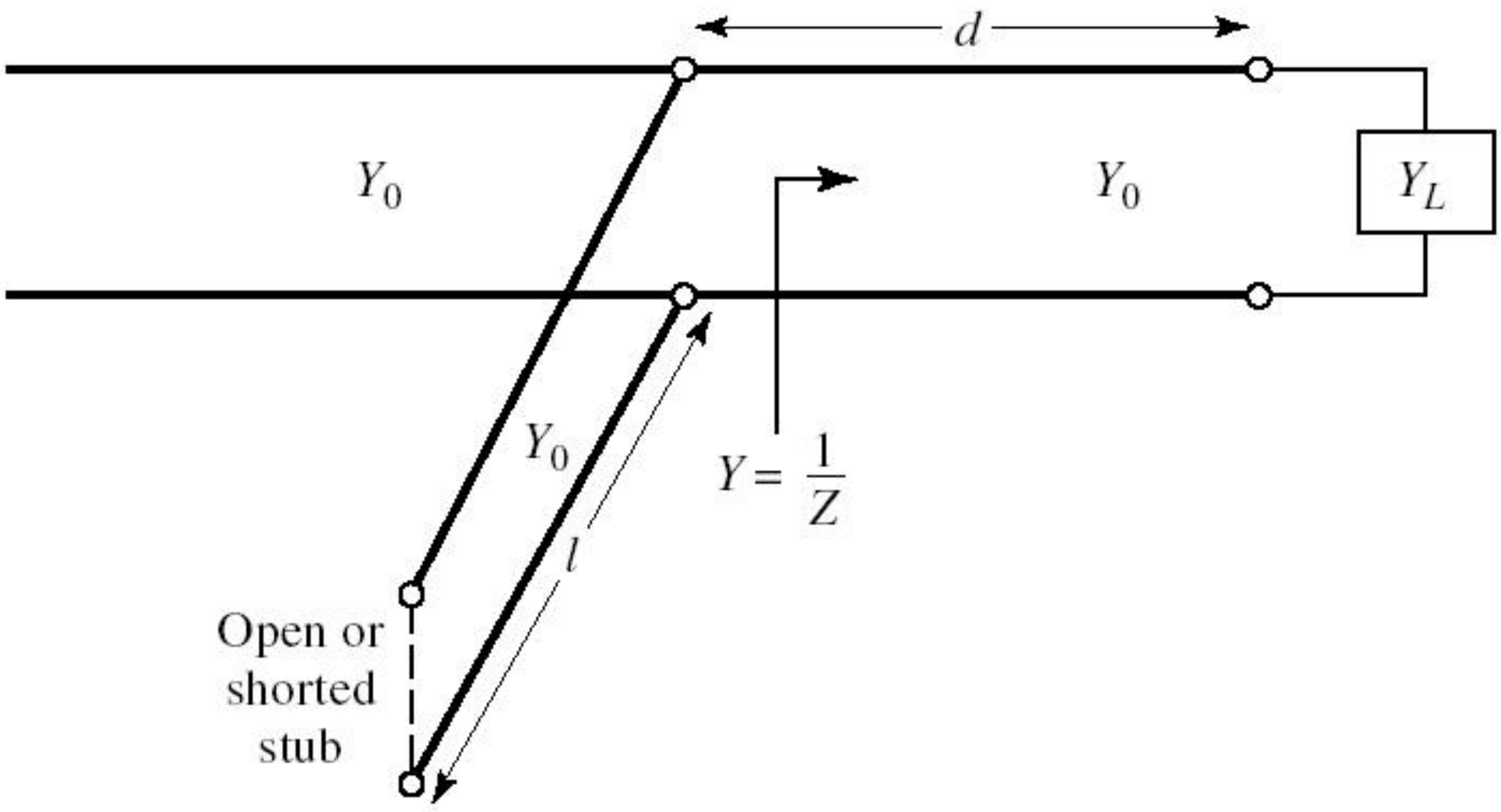


Impedance Matching with Stubs

Impedance Matching

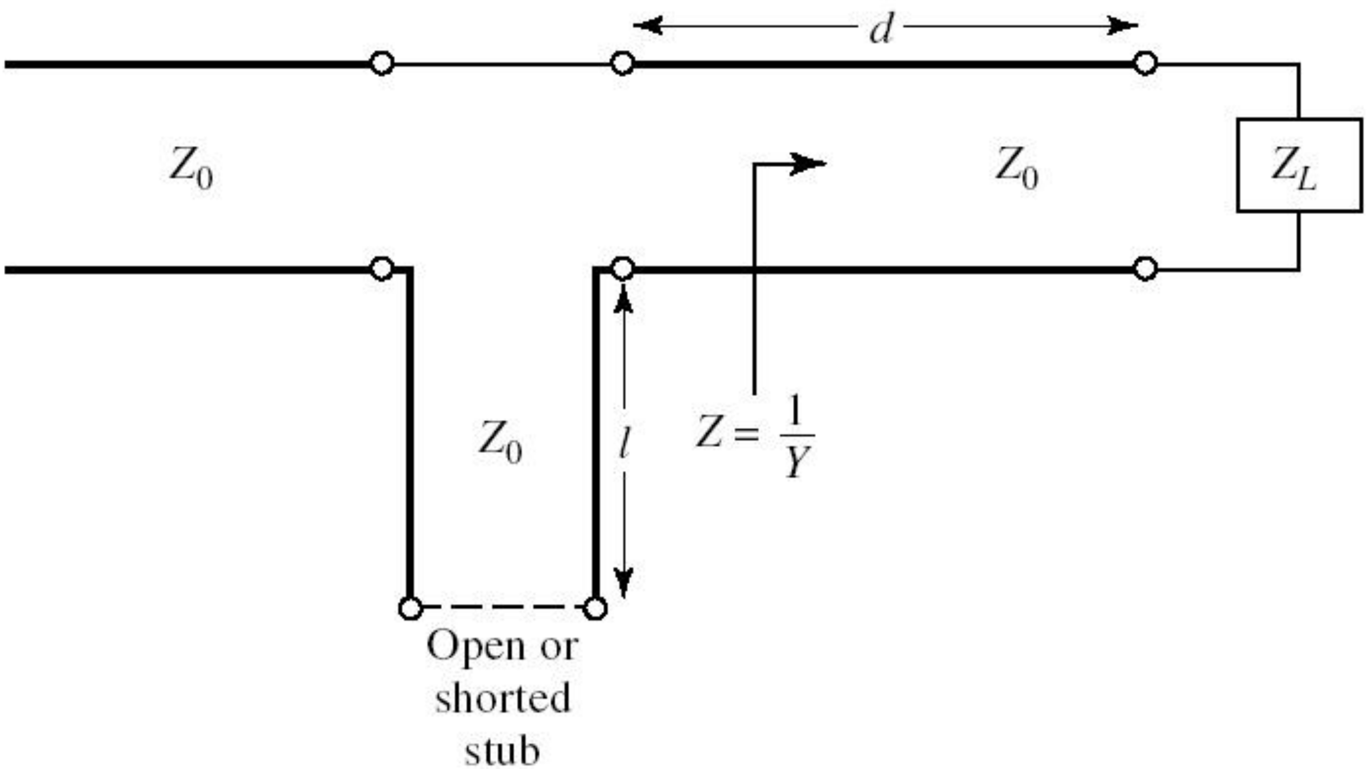
Single stub tuning

- Shunt Stub

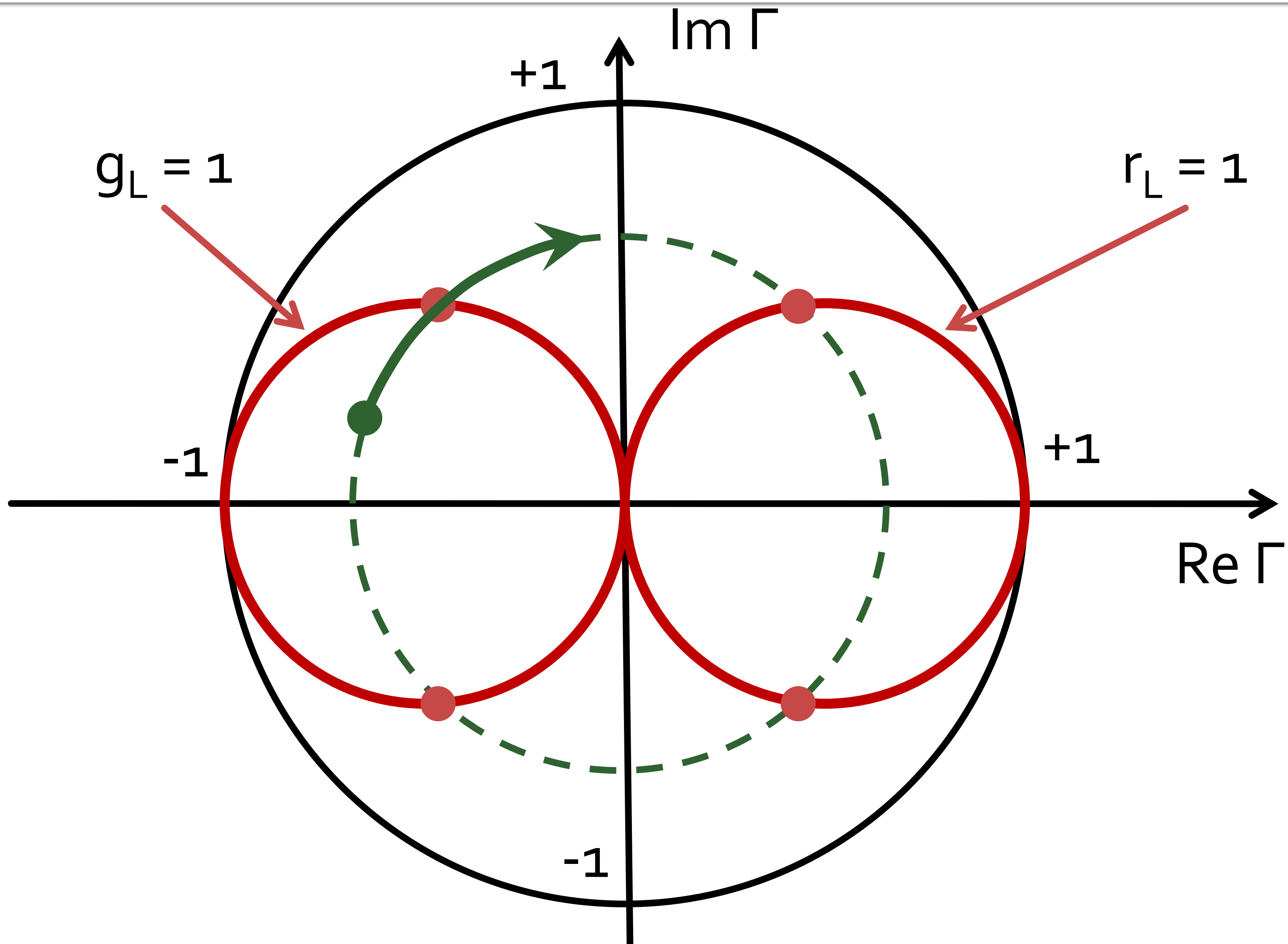


Single stub tuning

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

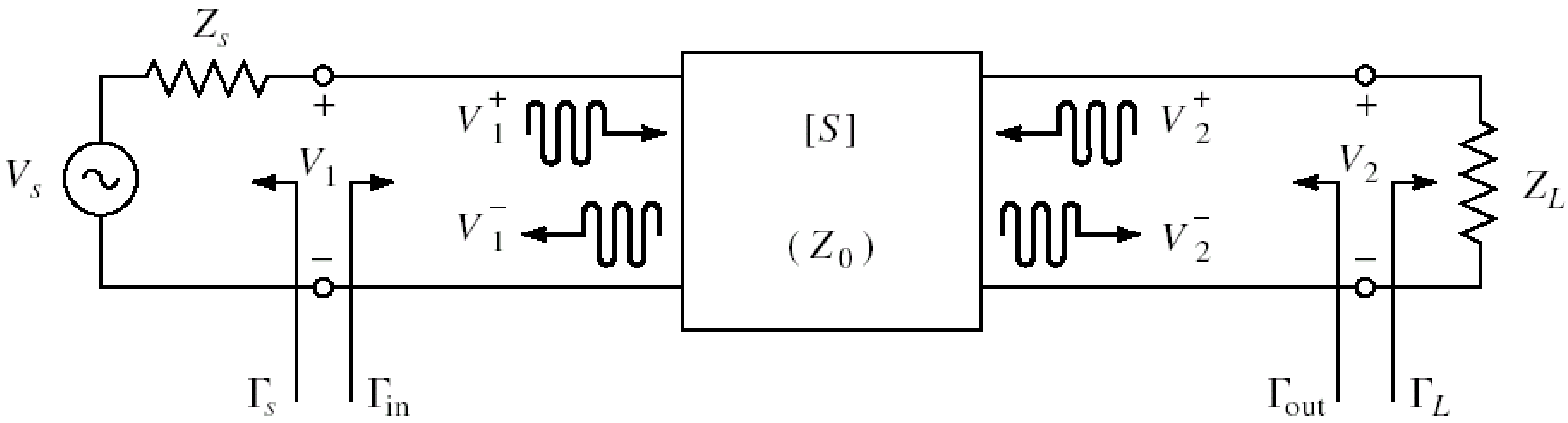


Smith chart, r=1 and g=1



Microwave Amplifiers

Amplifier as two-port



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

Datasheets

NE46100

V_{CE} = 5 V, I_C = 50 mA

FREQUENCY (MHz)	S_{11} MAG	S_{11} ANG	S_{21} MAG	S_{21} ANG	S_{12} MAG	S_{12} ANG	S_{22} MAG	S_{22} 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

V_{CE} = 5 V, I_C = 100 mA

FREQUENCY (MHz)	S_{11} MAG	S_{11} ANG	S_{21} MAG	S_{21} ANG	S_{12} MAG	S_{12} ANG	S_{22} MAG	S_{22} ANG	K	MAG[2] (dB)
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

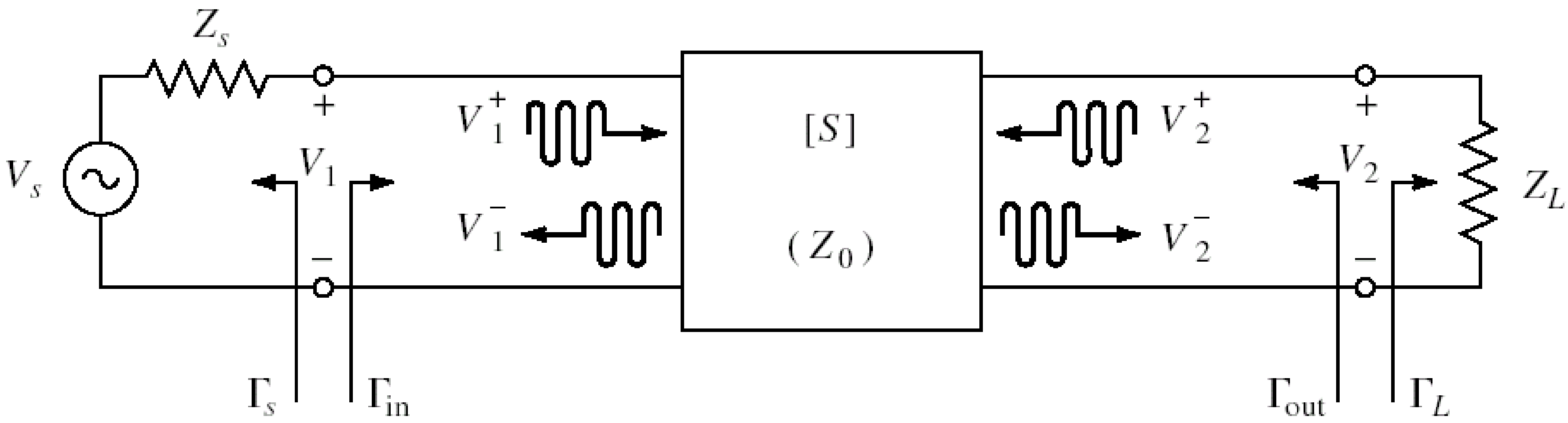
- Touchstone file format (*.s2p)

```
! 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
```

Stability

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)

Stability

$$|\Gamma_{in}| < 1 \qquad \left| S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_L}{1 - S_{22} \cdot \Gamma_L} \right| < 1$$

- The limit between stability/instability

$$|\Gamma_{in}| = 1 \qquad \log_{10}|\Gamma_{in}| = 0 \qquad \left| S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_L}{1 - S_{22} \cdot \Gamma_L} \right| = 1$$

$$|S_{11} \cdot (1 - S_{22} \cdot \Gamma_L) + S_{12} \cdot S_{21} \cdot \Gamma_L| = |1 - S_{22} \cdot \Gamma_L|$$

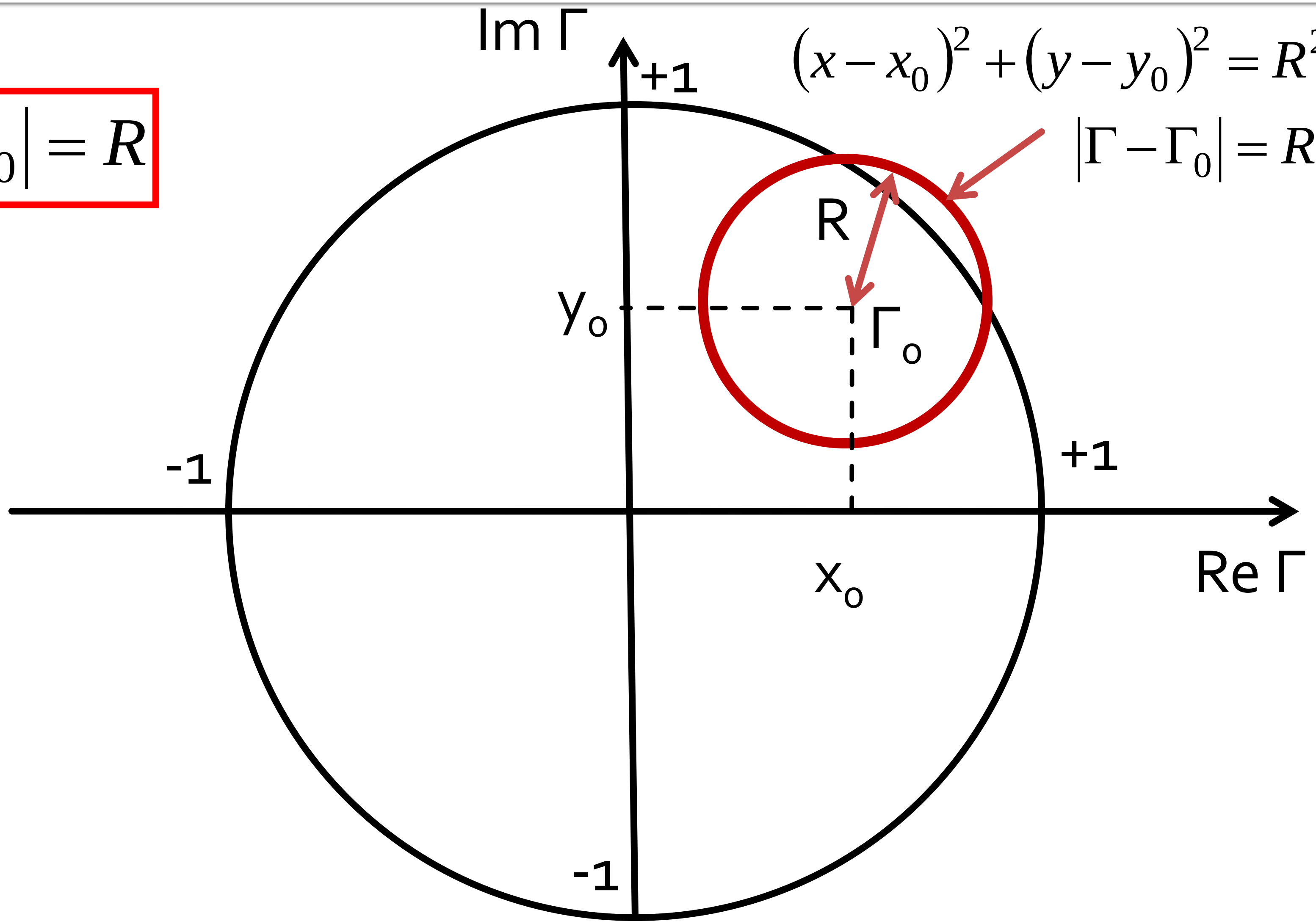
- determinant of the S matrix $\Delta = S_{11} \cdot S_{22} - S_{12} \cdot S_{21}$

$$|S_{11} - \Delta \cdot \Gamma_L| = |1 - S_{22} \cdot \Gamma_L|$$

$$|S_{11} - \Delta \cdot \Gamma_L|^2 = |1 - S_{22} \cdot \Gamma_L|^2$$

Stability

$$|\Gamma - \Gamma_0| = R$$



Output stability circle (CSOUT)

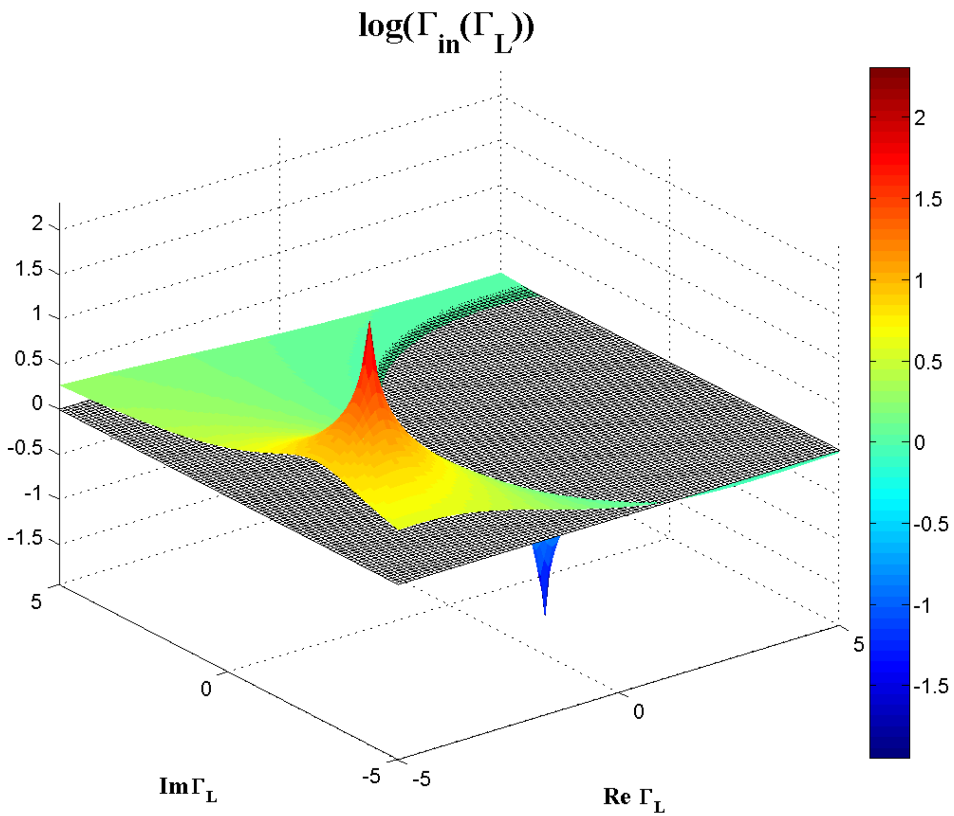
$$\left|\Gamma_L - \frac{\left(S_{22} - \Delta \cdot S_{11}^*\right)^*}{|S_{22}|^2 - |\Delta|^2}\right| = \left|\frac{S_{12} \cdot S_{21}}{|S_{22}|^2 - |\Delta|^2}\right| \qquad |\Gamma_L - C_L| = R_L$$

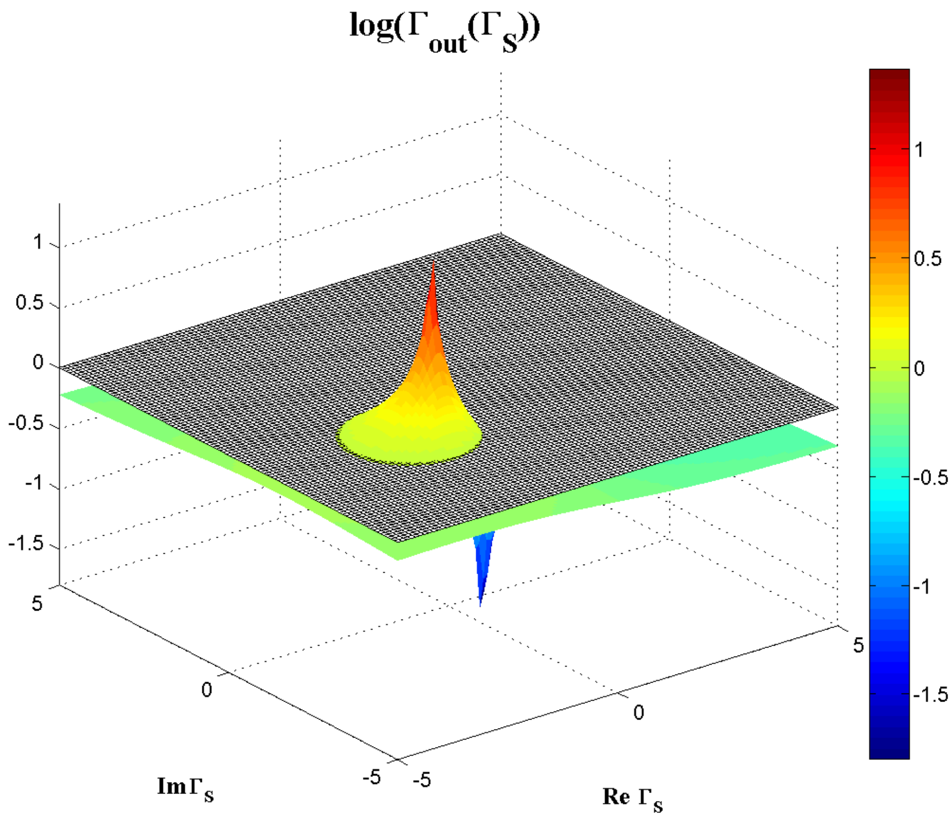
- We obtain the equation of a circle in the complex plane, which represents the locus of Γ_L for the **limit between stability and instability** ($|\Gamma_{in}| = 1$)
- This circle is the **output stability circle** (Γ_L)

$$C_L = \frac{\left(S_{22} - \Delta \cdot S_{11}^*\right)^*}{|S_{22}|^2 - |\Delta|^2} \qquad R_L = \frac{|S_{12} \cdot S_{21}|}{\left||S_{22}|^2 - |\Delta|^2\right|}$$

3D representation of $|\Gamma_{in}|$, $|\Gamma_{out}|$, $|\Gamma|=1$

- $|\Gamma| = 1 \rightarrow \log_{10}|\Gamma| = 0$, the intersection with the plane z = 0 is a circle





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:
  - $|S_{11}| < 1$
  - $|S_{22}| < 1$

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

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

μ Criterion

- Rollet's condition cannot be used to compare the relative stability of two or more devices because it involves constraints on two separate parameters, K and Δ

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

- The two-port is **unconditionally stable** if:
  - $\mu > 1$
- together with the implicit conditions:
  - $|S_{11}| < 1$
  - $|S_{22}| < 1$
- In addition, it can be said that larger values of μ imply greater stability
  - μ is the distance from the center of the Smith Chart to the closest output stability circle

μ′ Criterion

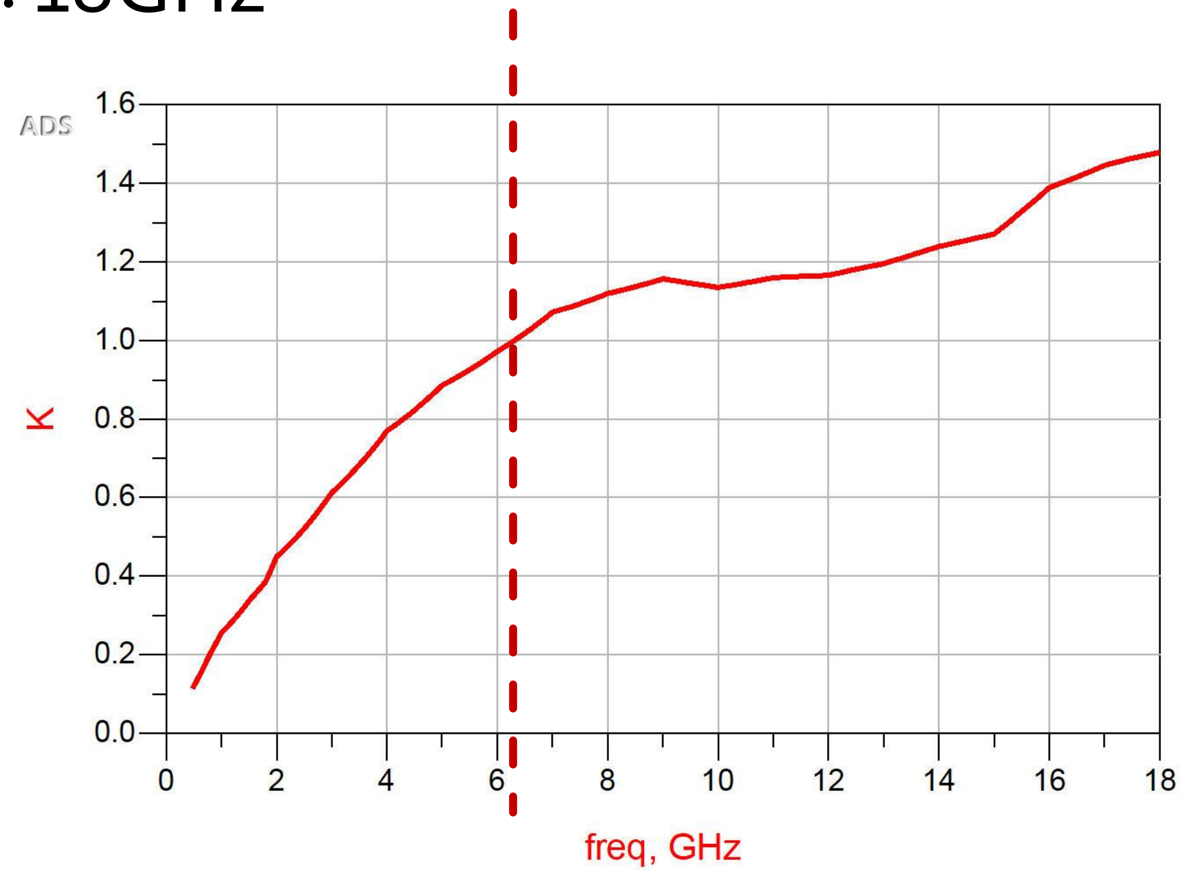
- Dual parameter to μ, determined in relation to the input stability circles

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

- The two-port is **unconditionally stable** if:
  - $\mu' > 1$
- together with the implicit conditions:
  - $|S_{11}| < 1$
  - $|S_{22}| < 1$
- In addition, it can be said that larger values of μ′ imply greater stability
  - μ′ is the distance from the center of the Smith Chart to the closest input stability circle

Rollet's condition

- ATF-34143 **at Vds=3V Id=20mA**.
- @0.5÷18GHz



μ Criterion

- ATF-34143 **at Vds=3V Id=20mA**.
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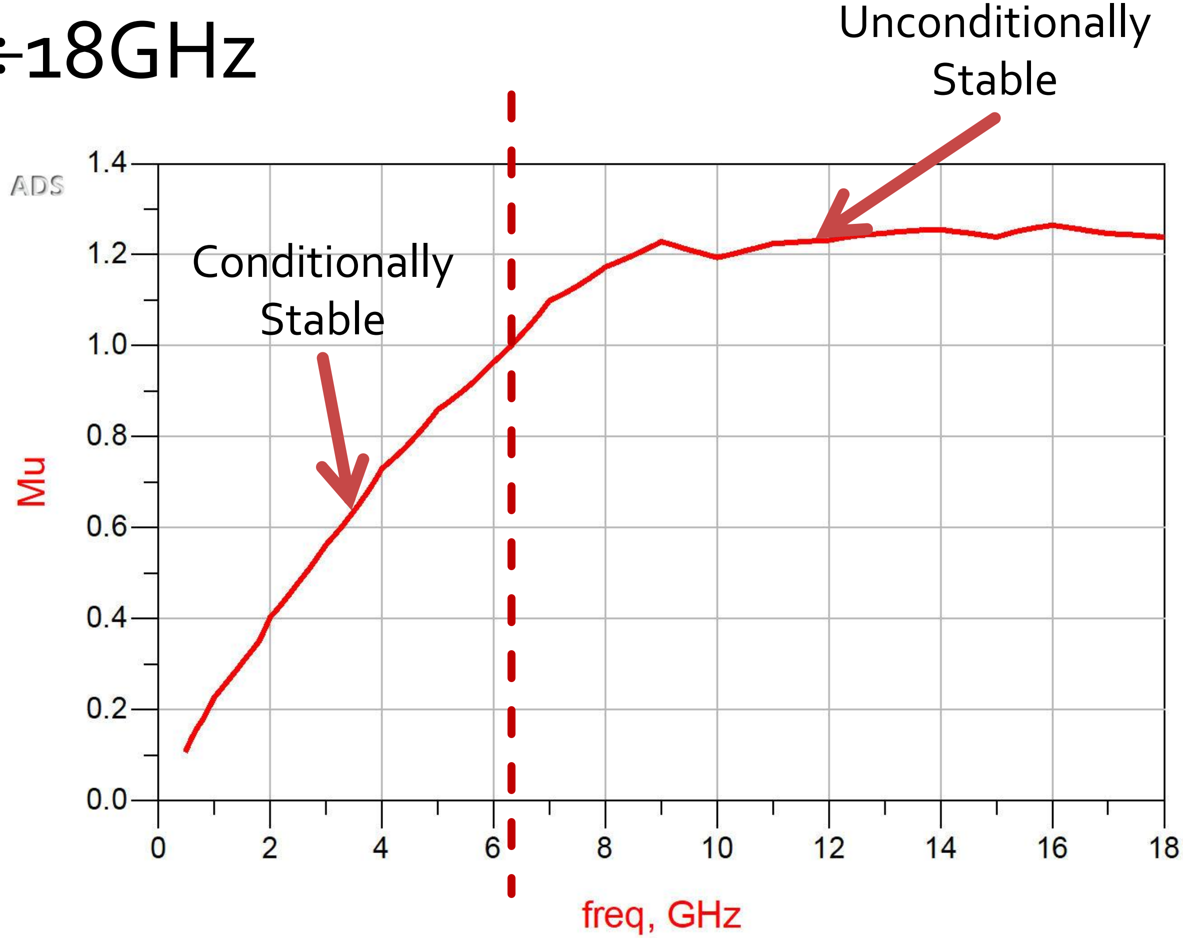
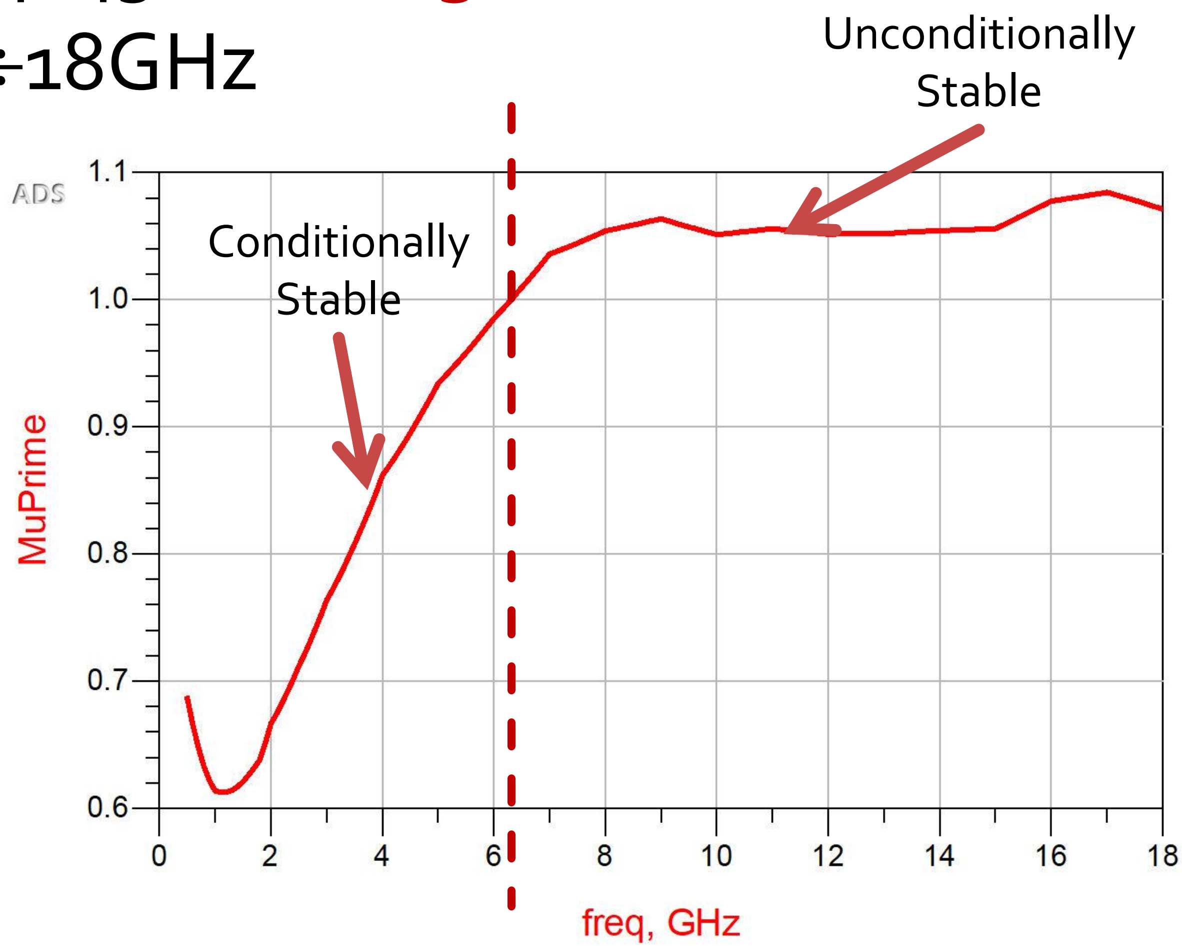


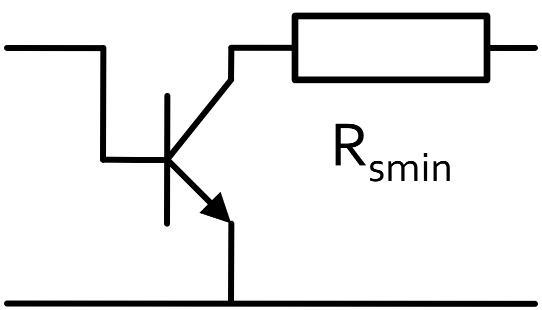
μ' Criterion

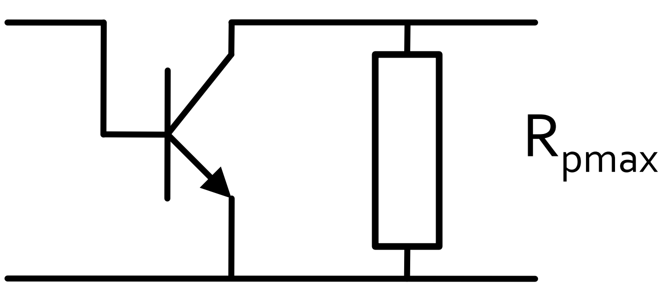
- ATF-34143 **at Vds=3V Id=20mA**.
- @0.5÷18GHz



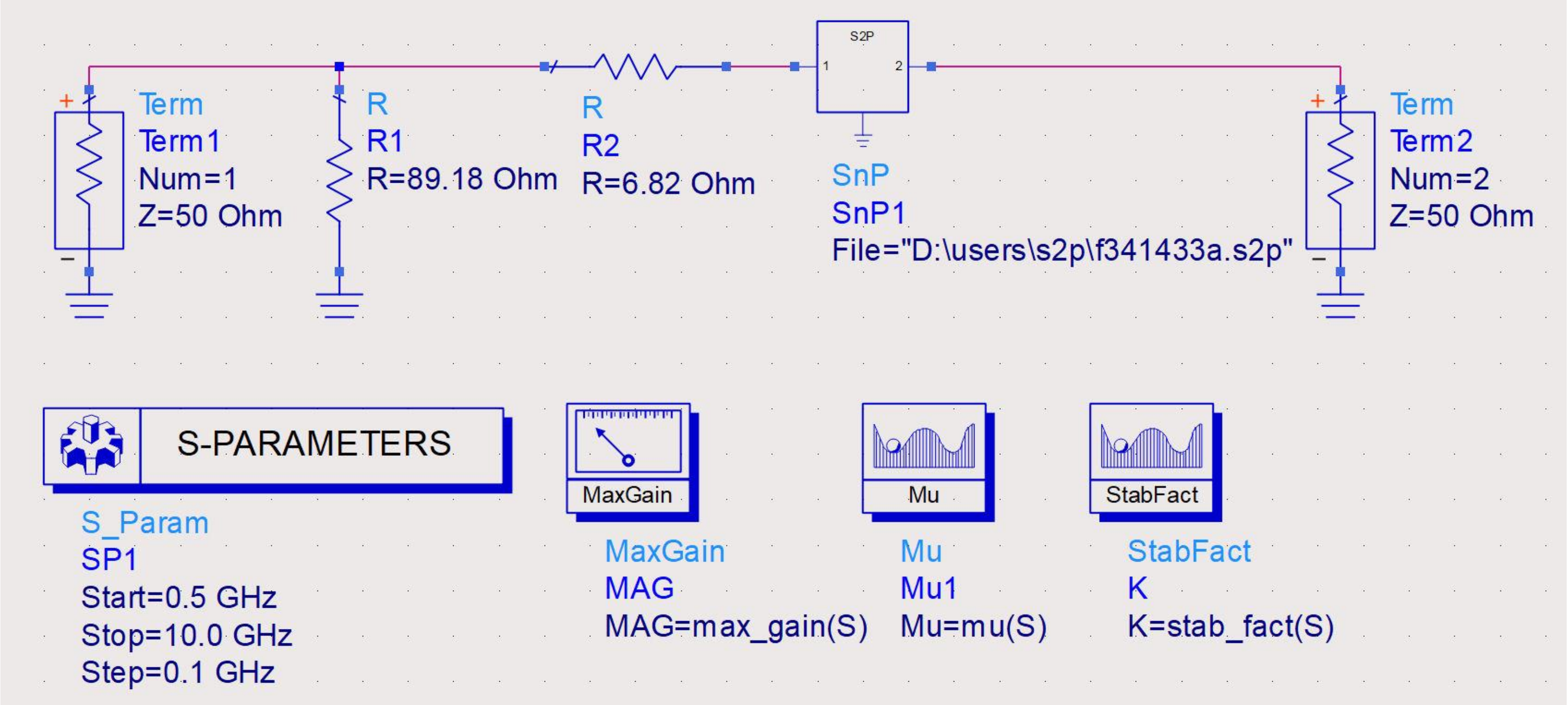
Output series/shunt resistor

- The procedure can be applied similarly at the output (finding g/r circles tangent to CSOUT)
- From previous examples, resistive loading at the input has a positive effect over output stability and vice versa (resistive loading at the output, effect over input stability)

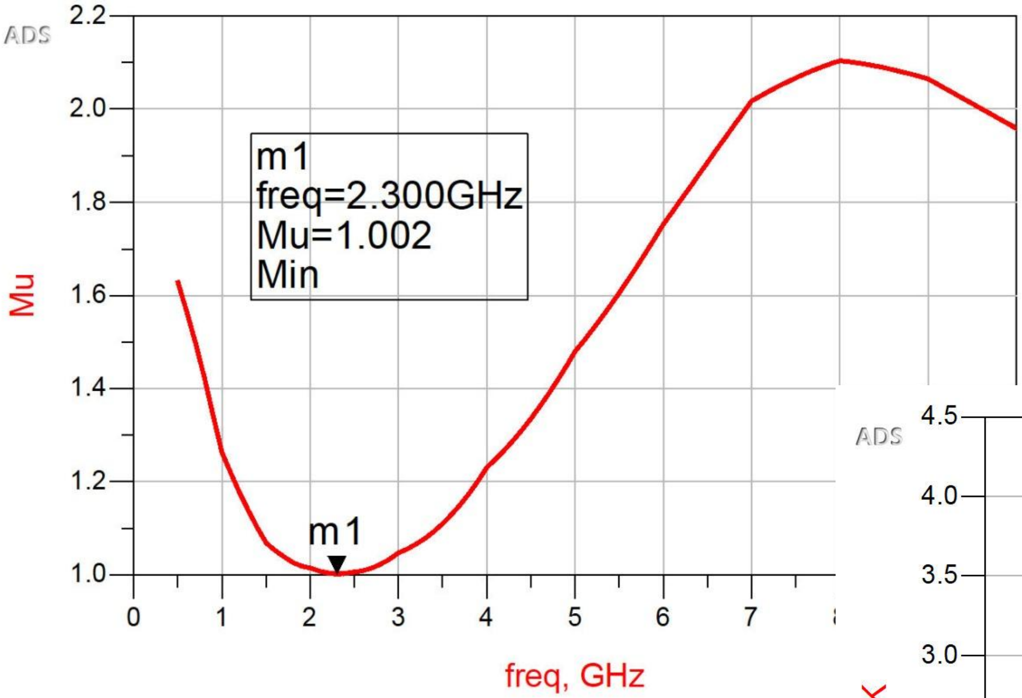


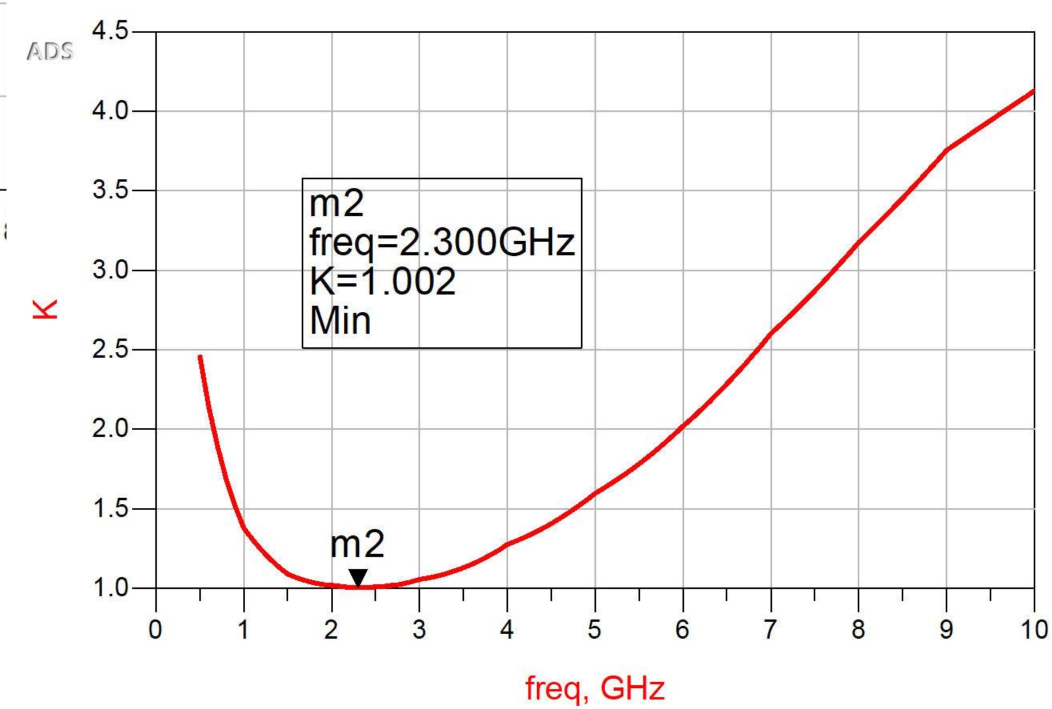


Stabilization of two-port



Stabilization of two-port

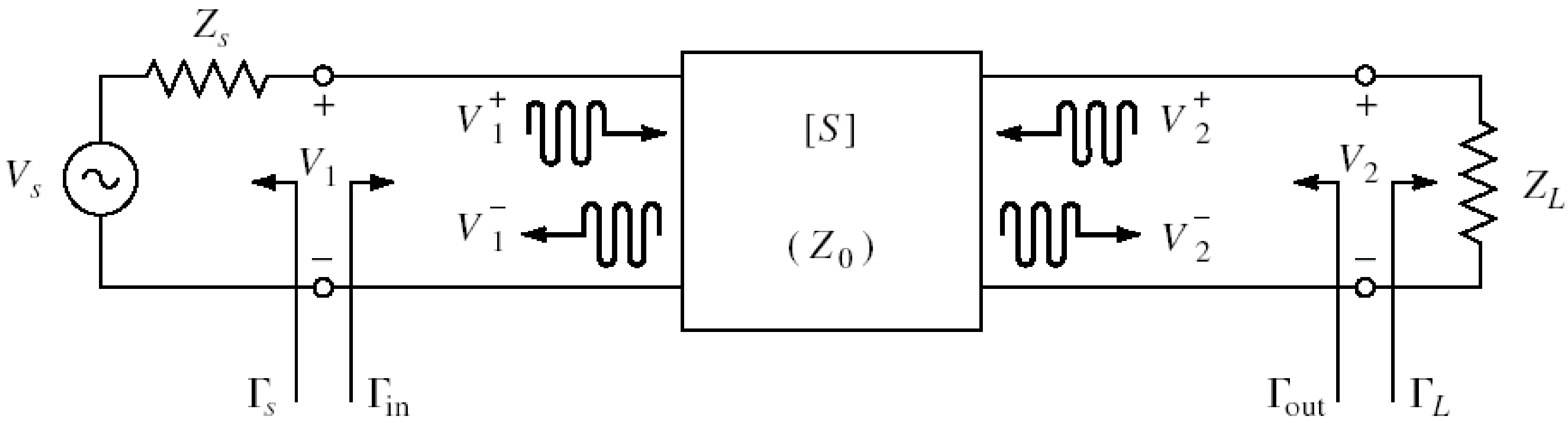




Power Gain of Microwave Amplifiers

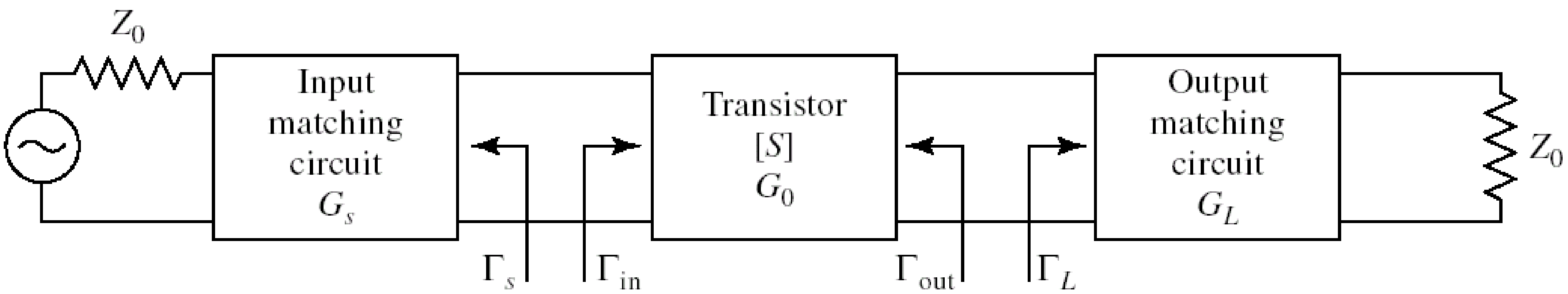
Microwave Amplifiers

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

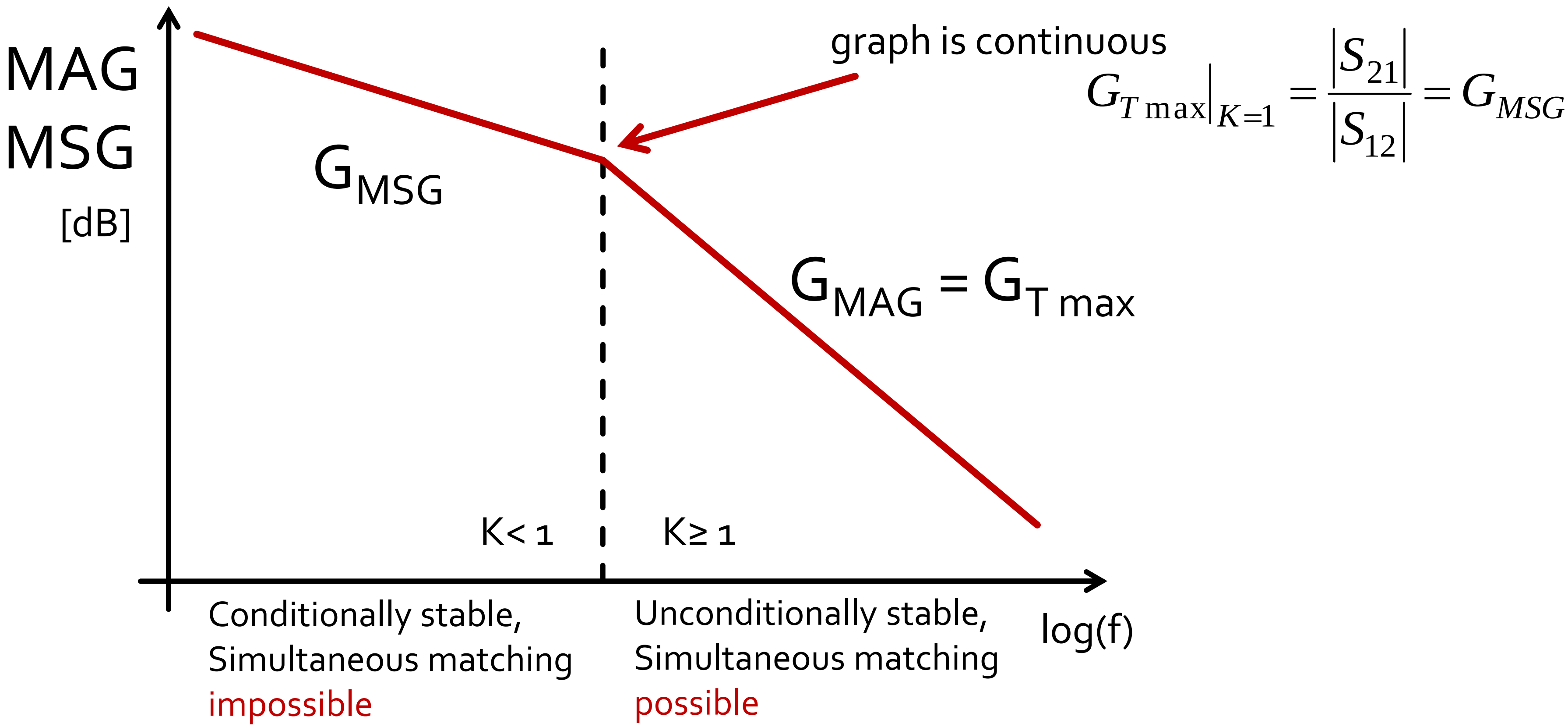
- Simultaneous matching can be achieved **if and only if** the amplifier is **unconditionally stable** at the operating frequency, and |Γ|<1 solutions are those with "–" sign of quadratic solutions

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

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

Maximum Available Gain

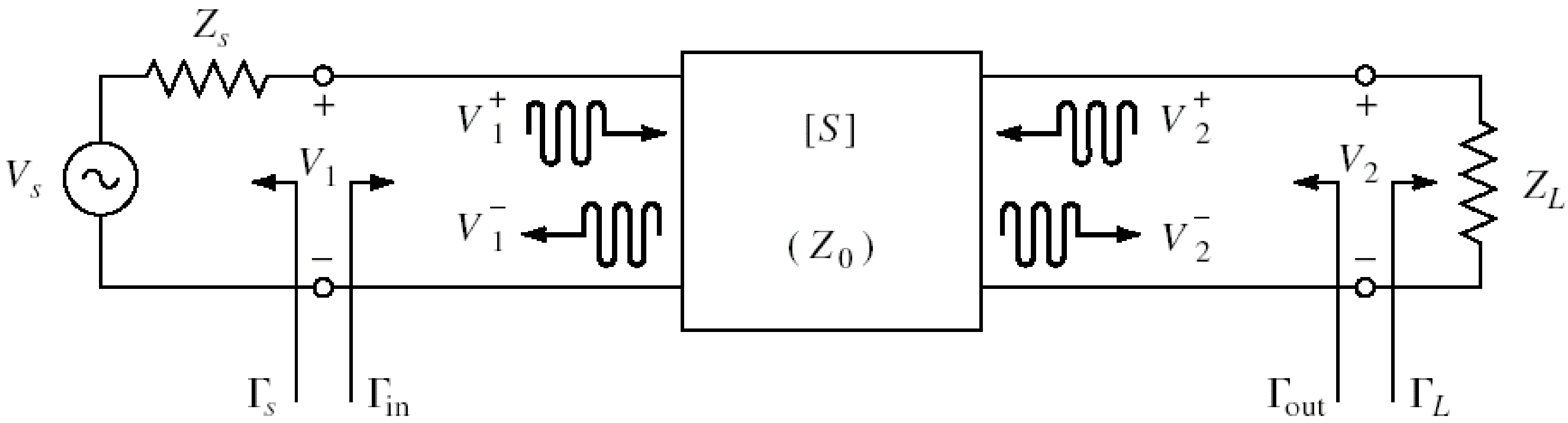
- Indicator across full frequency range of the capability to obtain a power gain



Design for Specified Gain

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)

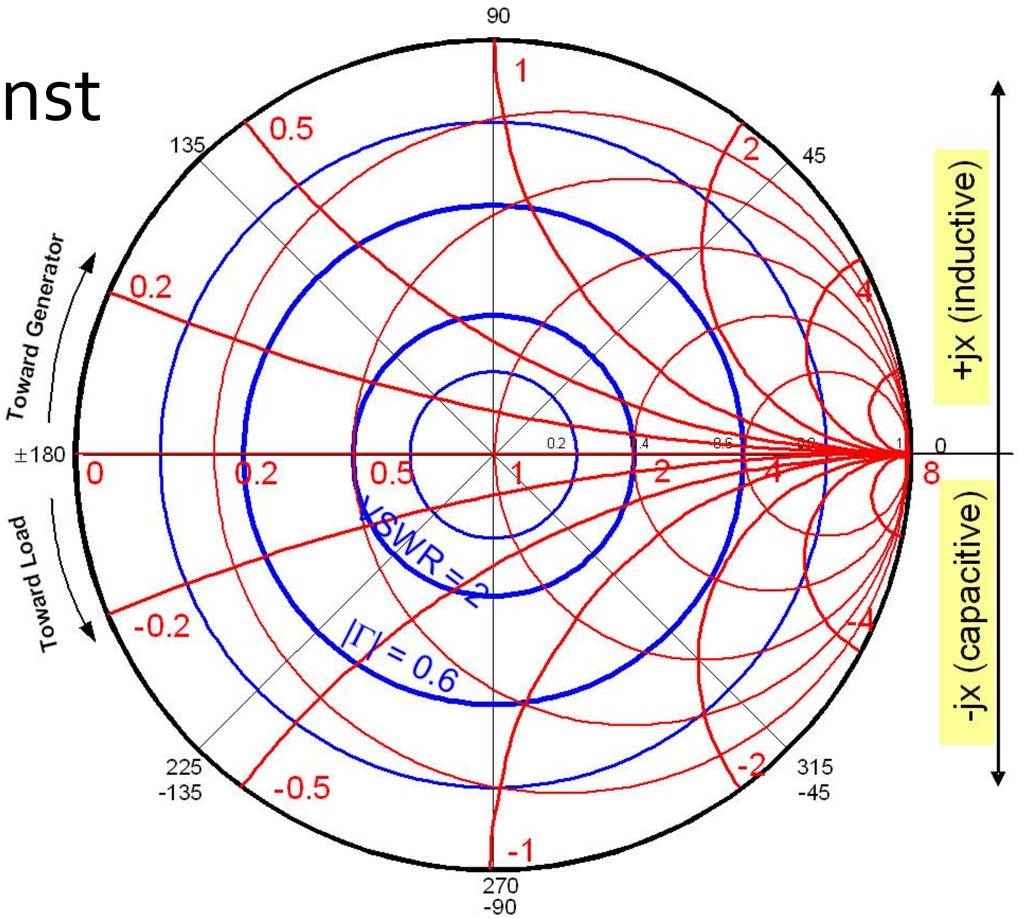
Design for Specified Gain

- 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

Constant VSWR circles

- Certain applications may require a certain ratio between maximum / minimum line voltage
  - VSWR = const → Γ = const

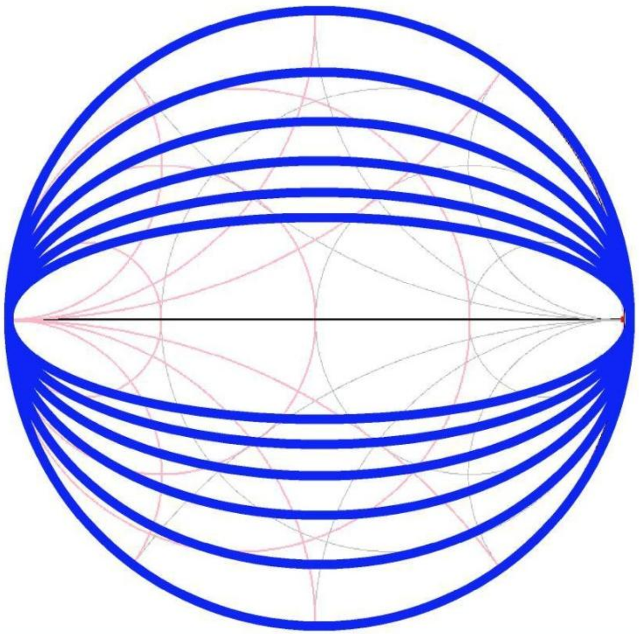
$$VSWR = \frac{V_{max}}{V_{min}} = \frac{1+|\Gamma|}{1-|\Gamma|}$$

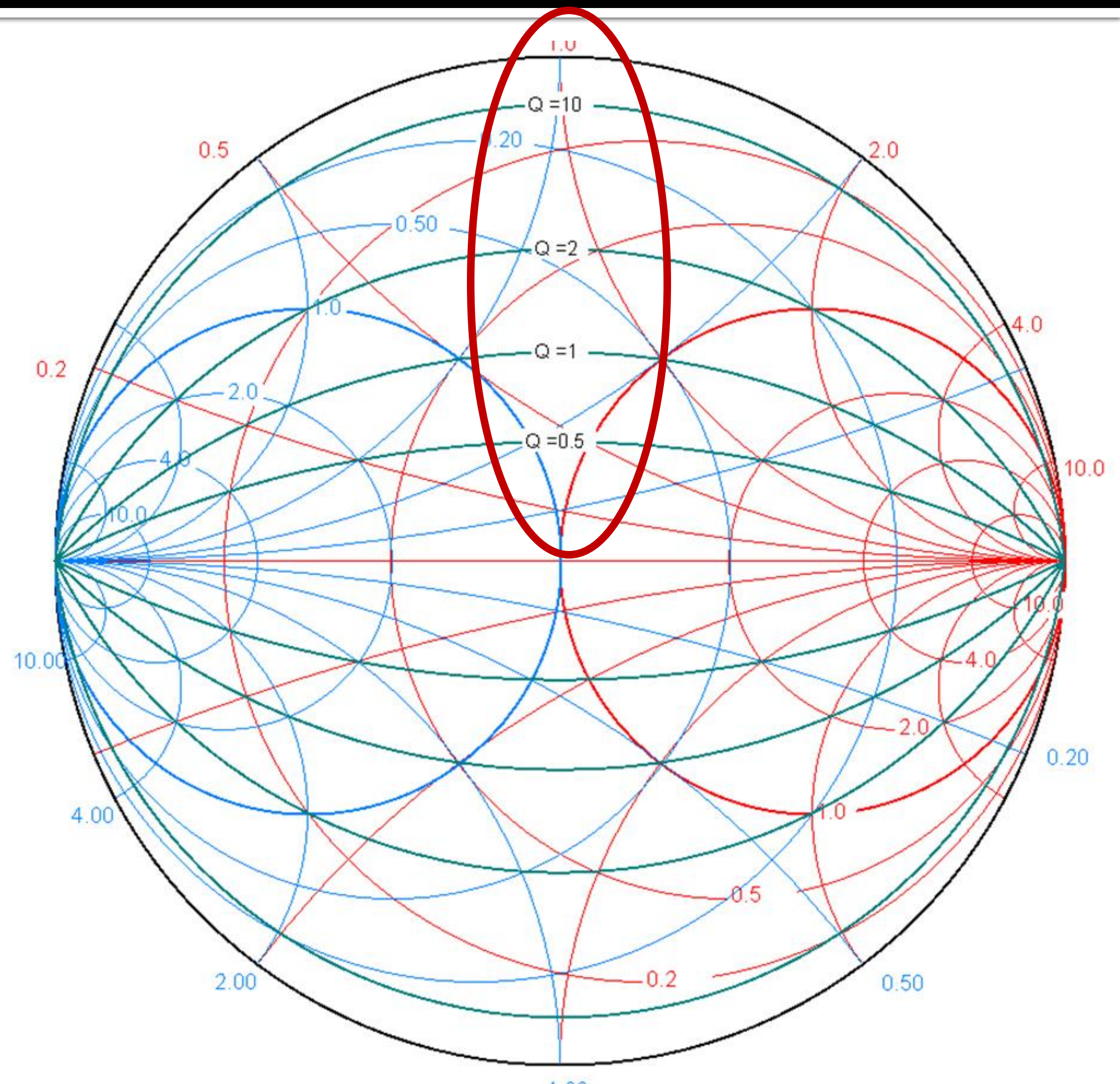


Constant Q circles

- Quality factor Q

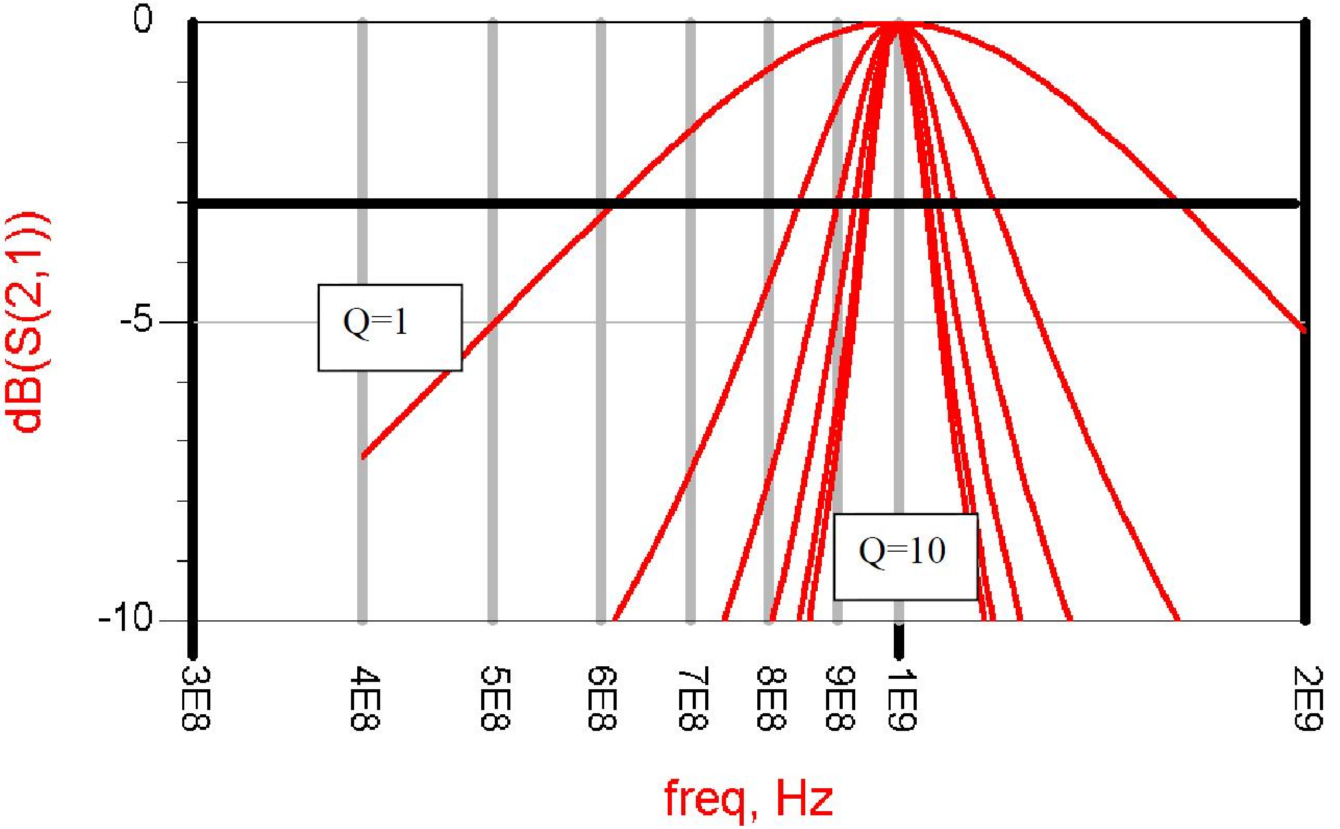
$$Q = \frac{X}{R} = \frac{G}{B} = const$$





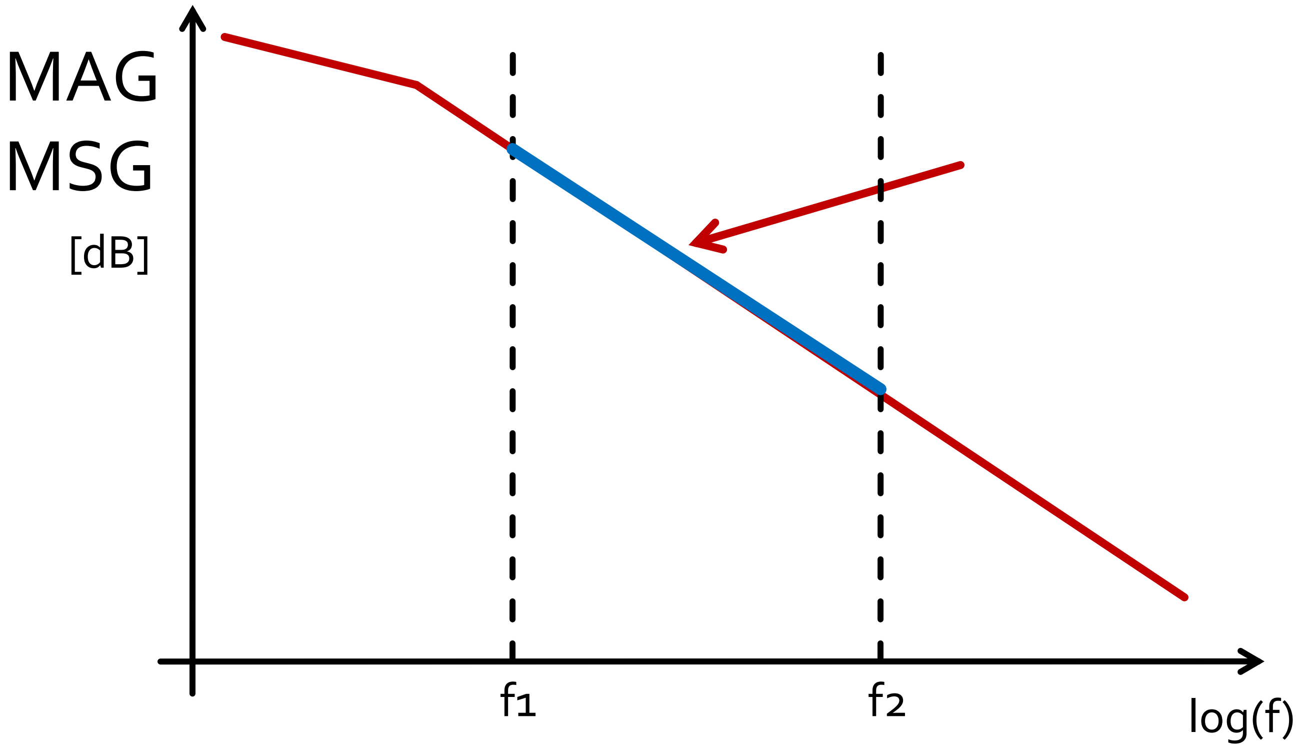
Quality factor - bandwidth

- High quality factor is equivalent with narrow bandwidth



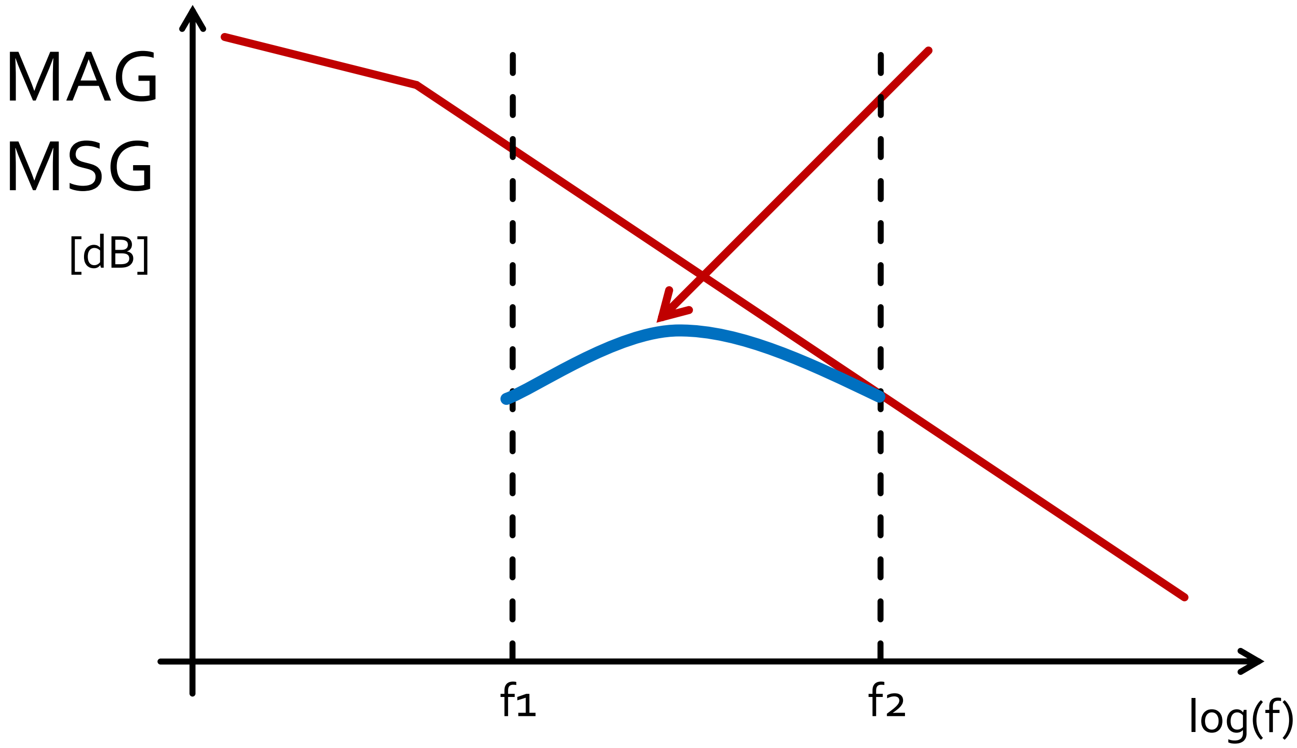
Wide bandwidth amplifier

- Design for maximum gain at two different frequencies creates an frequency unbalanced amplifier



Wide bandwidth amplifier

- Design for maximum gain at highest frequency
- Controlled mismatch at lower frequency
  - eventually at more frequencies inside the bandwidth



Design for Specified Gain

- Assumes the amplifier device **unilateral**

Input and output can be treated independently

$$G_{TU} = |S_{21}|^2 \cdot \frac{1-|\Gamma_S|^2}{|1-S_{11}\cdot\Gamma_S|^2} \cdot \frac{1-|\Gamma_L|^2}{|1-S_{22}\cdot\Gamma_L|^2} \qquad S_{12} \cong 0 \qquad \Gamma_{in} = S_{11}$$

- Maximum power gain

$$\Gamma_S = S_{11}^*$$

$$\Gamma_L = S_{22}^*$$

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

Unilateral figure of merit

- Allows estimation of the error introduced by the unilateral assumption

$$\frac{1}{(1+U)^2} < \frac{G_T}{G_{TU}} < \frac{1}{(1-U)^2} \qquad U = \frac{|S_{12}| \cdot |S_{21}| \cdot |S_{11}| \cdot |S_{22}|}{\left(1-|S_{11}|^2\right) \cdot \left(1-|S_{22}|^2\right)}$$

- We compute U then the maximum and minimum deviation of G_{TU} from G_T
  - this deviation must be accounted in the design as a reserve gain against the target gain

$$-20 \cdot \log(1+U) < G_T[dB] - G_{TU}[dB] < -20 \cdot \log(1-U)$$

Example

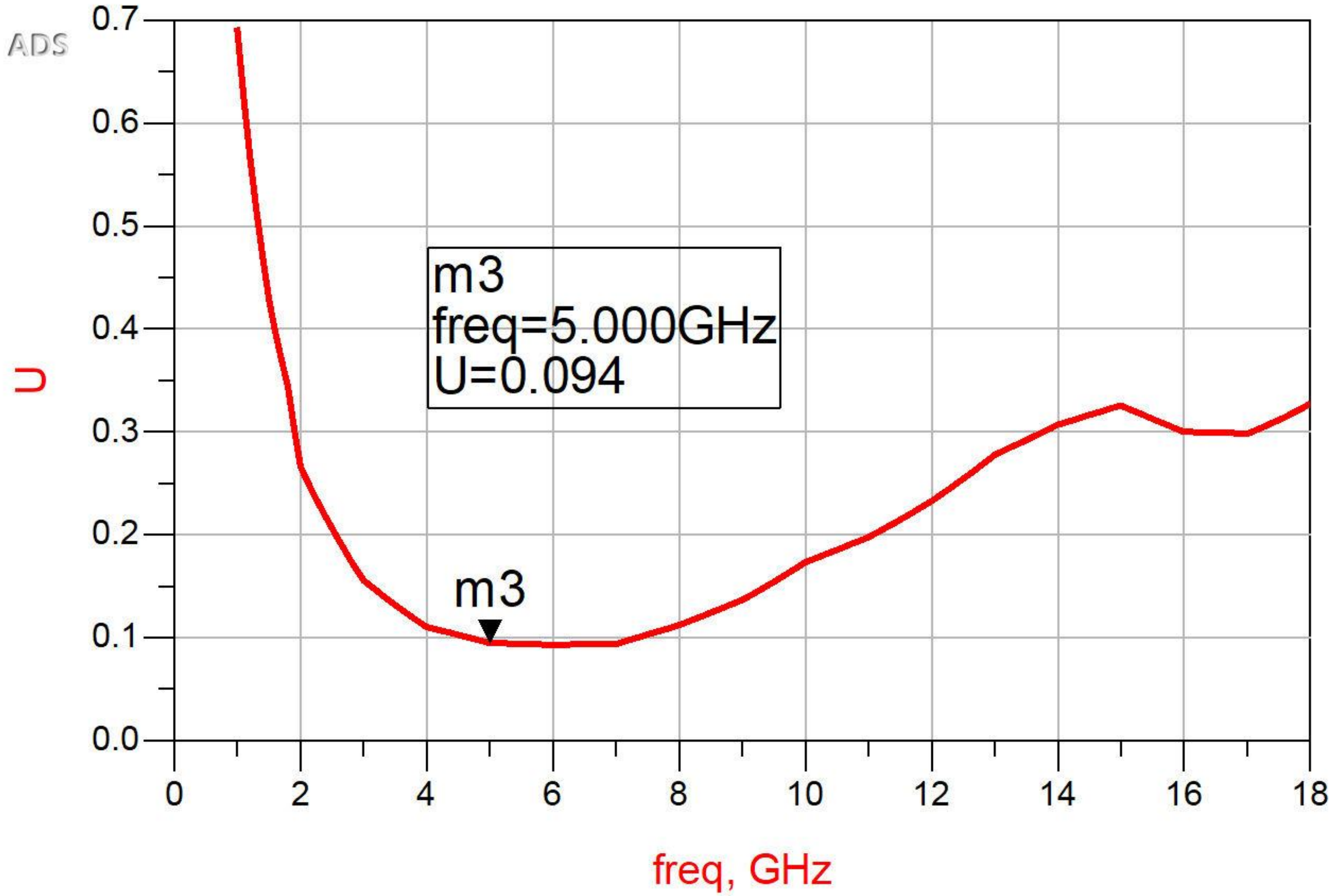
- ATF-34143 **at Vds=3V Id=20mA**.
- @5GHz
  - S11 = 0.64∠139°
  - S12 = 0.119∠-21°
  - S21 = 3.165 ∠16°
  - S22 = 0.22 ∠146°

$$U = \frac{|S_{12}|\cdot|S_{21}|\cdot|S_{11}|\cdot|S_{22}|}{\left(1-|S_{11}|^2\right)\cdot\left(1-|S_{22}|^2\right)} = 0.094$$

$$-0.783\, dB < G_T[dB] - G_{TU}[dB] < 0.861\, dB$$

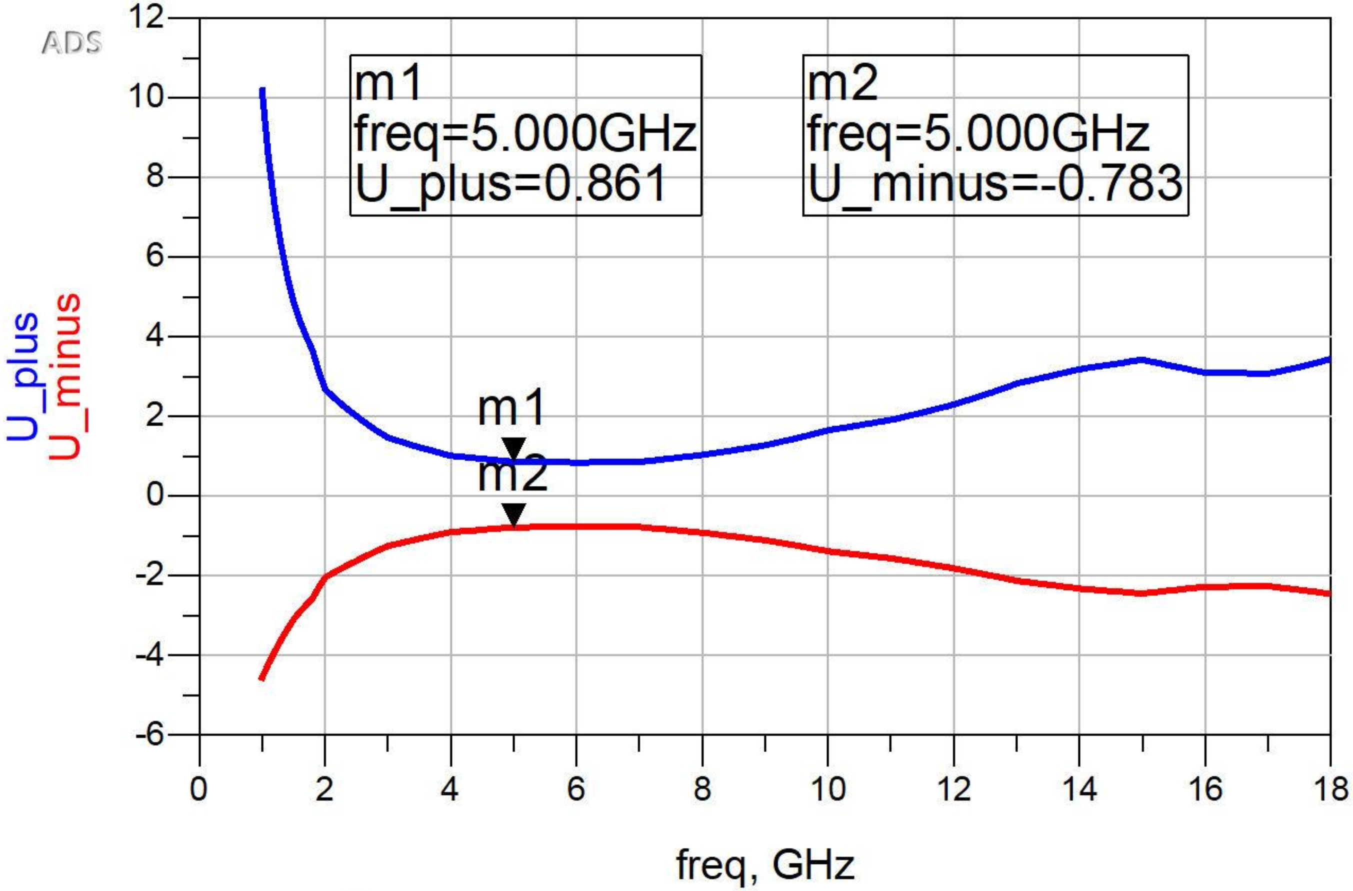
Example

- ATF-34143 **at Vds=3V Id=20mA**.
- @5GHz

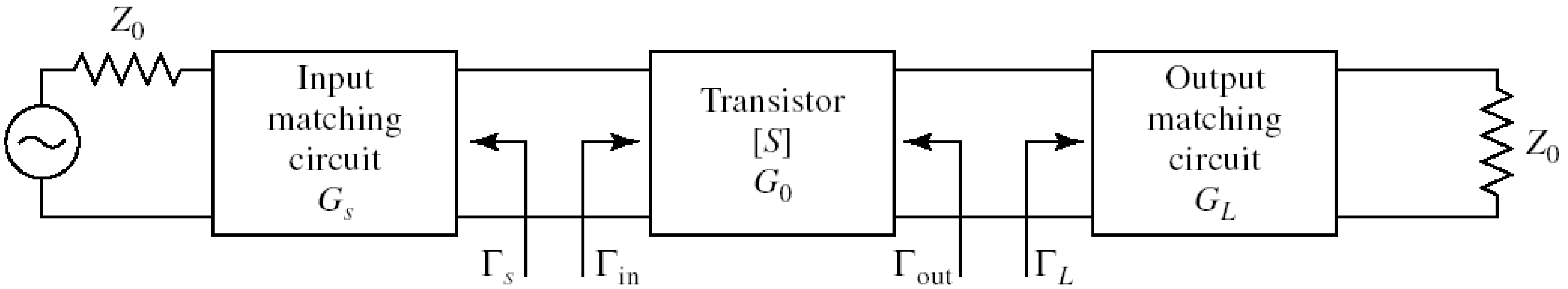


Example

- ATF-34143 **at Vds=3V Id=20mA**.
- @5GHz



Design for Specified Gain



- In the unilateral assumption:

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

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

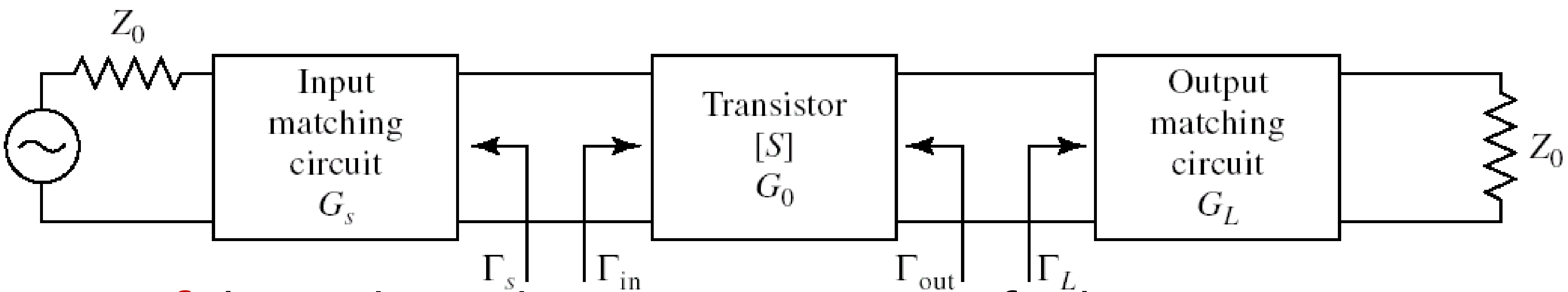
$$G_S = G_S(\Gamma_S)$$

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

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

$$G_L = G_L(\Gamma_L)$$

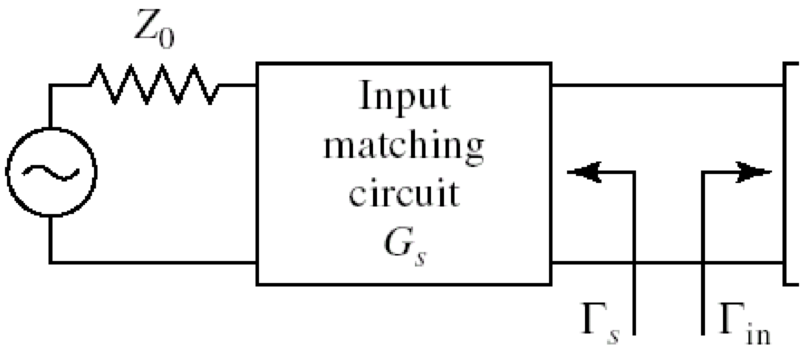
Design for Specified Gain



- **If** the unilateral assumption is justified :
  - power gain added by the input matching circuit **is not** influenced by the output matching circuit $G_S = G_S(\Gamma_S)$
  - power gain added by the output matching circuit **is not** influenced by the input matching circuit $G_L = G_L(\Gamma_L)$
- Output /Input match can be designed independently
  - We can impose different demands for input/output
  - Total gain is:

$$G_T = G_S \cdot G_0 \cdot G_L \qquad G_T[dB] = G_S[dB] + G_0[dB] + G_L[dB]$$

Input matching circuit



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

- Maximum gain in the case of complex conjugate match

$$\Gamma_S = S_{11}^* \Rightarrow \qquad G_{S\max} = \frac{1}{1-|S_{11}|^2}$$

- For any other input matching circuit:

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

Example

- ATF-34143 **at Vds=3V Id=20mA**.
- @5GHz
  - S11 = 0.64∠139°
  - S12 = 0.119∠-21°
  - S21 = 3.165 ∠16°
  - S22 = 0.22 ∠146°

$$U = \frac{|S_{12}|\cdot|S_{21}|\cdot|S_{11}|\cdot|S_{22}|}{\left(1-|S_{11}|^2\right)\cdot\left(1-|S_{22}|^2\right)} = 0.094$$

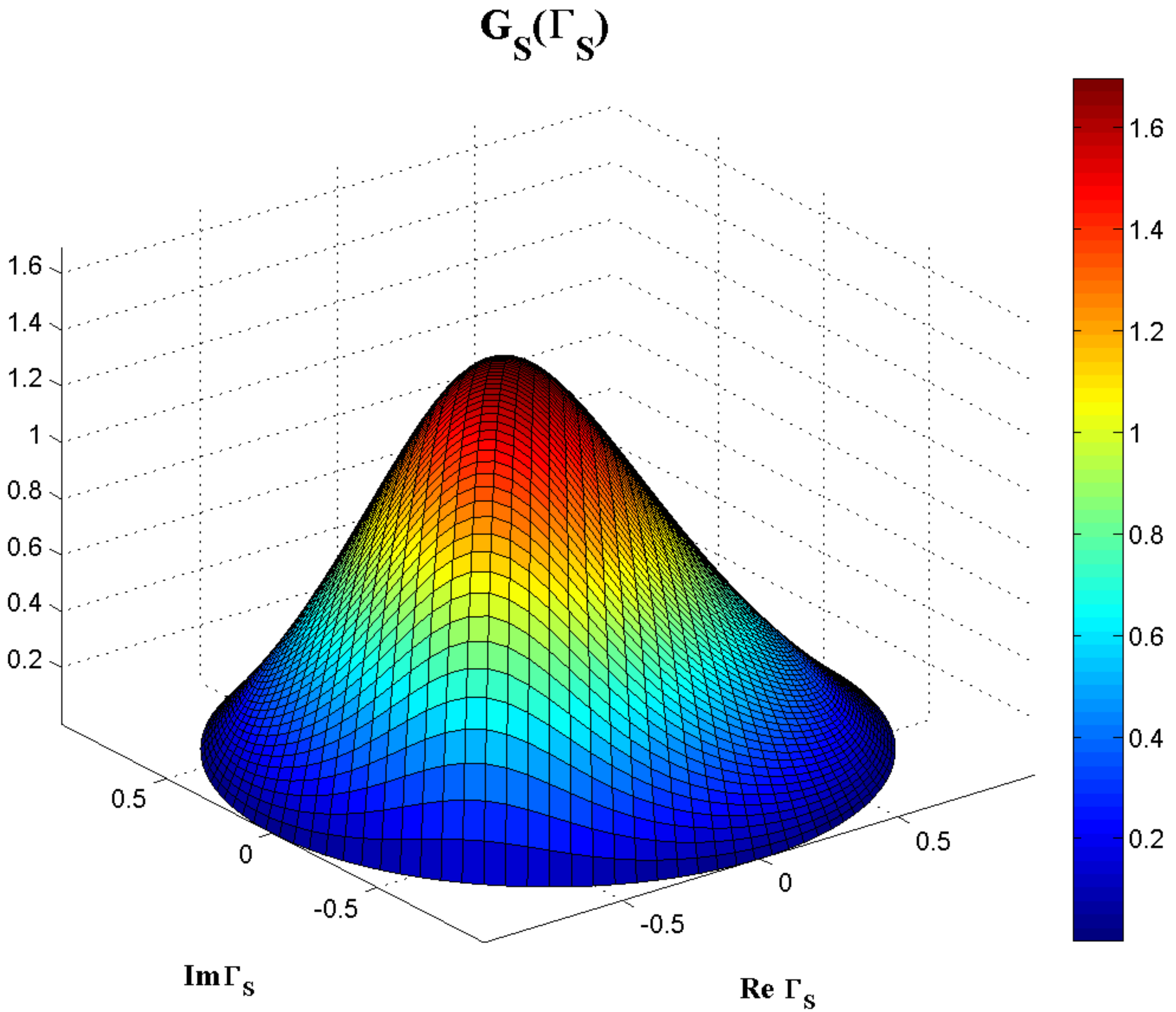
$$-0.783\,dB < G_T[dB] - G_{TU}[dB] < 0.861\,dB$$

$$G_{TU\,\max} = \frac{1}{1-|S_{11}|^2}\cdot|S_{21}|^2\cdot\frac{1}{1-|S_{22}|^2} = 17.83$$

$$G_{TU\,\max}[dB] = 12.511\,dB$$

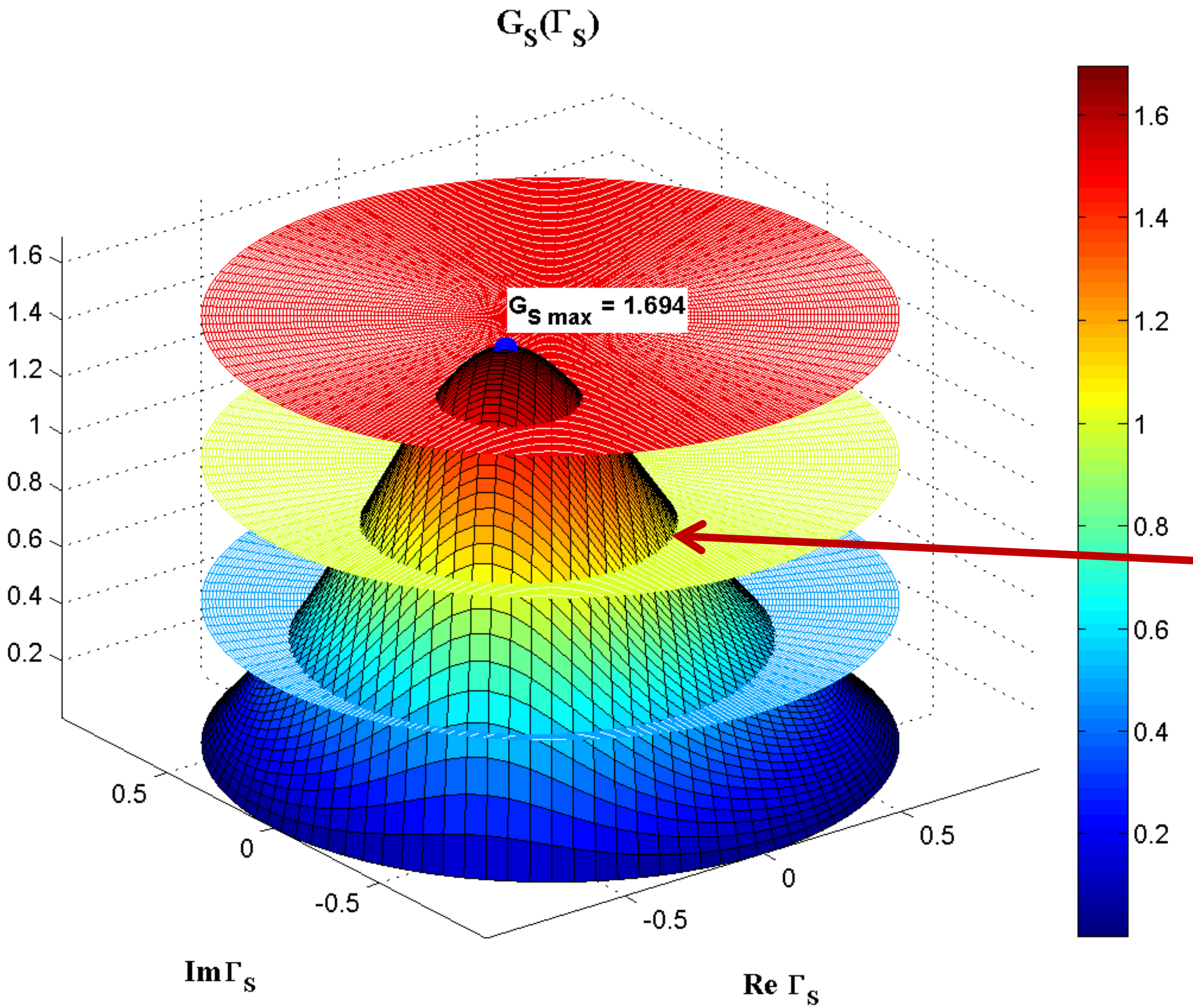
$$G_{S\,\max} = \frac{1}{1-|S_{11}|^2} = 1.694 = 2.289\,dB$$

$G_S(\Gamma_S)$



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

$G_S(\Gamma_S)$, constant value contours



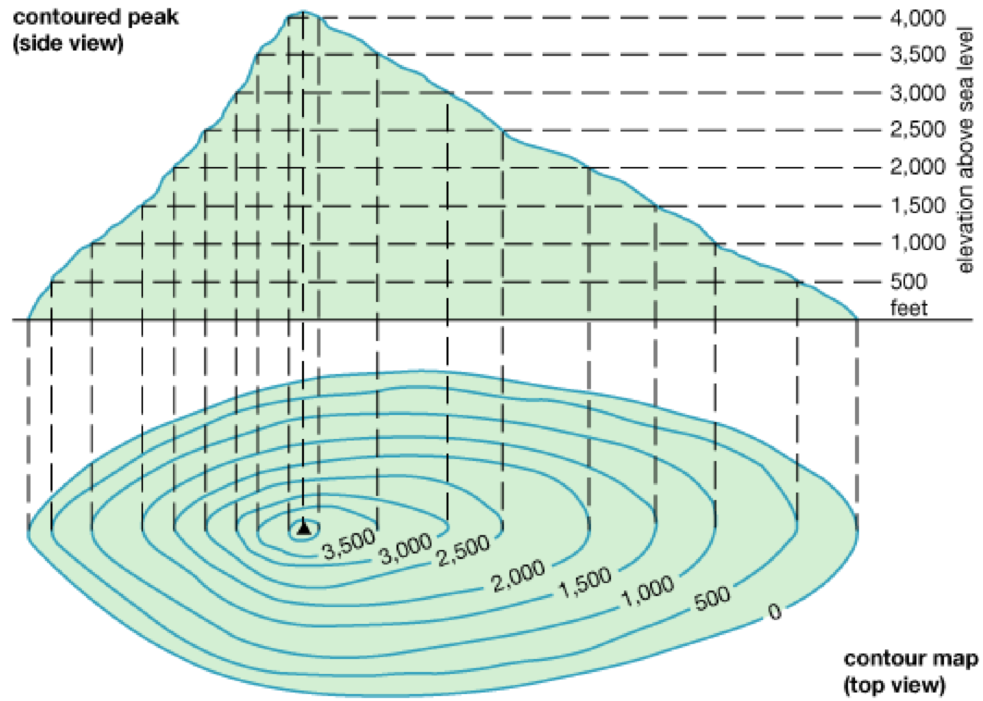
$G_S = 1.5$

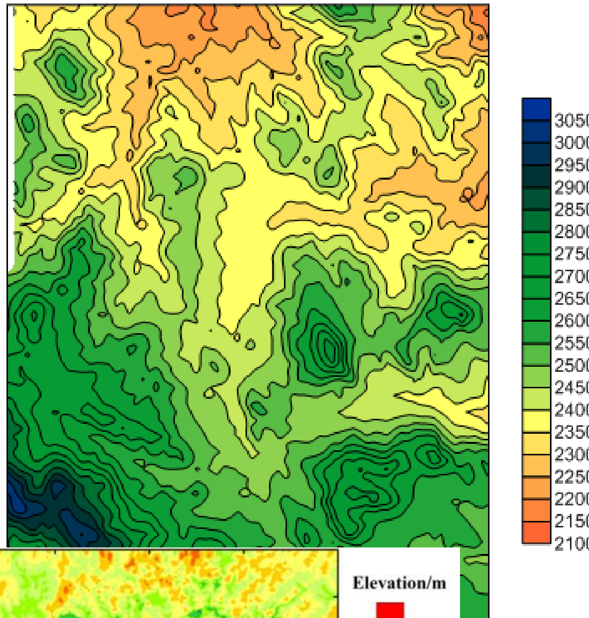
$G_S = 1.0$

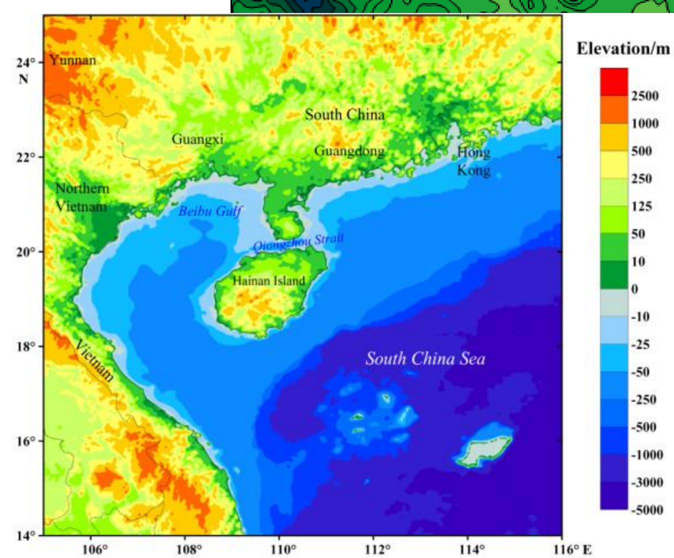
$G_S = 0.5$

Circles

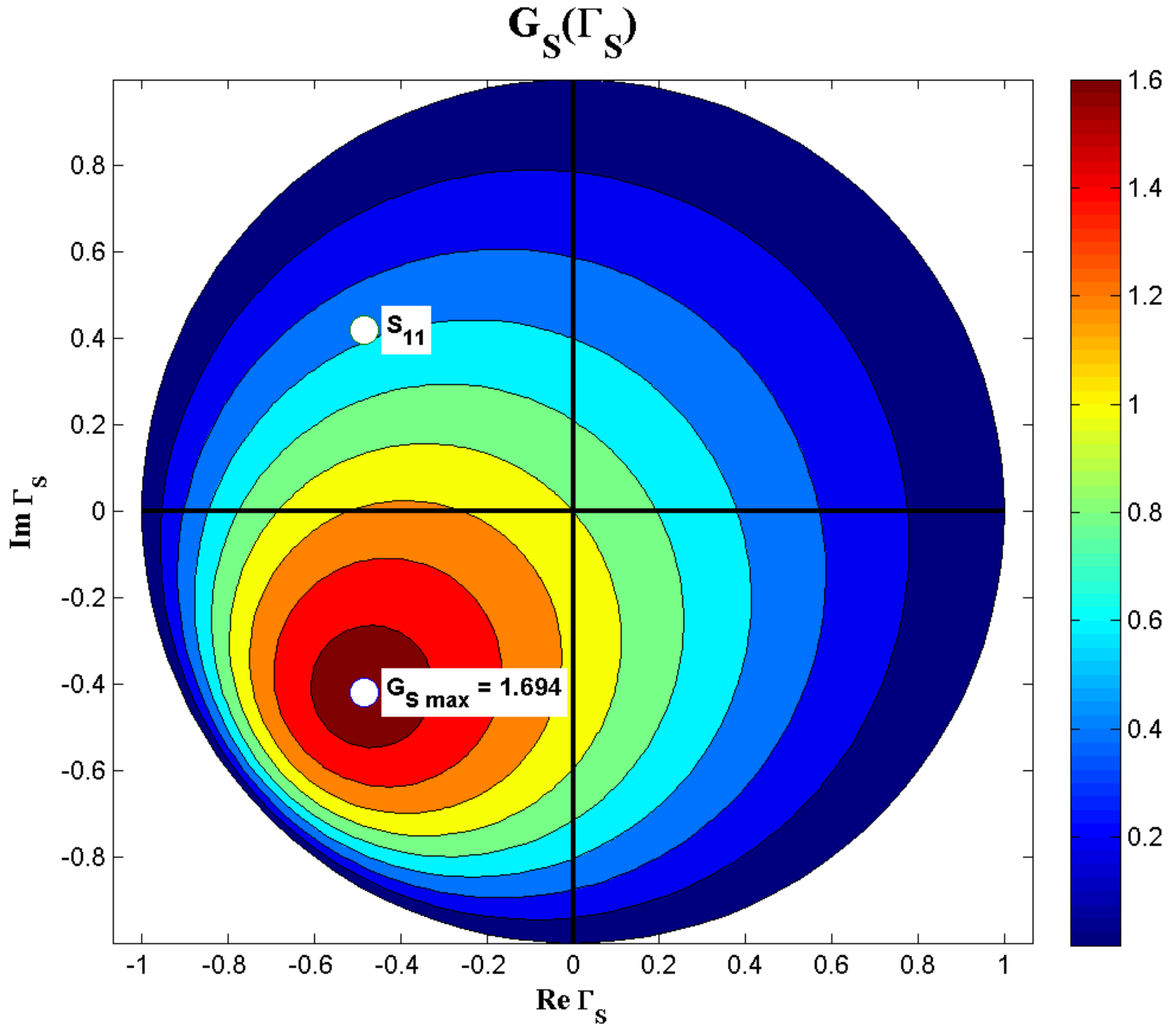
Contour map/lines







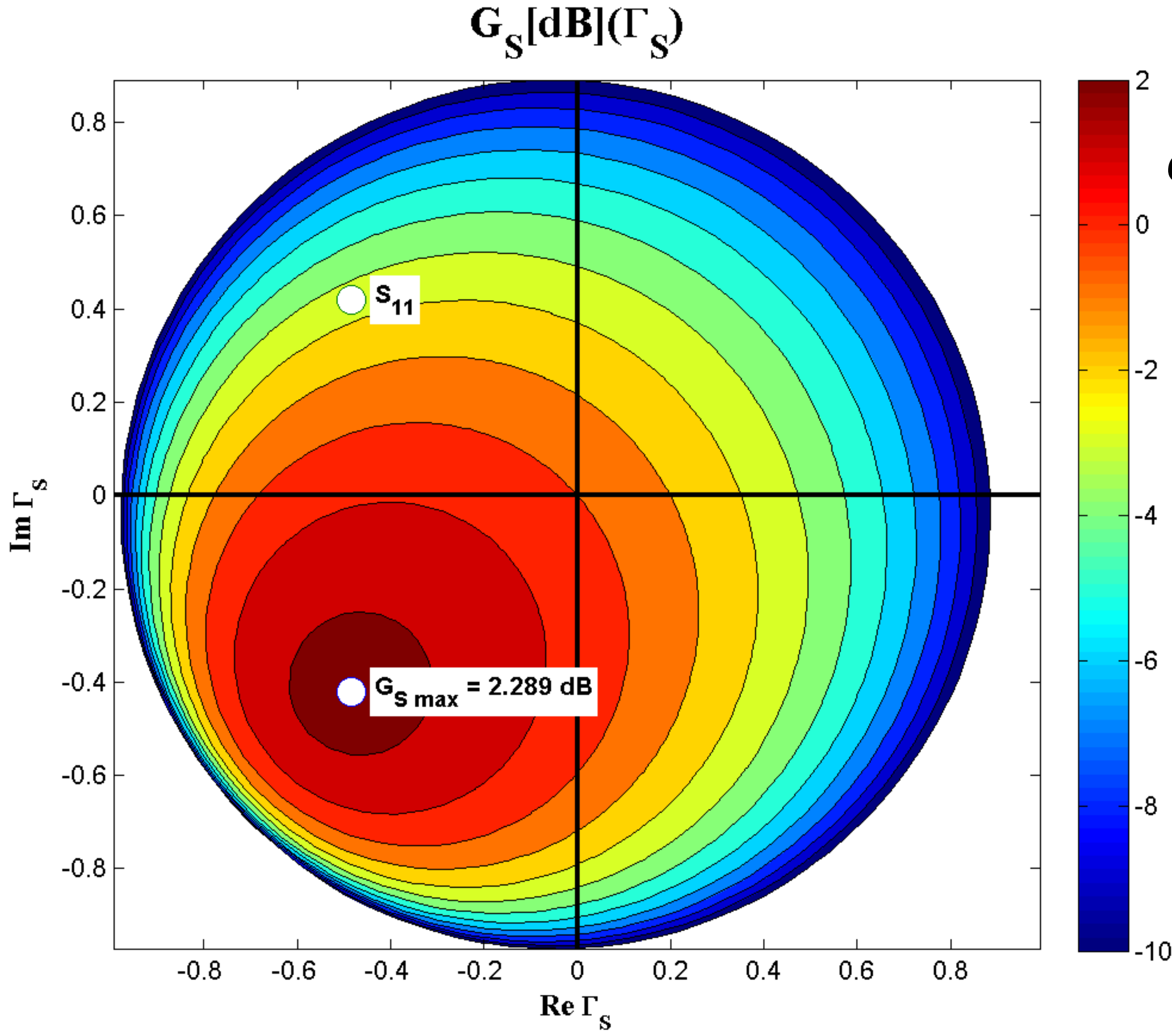
$G_S(\Gamma_S)$, constant value contours



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

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

G_S[dB](Γ_S), constant value contours



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

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

Input section constant gain circles

- The normalized gain factor (**linear scale!**)

$$g_S = \frac{G_S}{G_{S\max}} = \frac{1-|\Gamma_S|^2}{|1-S_{11}\cdot\Gamma_S|^2}\cdot\left(1-|S_{11}|^2\right) < 1$$

- Locus of the points with fixed values $g_s<1$

$$g_S\cdot|1-S_{11}\cdot\Gamma_S|^2 = \left(1-|\Gamma_S|^2\right)\cdot\left(1-|S_{11}|^2\right)$$

$$\left(g_S\cdot|S_{11}|^2+1-|S_{11}|^2\right)\cdot|\Gamma_S|^2 - g_S\cdot\left(S_{11}\cdot\Gamma_S + S_{11}^*\cdot\Gamma_S^*\right) = 1-|S_{11}|^2 - g_S$$

$$\Gamma_S\cdot\Gamma_S^* - \frac{g_S\cdot\left(S_{11}\cdot\Gamma_S + S_{11}^*\cdot\Gamma_S^*\right)}{1-(1-g_S)\cdot|S_{11}|^2} = \frac{1-|S_{11}|^2-g_S}{1-(1-g_S)\cdot|S_{11}|^2} \quad \Big/ + \frac{g_S^2\cdot|S_{11}|^2}{\left[1-(1-g_S)\cdot|S_{11}|^2\right]^2}$$

Input section constant gain circles

$$\left|\Gamma_S - \frac{g_S \cdot S_{11}^*}{1-(1-g_S)\cdot|S_{11}|^2}\right| = \frac{\sqrt{1-g_S}\cdot\left(1-|S_{11}|^2\right)}{1-(1-g_S)\cdot|S_{11}|^2} \qquad |\Gamma_S - C_S| = R_S$$

$$C_S = \frac{g_S \cdot S_{11}^*}{1-(1-g_S)\cdot|S_{11}|^2} \qquad R_S = \frac{\sqrt{1-g_S}\cdot\left(1-|S_{11}|^2\right)}{1-(1-g_S)\cdot|S_{11}|^2}$$

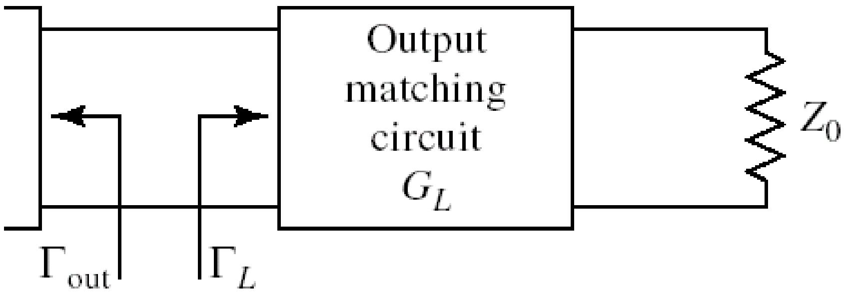
- Equation of a circle in the complex plane where Γ_S is plotted
- **Interpretation**: Any reflection coefficient Γ_S which plotted in the complex plane lies **on** the circle drawn for $g_{circle} = G_{circle}/G_{Smax}$ will lead to a gain $G_S = G_{circle}$
  - Any reflection coefficient Γ_S plotted **outside** this circle will lead to a gain $G_S < G_{circle}$
  - Any reflection coefficient Γ_S plotted **inside** this circle will lead to a gain $G_S > G_{circle}$

Input section constant gain circles

$$C_S = \frac{g_S \cdot S_{11}^*}{1-(1-g_S)\cdot|S_{11}|^2} \qquad R_S = \frac{\sqrt{1-g_S}\cdot\left(1-|S_{11}|^2\right)}{1-(1-g_S)\cdot|S_{11}|^2}$$

- The centers of each family of circles lie along straight lines given by the angle of $\Gamma_{S\max} = S_{11}^*$
- Circles are plotted (traditionally, CAD) in **logarithmic scale** ([dB])
  - formulas are in **linear scale!**
- The circle for G_S = 0 dB will always pass through the origin of the complex plane (center of the Smith chart)

Output section constant gain circles



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

- Maximum gain for $\Gamma_L = S_{22}^* \Rightarrow$ $G_{L\max} = \dfrac{1}{1-|S_{22}|^2}$

$$g_L = \frac{G_L}{G_{L\max}} = \frac{1-|\Gamma_L|^2}{|1-S_{22}\cdot\Gamma_L|^2}\cdot\left(1-|S_{22}|^2\right) < 1$$

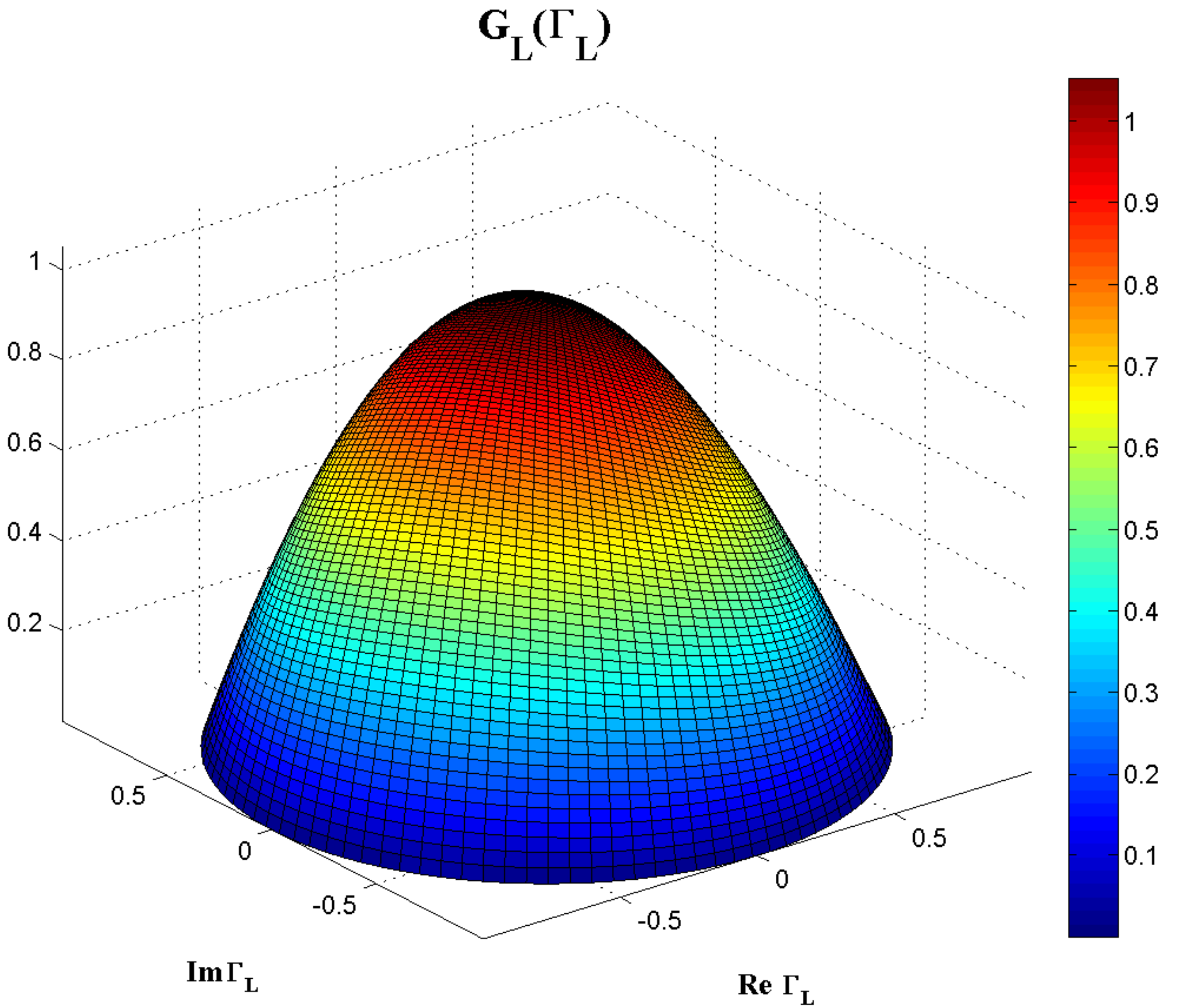
- Similar computations

$$C_L = \frac{g_L \cdot S_{22}^*}{1-(1-g_L)\cdot|S_{22}|^2} \qquad R_L = \frac{\sqrt{1-g_L}\cdot\left(1-|S_{22}|^2\right)}{1-(1-g_L)\cdot|S_{22}|^2}$$

- Example

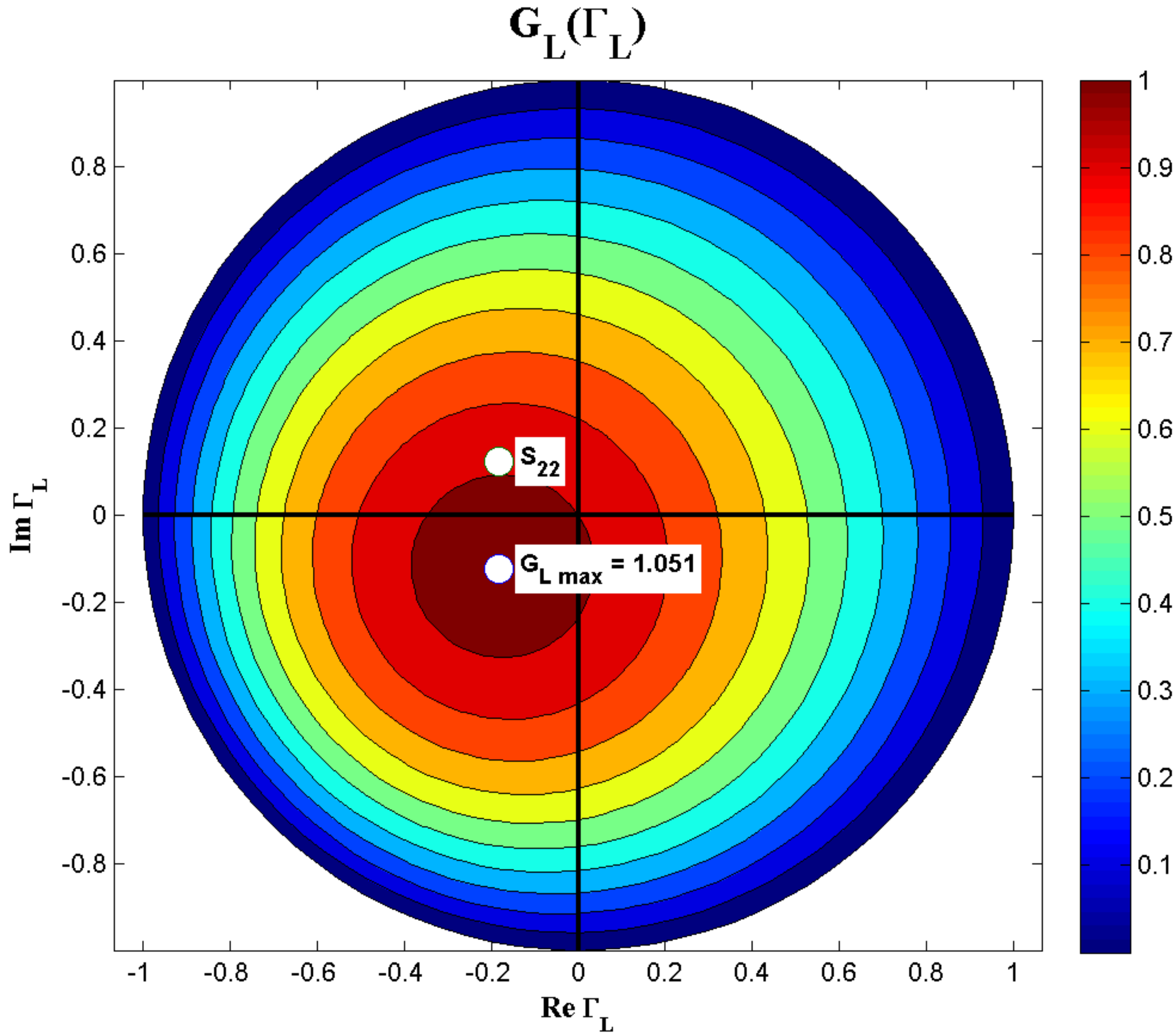
$$G_{L\max} = \frac{1}{1-|S_{22}|^2} = 1.051 = 0.215\, dB$$

$G_L(\Gamma_L)$



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

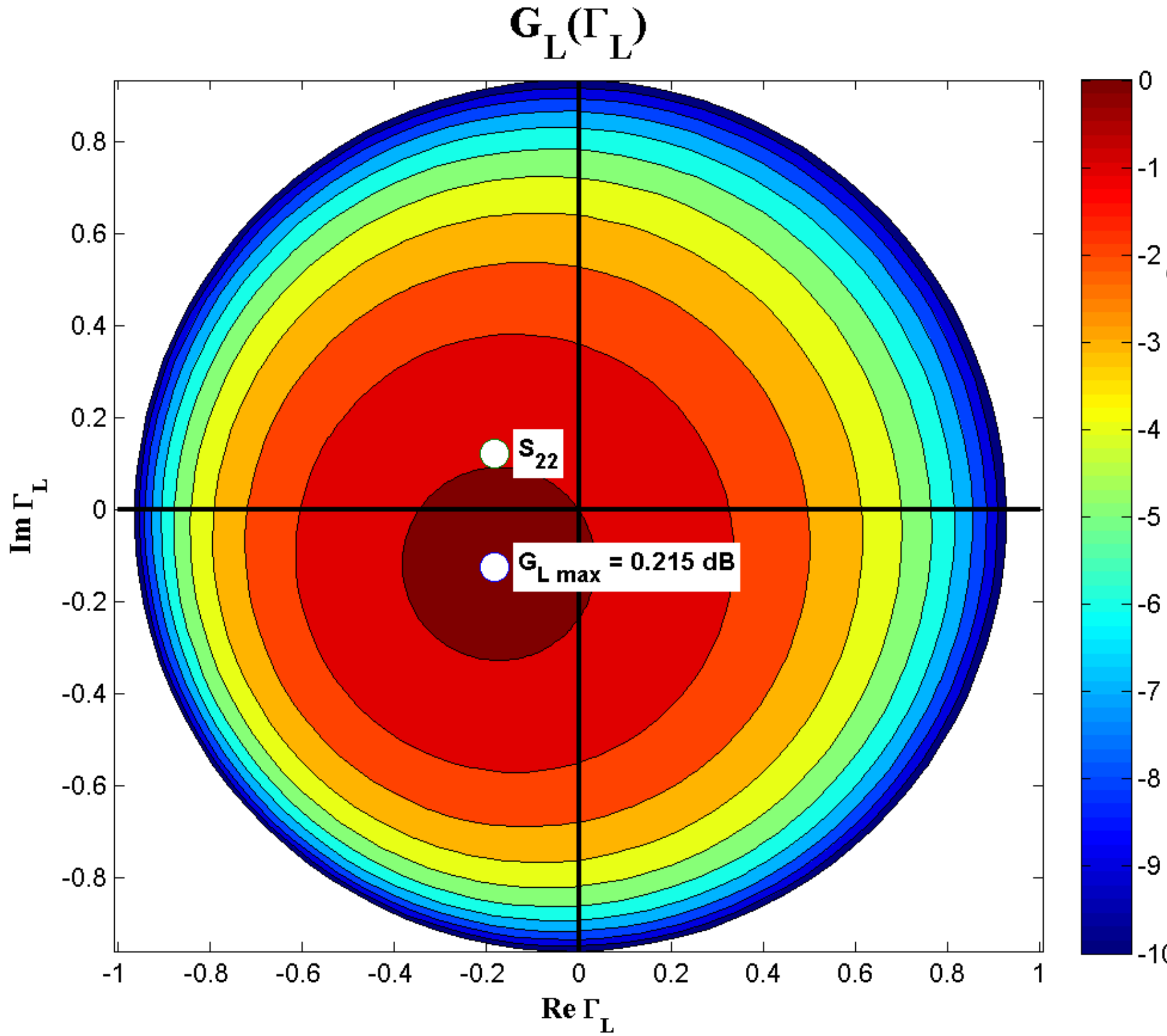
$G_L(\Gamma_L)$, constant value contours



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

$$G_{L\max} = G_L\Big|_{\Gamma_L = S_{22}^*}$$

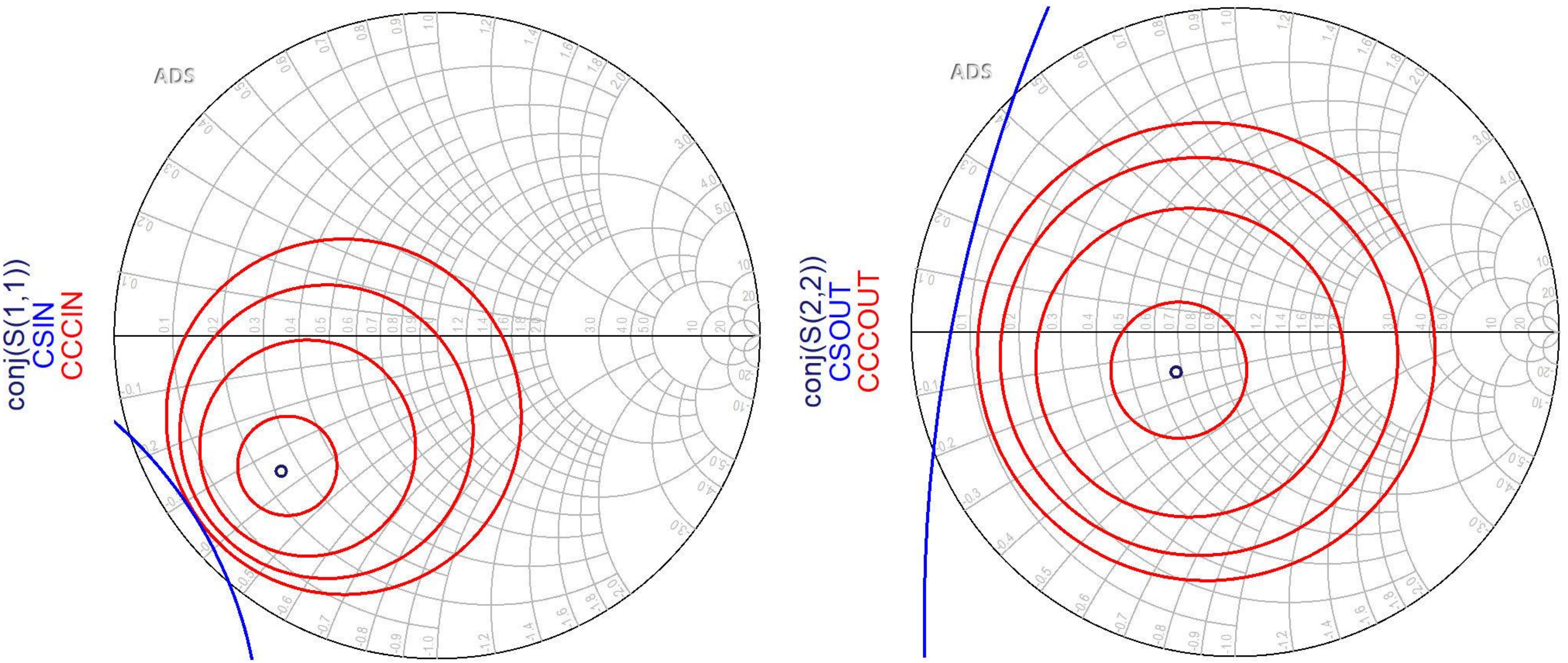
$G_L[dB](\Gamma_L)$, constant value contours



$$G_L[dB] = 10 \cdot \log\left(\frac{1-|\Gamma_L|^2}{|1-S_{22}\cdot\Gamma_L|^2}\right)$$

$$G_{L\max} = G_L\Big|_{\Gamma_L = S_{22}^*}$$

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

Design for Specified Gain

- We compute G_o, G_{Smax}, G_{Lmax}
- To obtain the design gain we **choose** supplemental gain needed (supplemental to constant G_o)
  - we account for the deviation that might arise from the unilateral assumption (using unilateral figure of merit U)

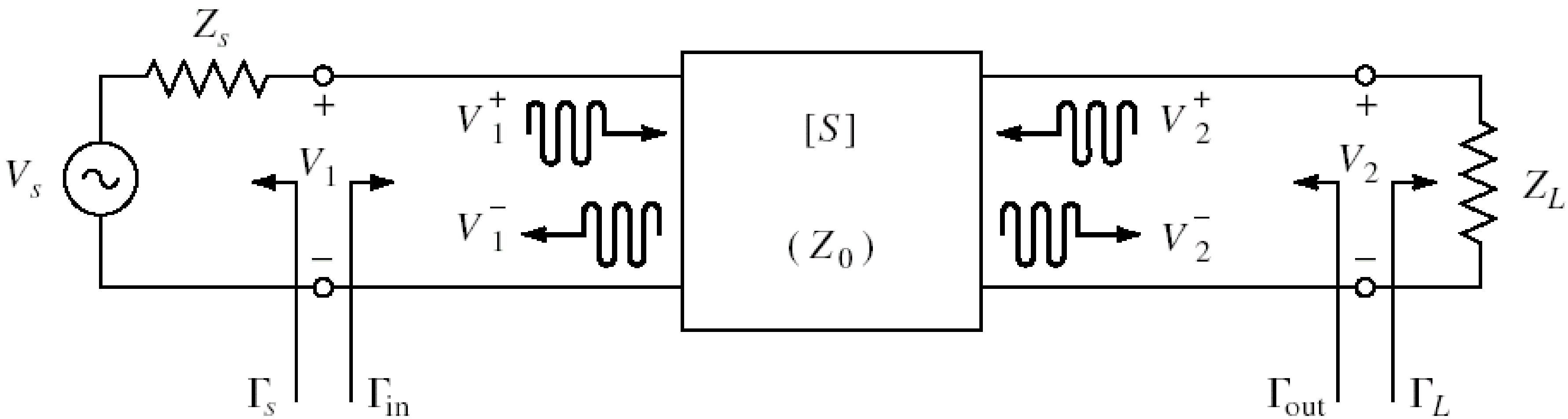
$$G_{design}[dB] = G_{S\_design}[dB] + G_0[dB] + G_{L\_design}[dB]$$

- We plot the circles for design (chosen) values G_{S_design}, G_{L_design}
- We design input and output matching circuits which move the reflection coefficient **on** or **inside** the design circles (depending on specific application requirements)

Low-Noise Amplifier Design

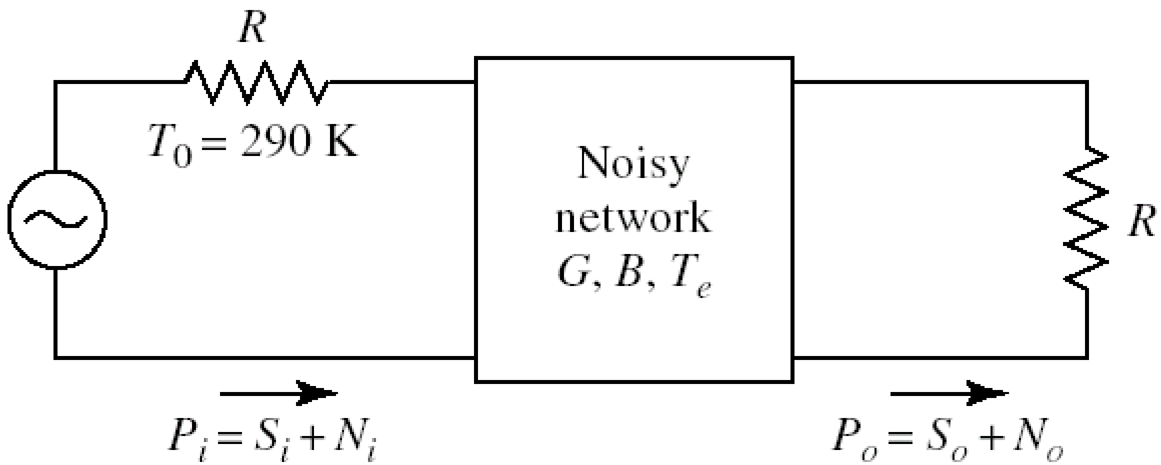
Microwave Amplifiers

Amplifier as two-port



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

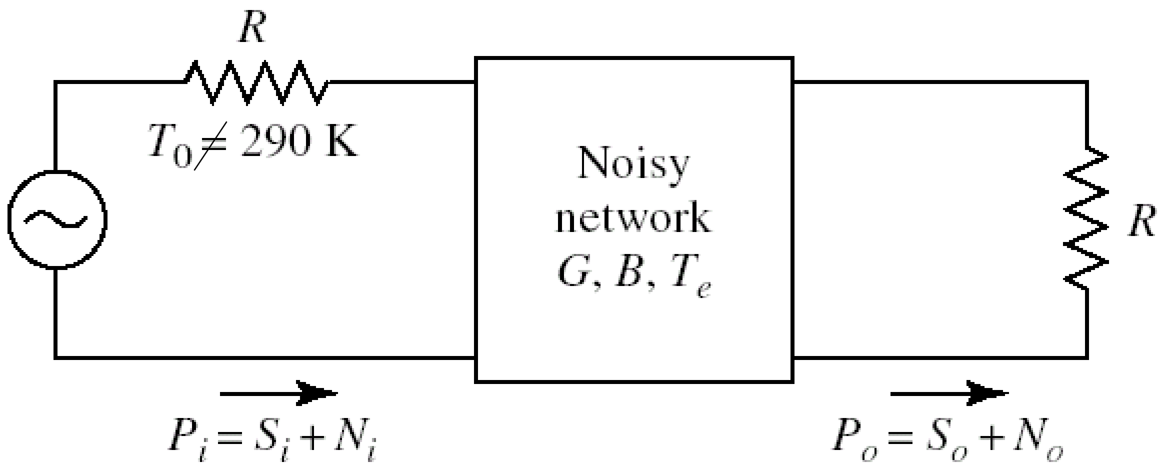
Noise Figure F



- 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)

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

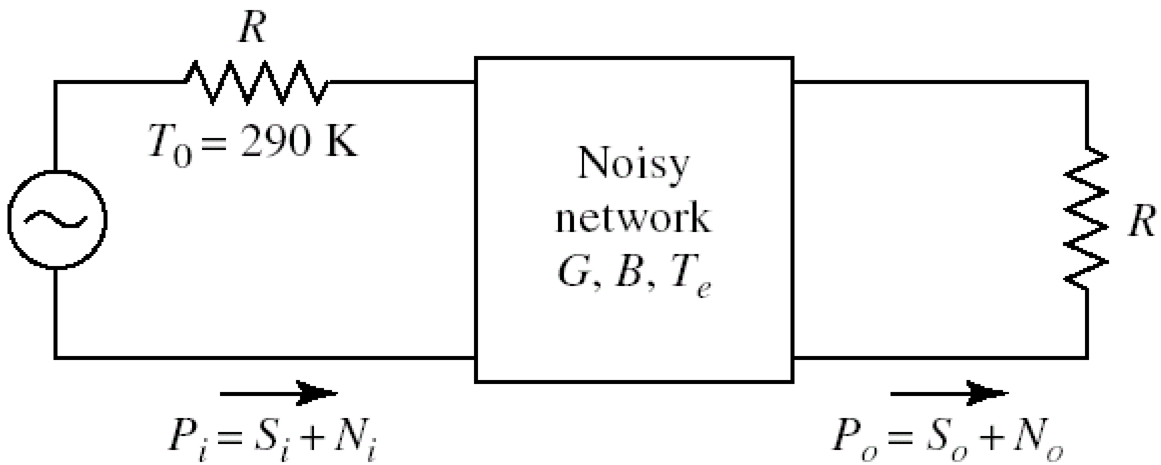
Noise Figure F



- 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

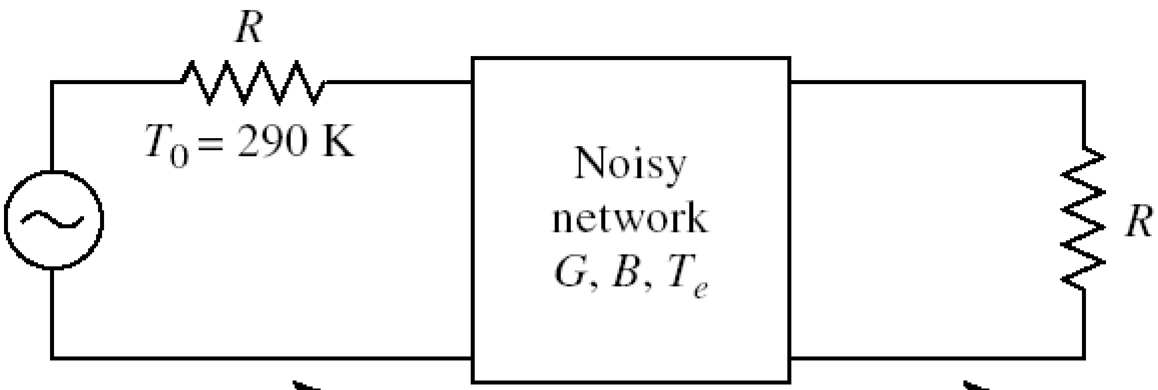
$$F \neq \left. \frac{S_i / N_i}{S_o / N_o} \right|_{T_0 \neq 290K}$$

Noise Figure F



- In general, the output noise power consists of two elements:
  - the input noise power amplified or attenuated by the device (for example amplified with the power gain G applied also to the desired signal)
  - a noise power generated internally by the network if the network is noisy (this power **does not** depend on the input noise power)

Noise Figure F



$$P_1 = S_1 + N_1 \qquad P_2 = S_2 + N_2$$

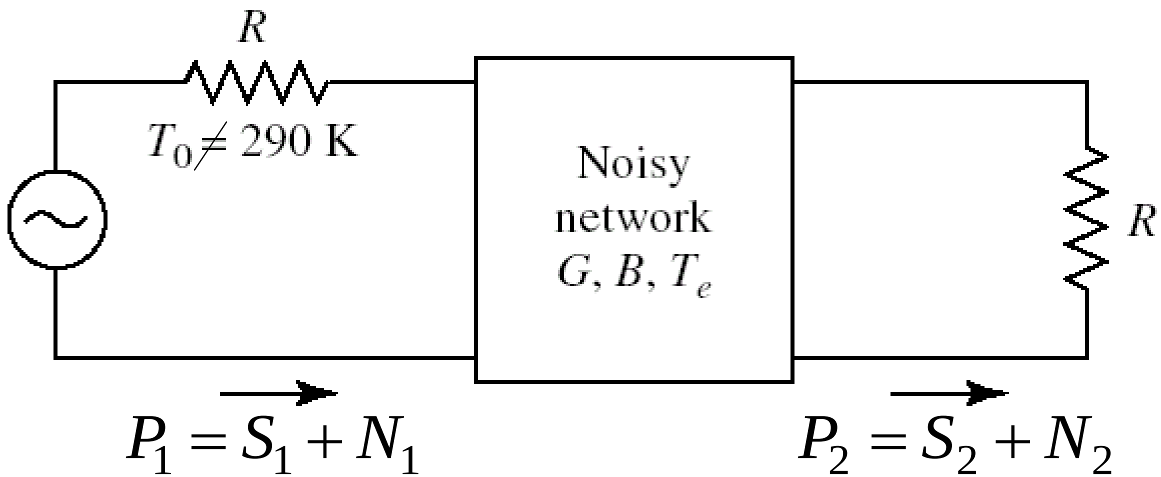
- Estimation of the internally generated noise power can be done using the Noise Figure F definition:

$$F = \left. \frac{S_1/N_1}{S_2/N_2} \right|_{T_0 = 290K,\, N_1 = N_0}$$

$$N_2 = F \cdot N_0 \cdot \frac{S_2}{S_1} = F \cdot N_0 \cdot G$$

$$N_2 = N_0 \cdot G + (F - 1) \cdot N_0 \cdot G$$

Noise Figure F

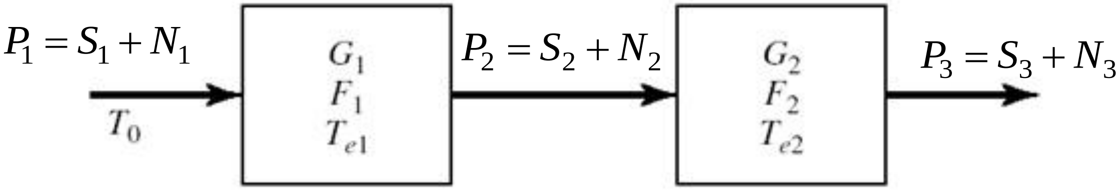


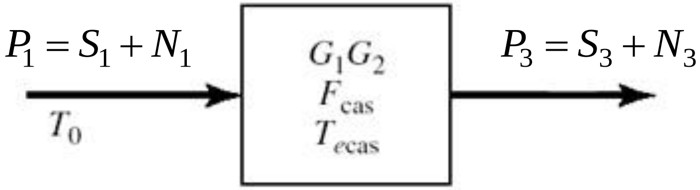
- We identify the two terms:
  - amplified input noise
  - internally generated noise
- When the input noise does not correspond to reference noise conditions (N1 ≠ No)
  - the internally generated noise does not change

$$N_2 = N_0 \cdot G + (F-1) \cdot N_0 \cdot G$$

$$N_2 = N_1 \cdot G + (F-1) \cdot N_0 \cdot G$$

Noise figure of a cascaded system





$$N_2 = N_1 \cdot G_1 + (F_1 - 1) \cdot N_0 \cdot G_1$$

$$N_3 = N_2 \cdot G_2 + (F_2 - 1) \cdot N_0 \cdot G_2$$

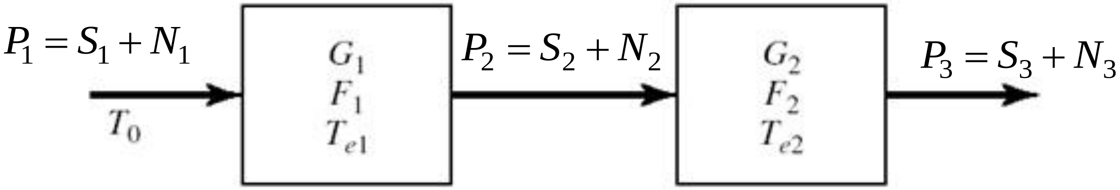
$$G_{cas} = G_1 \cdot G_2$$

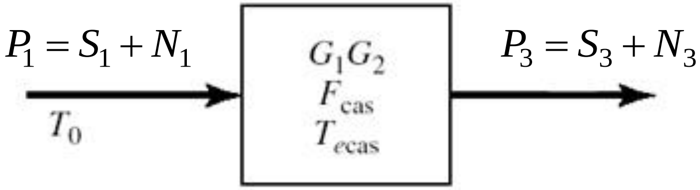
$$N_3 = N_1 \cdot G_{cas} + (F_{cas} - 1) \cdot N_0 \cdot G_{cas}$$

$$N_3 = [N_1 \cdot G_1 + (F_1 - 1) \cdot N_0 \cdot G_1] \cdot G_2 + (F_2 - 1) \cdot N_0 \cdot G_2$$

$$N_3 = N_1 \cdot G_1 \cdot G_2 + (F_1 - 1) \cdot N_0 \cdot G_1 \cdot G_2 + (F_2 - 1) \cdot N_0 \cdot G_2$$

Noise figure of a cascaded system





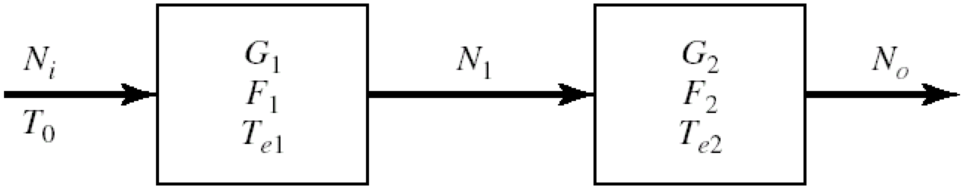
$$N_3 = N_1 \cdot G_1 \cdot G_2 + (F_1 - 1) \cdot N_0 \cdot G_1 \cdot G_2 + (F_2 - 1) \cdot N_0 \cdot G_2$$

$$G_{cas} = G_1 \cdot G_2 \qquad N_3 = N_1 \cdot G_{cas} + (F_{cas} - 1) \cdot N_0 \cdot G_{cas}$$

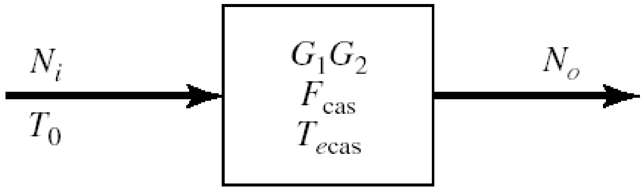
$$(F_1 - 1) \cdot N_0 \cdot G_1 \cdot G_2 + (F_2 - 1) \cdot N_0 \cdot G_2 = (F_{cas} - 1) \cdot N_0 \cdot G_1 \cdot G_2$$

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

Noise figure of a cascaded system



(a)



(b)

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

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

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

Friis Formula (noise)

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

- Friis Formula shows that:
  - the overall noise figure of a cascaded system is largely determined by the noise characteristics of the first stage
  - the noise introduced by the following stages is reduced:
    - -1
    - division by G (usually G > 1)

Friis Formula (noise)

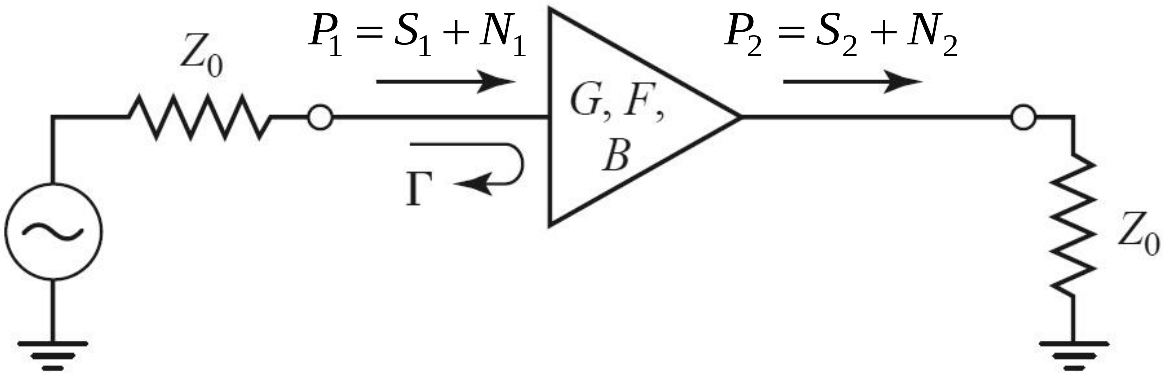
$$F_{cas} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 \cdot G_2} + \frac{F_4 - 1}{G_1 \cdot G_2 \cdot G_3} + \cdots$$

- Effects of Friis Formula:
- in multi stage amplifiers:
  - it’s essential that the first stage is as noiseless as possible even if that means sacrificing power gain
  - the following stages can be optimized for power gain
- in single stage amplifiers:
  - in the input matching circuit it’s important to have noiseless elements (pure reactance, lossless lines)
  - output matching circuit has less influence on the noise (noise generated at this level appears when the desired signal has already been amplified by the transistor)

$$V_{n(ef)} = \sqrt{4kTBR} \qquad P_n = kTB$$

Noise Figure of a Mismatched Amplifier

- An input mismatched amplifier($\Gamma \neq 0$)



$$G_{ech} = G \cdot \left(1 - |\Gamma|^2\right)$$

$$N_2 = N_1 \cdot G \cdot \left(1 - |\Gamma|^2\right) + (F - 1) \cdot N_0 \cdot G = N_1 \cdot G \cdot \left(1 - |\Gamma|^2\right) + \frac{F - 1}{1 - |\Gamma|^2} \cdot N_0 \cdot G \cdot \left(1 - |\Gamma|^2\right)$$

$$N_2 = N_1 \cdot G_{ech} + (F_{ech} - 1) \cdot N_0 \cdot G_{ech}$$

$$F_{ech} = 1 + \frac{F - 1}{1 - |\Gamma|^2} \geq F$$

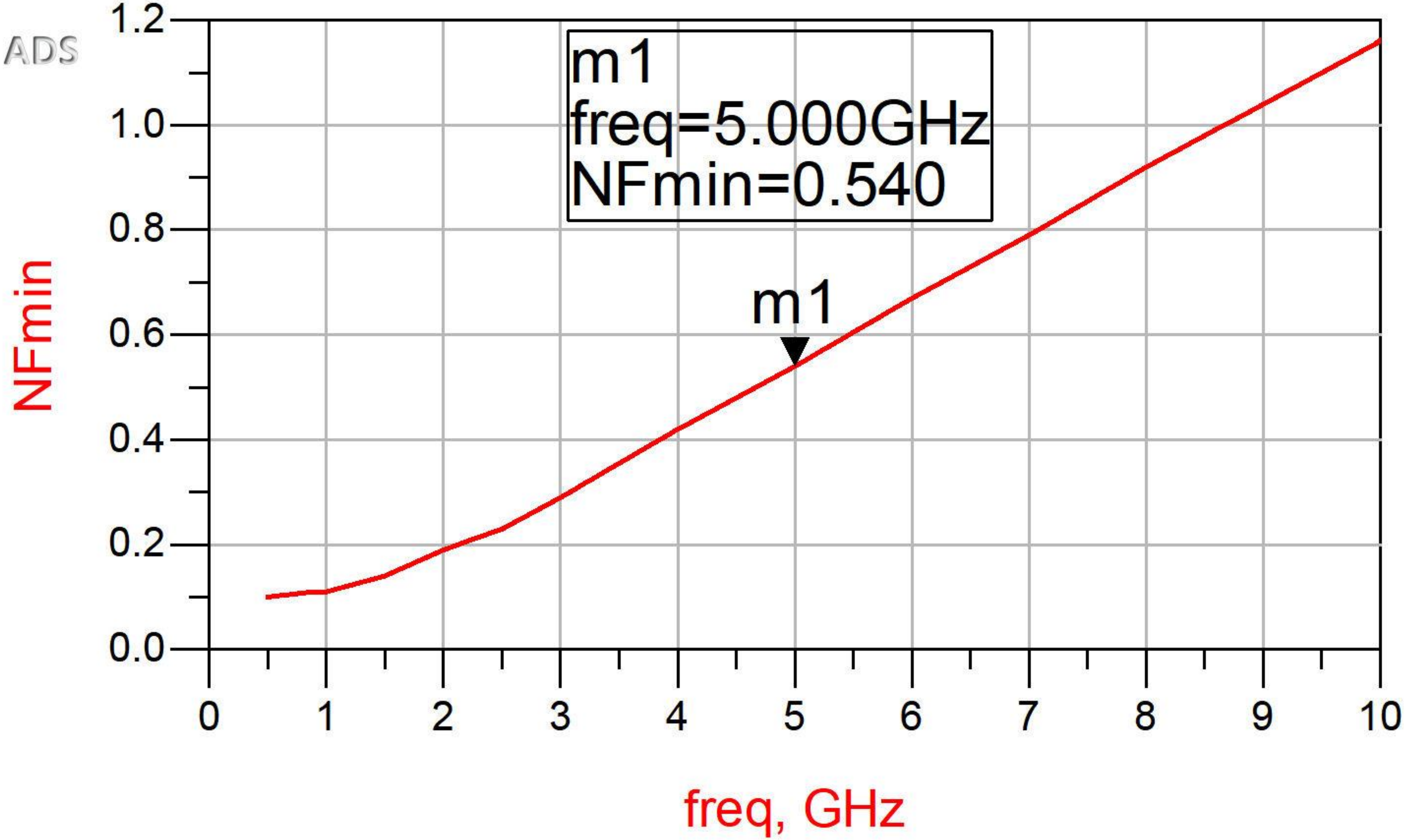
- Good noise figure **requires** good impedance matching

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 (**tipic [dB]**)
  - $\Gamma_{opt} = 0.45\angle 174°$
  - $r_n = 0.03$

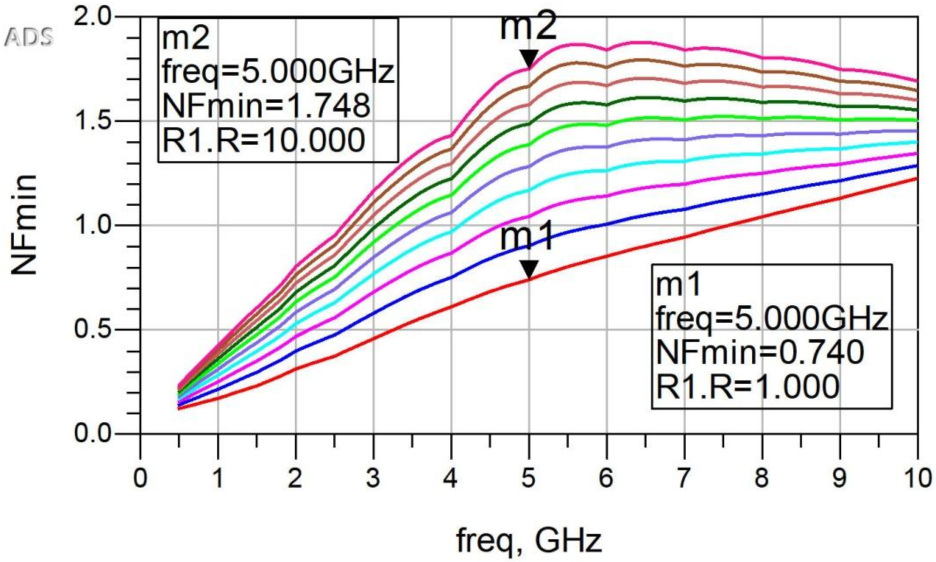
```
!ATF-34143
!S-PARAMETERS at Vds=3V  Id=20mA.   LAST UPDATED 01-29-99

# ghz s ma r 50

2.0  0.75  -126  6.306  90  0.088  23  0.26  -120
2.5  0.72  -145  5.438  75  0.095  15  0.25  -140
3.0  0.69  -162  4.762  62  0.102  7  0.23  -156
4.0  0.65  166  3.806  38  0.111  -8  0.22  174
5.0  0.64  139  3.165  16  0.119  -21  0.22  146
6.0  0.65  114  2.706  -5  0.125  -35  0.23  118
7.0  0.66  89  2.326  -27  0.129  -49  0.25  91
8.0  0.69  67  2.017  -47  0.133  -62  0.29  67
9.0  0.72  48  1.758  -66  0.135  -75  0.34  46

!FREQ   Fopt   GAMMA OPT      RN/Zo
!GHZ    dB    MAG    ANG      -

2.0  0.19  0.71  66  0.09
2.5  0.23  0.65  83  0.07
3.0  0.29  0.59  102  0.06
4.0  0.42  0.51  138  0.03
5.0  0.54  0.45  174  0.03
6.0  0.67  0.42  -151  0.05
7.0  0.79  0.42  -118  0.10
8.0  0.92  0.45  -88  0.18
9.0  1.04  0.51  -63  0.30
10.0  1.16  0.61  -43  0.46
```

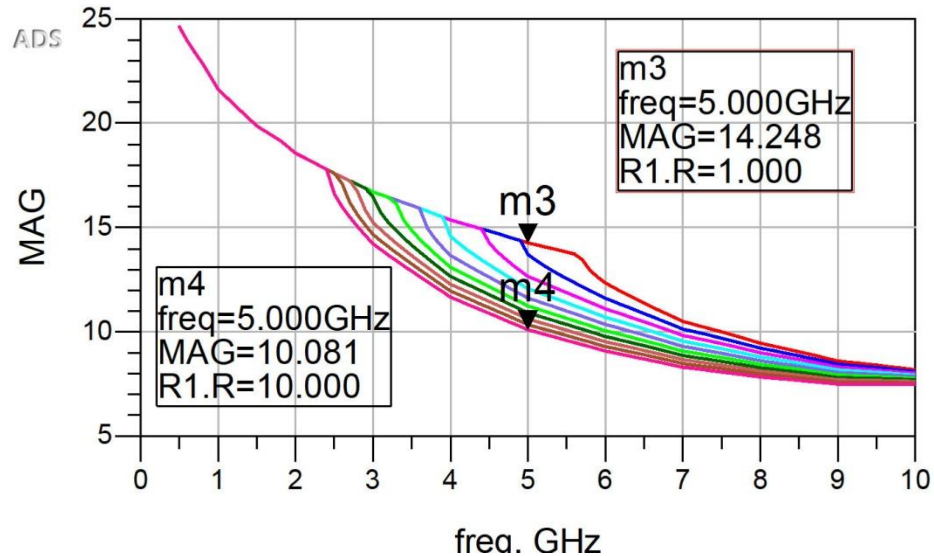
Example

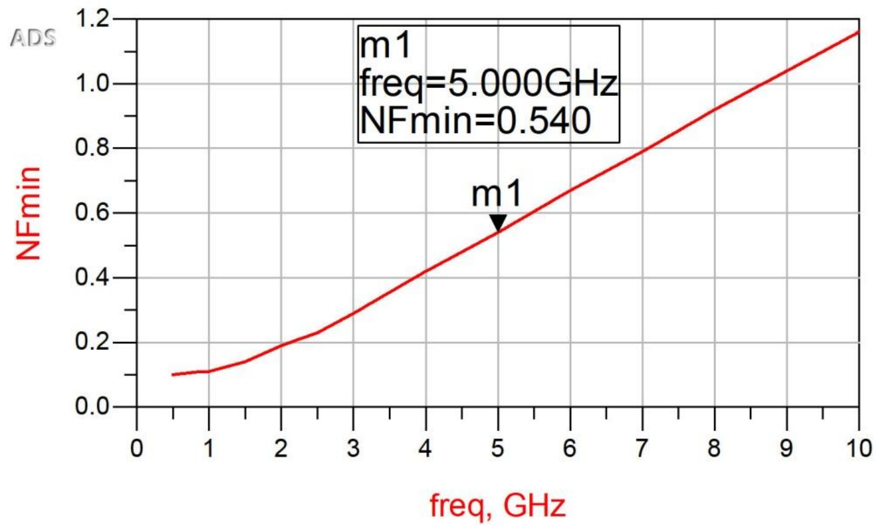


ADS
m1
freq=5.000GHz
NFmin=0.540
m1
NFmin
freq, GHz
1.2
1.0
0.8
0.6
0.4
0.2
0.0
0
1
2
3
4
5
6
7
8
9
10

Stabilization, input series resistor

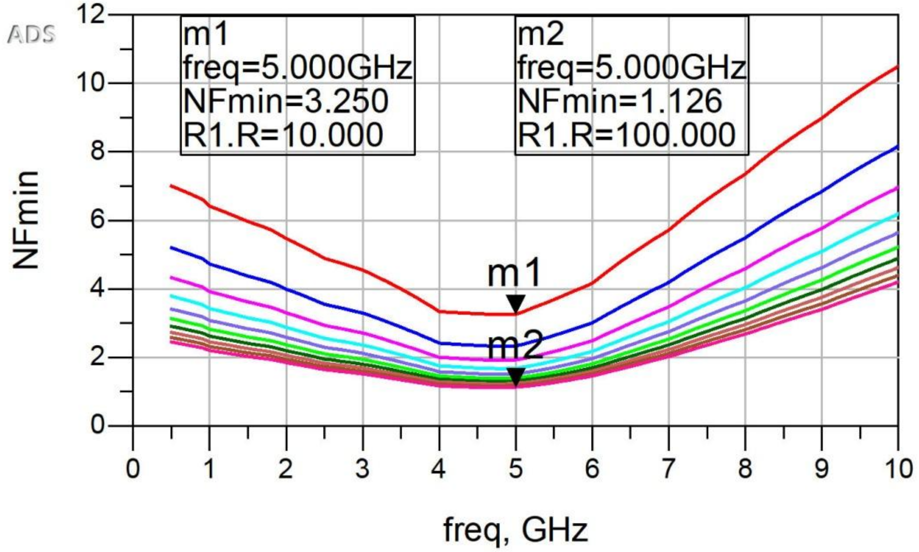


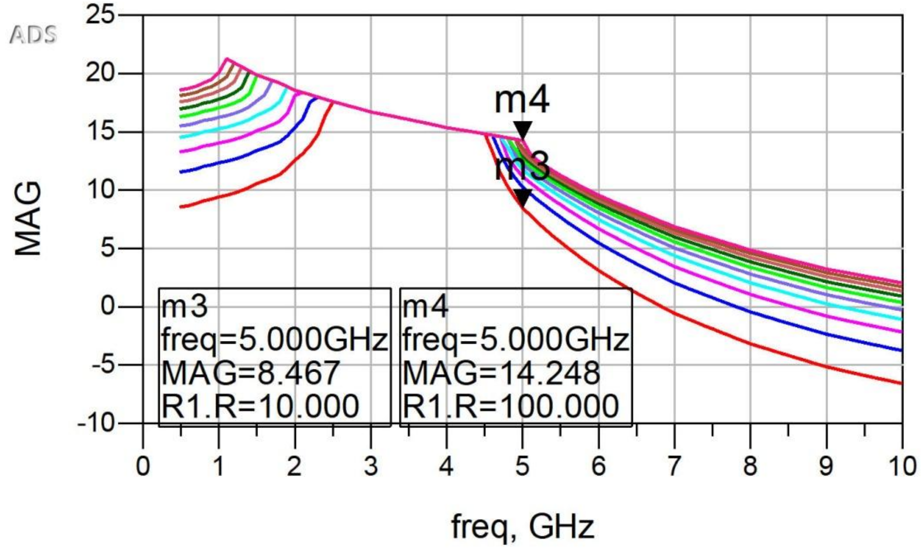




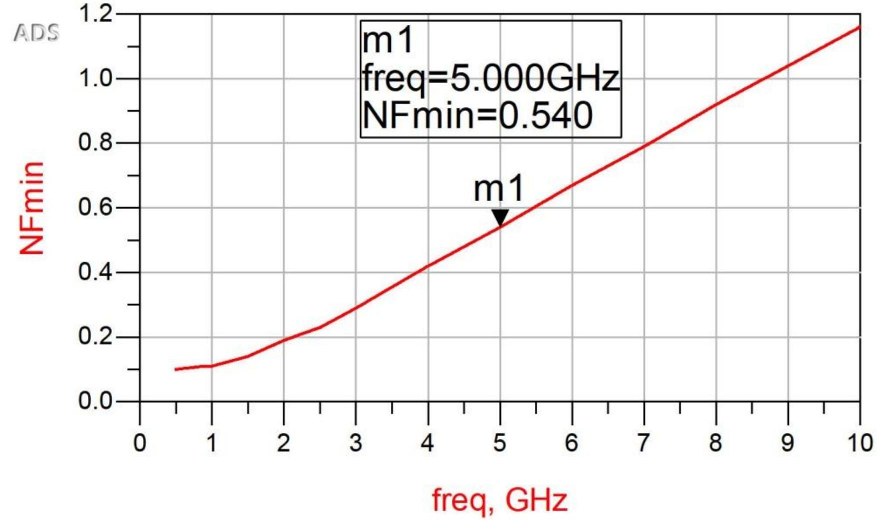
$$R_{SS} = 1 \div 10\,\Omega$$

Stabilization, input shunt resistor

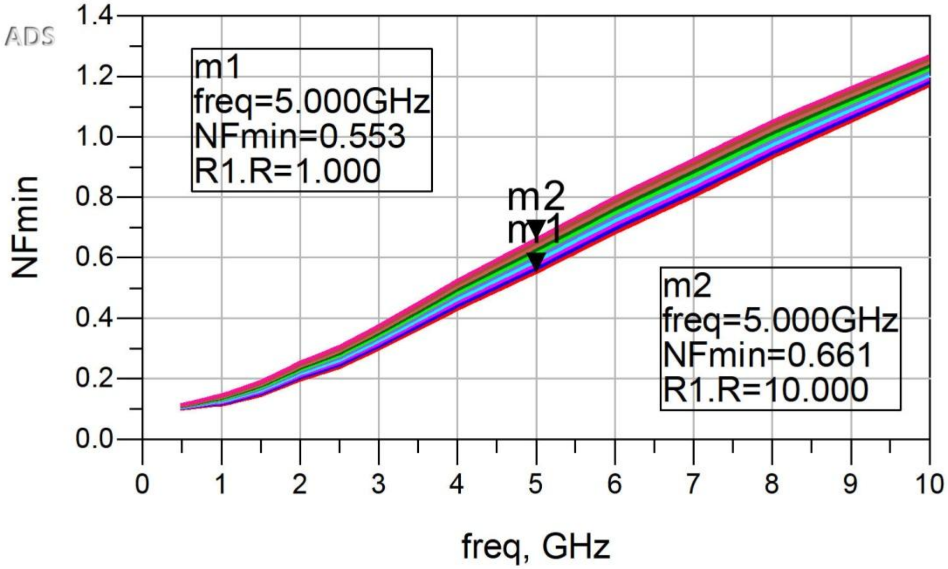


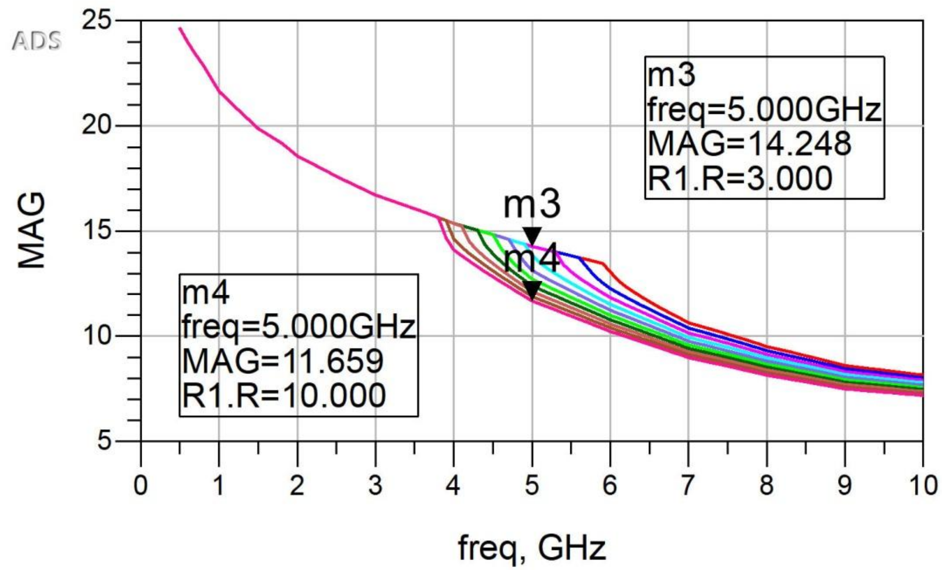


$$R_{PS} = 10 \div 100\,\Omega$$

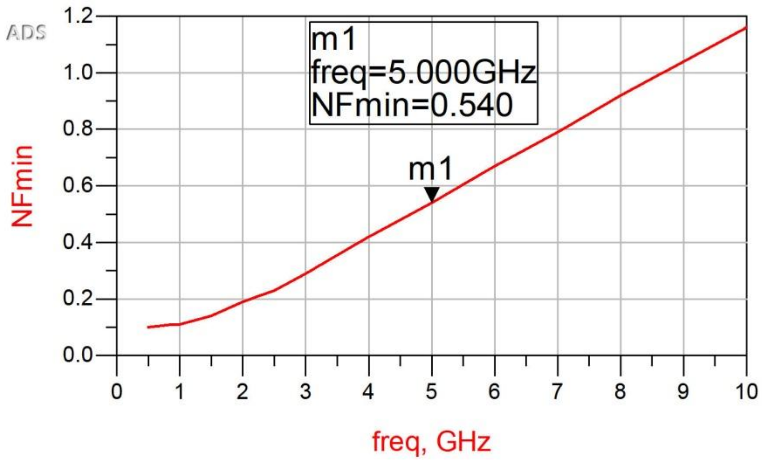


Stabilization, output series resistor

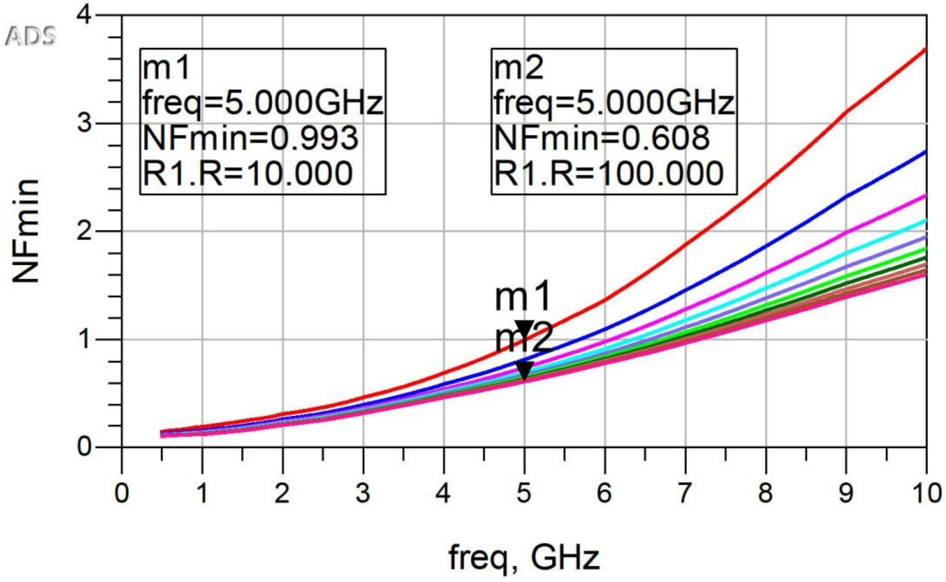


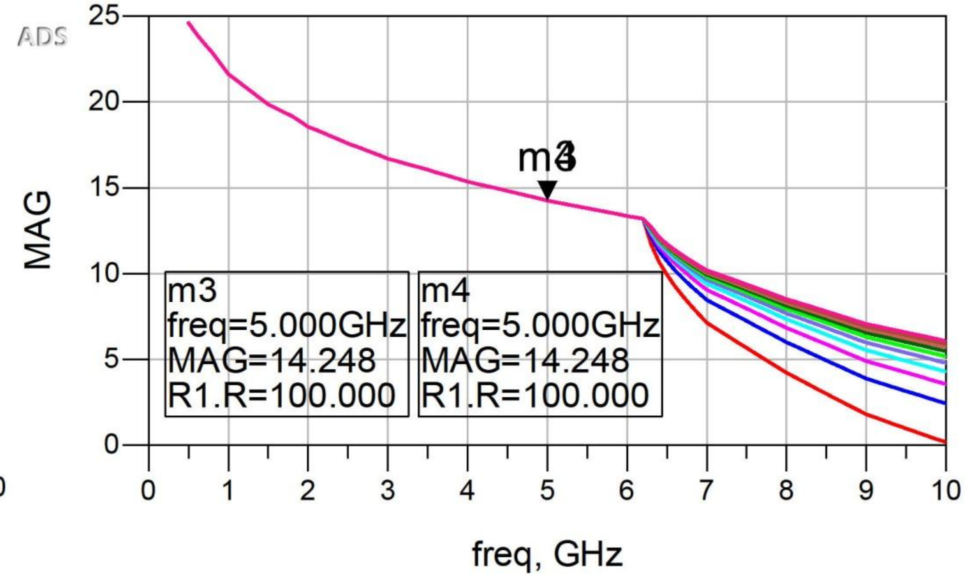


$$R_{SL} = 1 \div 10\,\Omega$$

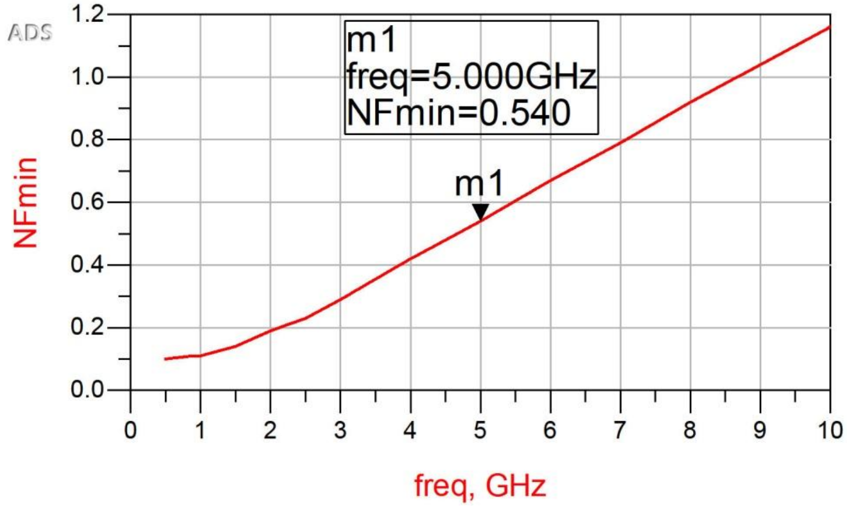


Stabilization, output shunt resistor





$$R_{PL} = 10 \div 100\,\Omega$$



Noise figure of a two-port amplifier

- 3 noise parameters (2reals + 1 complex):

$$F_{\min}, r_n = \frac{R_N}{Z_0}, \Gamma_{opt}$$

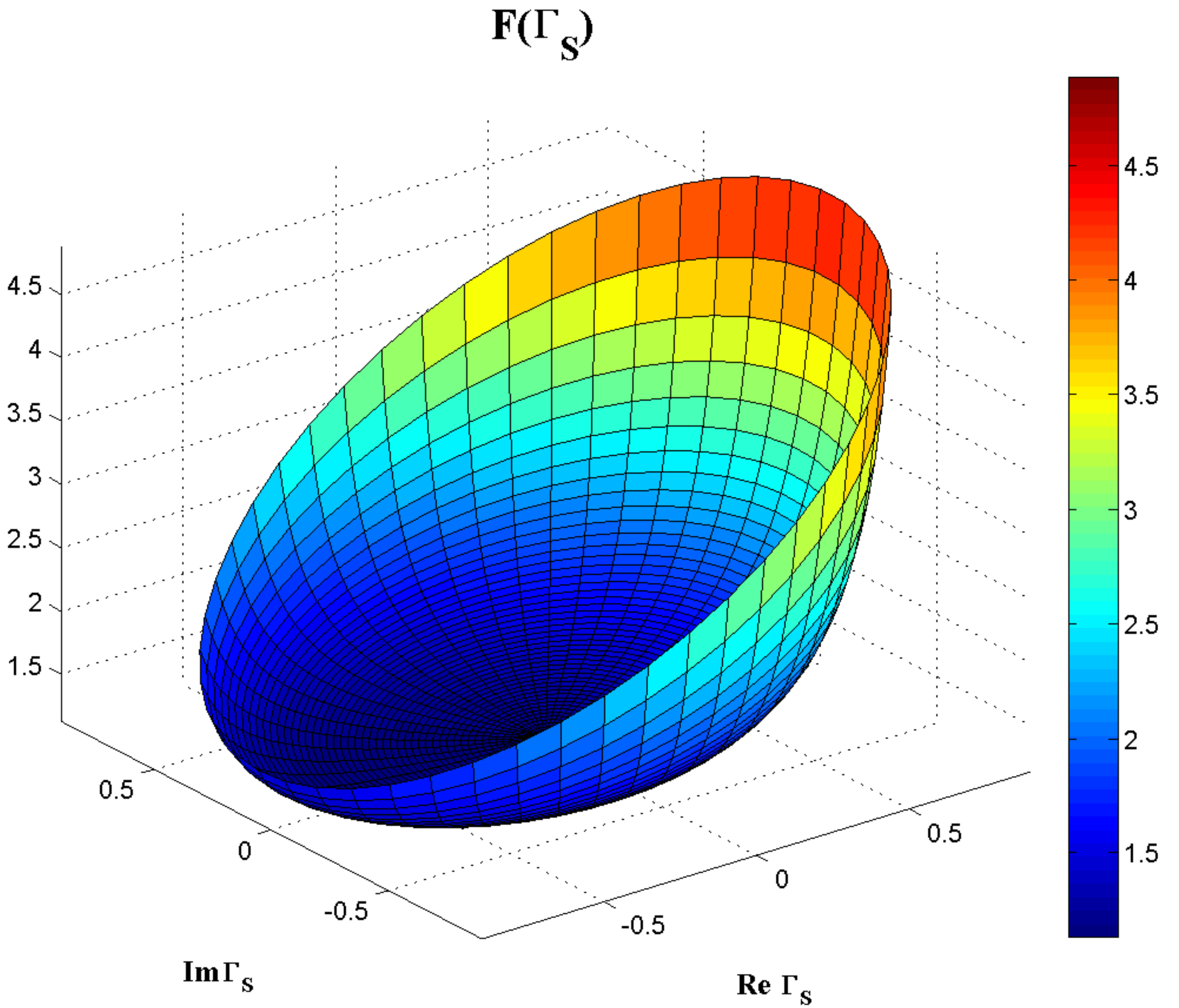
$$F = F_{\min} + \frac{R_N}{G_S} \cdot \left| Y_S - Y_{opt} \right|^2 \qquad Y_S = \frac{1}{Z_0} \cdot \frac{1 - \Gamma_S}{1 + \Gamma_S} \qquad Y_{opt} = \frac{1}{Z_0} \cdot \frac{1 - \Gamma_{opt}}{1 + \Gamma_{opt}}$$

$$F = F_{\min} + 4 \cdot r_n \cdot \frac{\left| \Gamma_S - \Gamma_{opt} \right|^2}{\left(1 - \left| \Gamma_S \right|^2\right) \cdot \left| 1 + \Gamma_{opt} \right|^2}$$

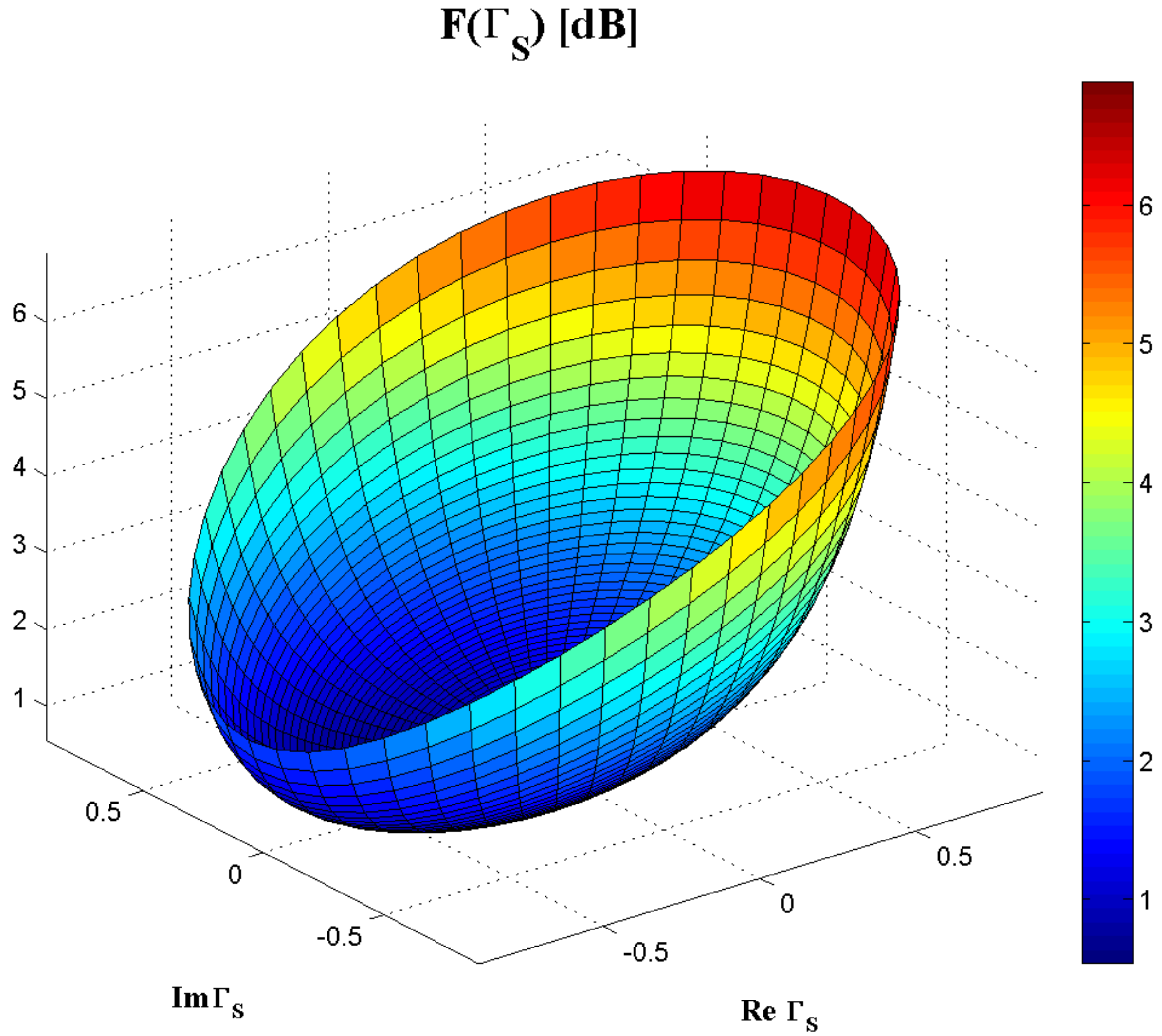
- Γ_{opt} optimum source reflection coefficient that results in minimum noise figure

$$\Gamma_S = \Gamma_{opt} \Rightarrow F = F_{\min}$$

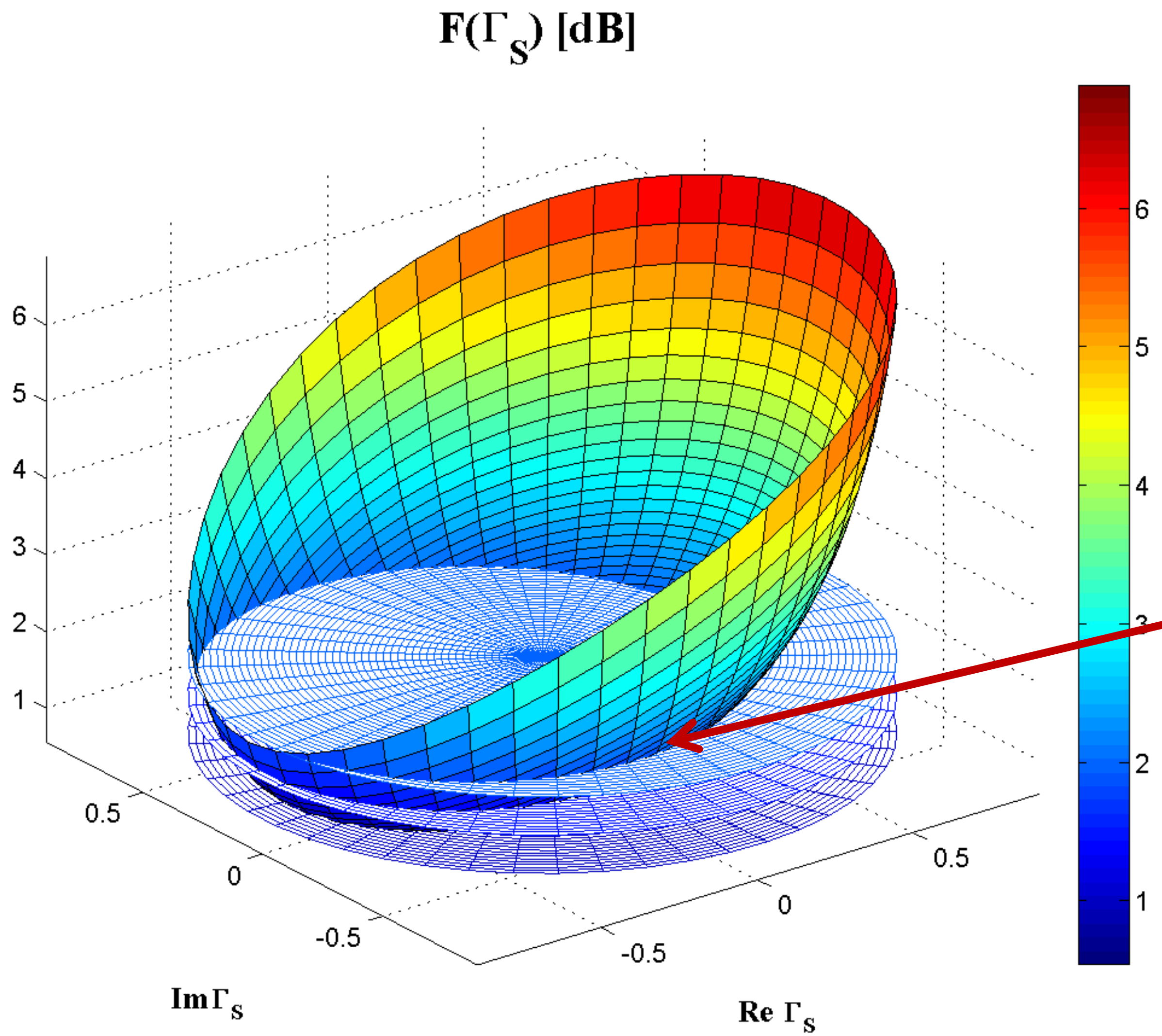
$F(\Gamma_S)$



$F[dB](\Gamma_S)$

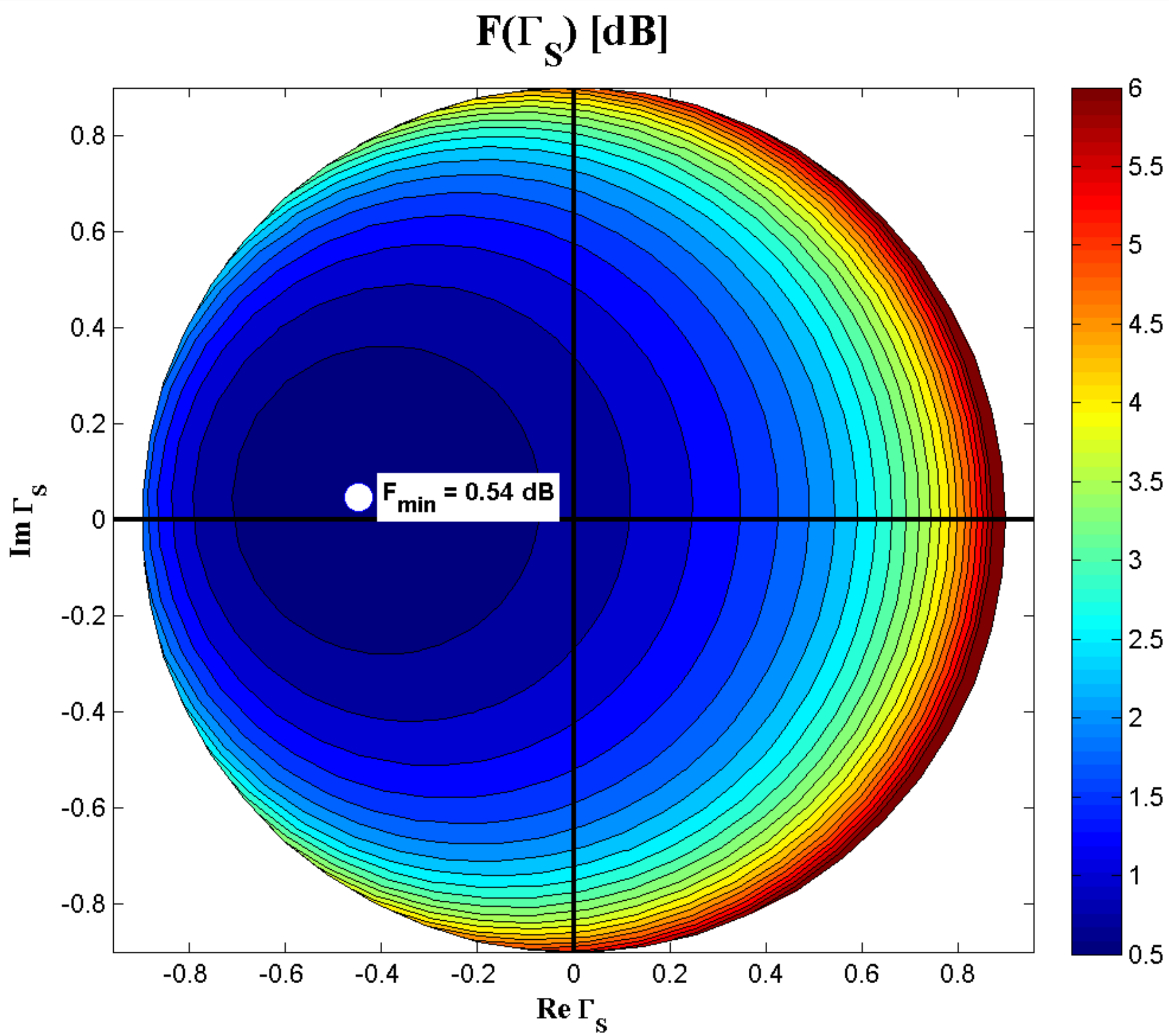


F[dB](Γ_S), constant value contours



Circles!!

G_S[dB](Γ_S), constant value contours



$\Gamma opt = 0.45 \angle 174°$

Circles of constant noise figure

- We define N (noise figure parameter)
  - N constant for F constant

$$N = \frac{\left|\Gamma_S - \Gamma_{opt}\right|^2}{1 - \left|\Gamma_S\right|^2} = \frac{F - F_{\min}}{4 \cdot r_n} \cdot \left|1 + \Gamma_{opt}\right|^2$$

$$\left(\Gamma_S - \Gamma_{opt}\right) \cdot \left(\Gamma_S^* - \Gamma_{opt}^*\right) = N \cdot \left(1 - \left|\Gamma_S\right|^2\right)$$

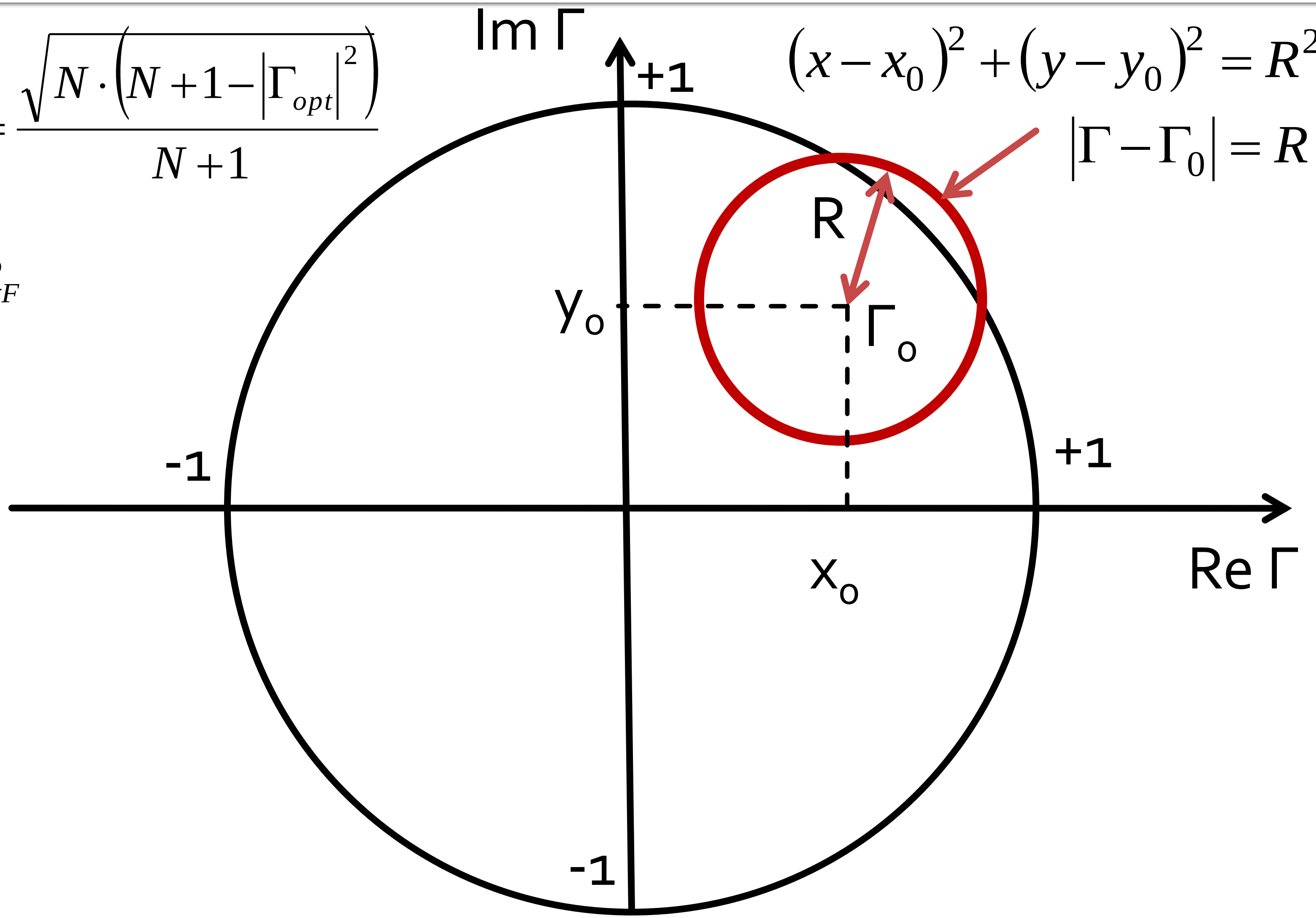
$$\Gamma_S \cdot \Gamma_S^* + N \cdot \left|\Gamma_S\right|^2 - \left(\Gamma_S \cdot \Gamma_{opt}^* - \Gamma_S^* \cdot \Gamma_{opt}\right) + \Gamma_{opt} \cdot \Gamma_{opt}^* = N$$

$$\Gamma_S \cdot \Gamma_S^* - \frac{\Gamma_S \cdot \Gamma_{opt}^* - \Gamma_S^* \cdot \Gamma_{opt}}{N+1} + \Gamma_{opt} \cdot \Gamma_{opt}^* = \frac{N - \left|\Gamma_{opt}\right|^2}{N+1} \bigg/ + \frac{\left|\Gamma_{opt}\right|^2}{(N+1)^2}$$

Circles of constant noise figure

$$\left|\Gamma_S - \frac{\Gamma_{opt}}{N+1}\right| = \frac{\sqrt{N \cdot \left(N + 1 - \left|\Gamma_{opt}\right|^2\right)}}{N+1}$$

$$\left|\Gamma_S - C_F\right| = R_F$$



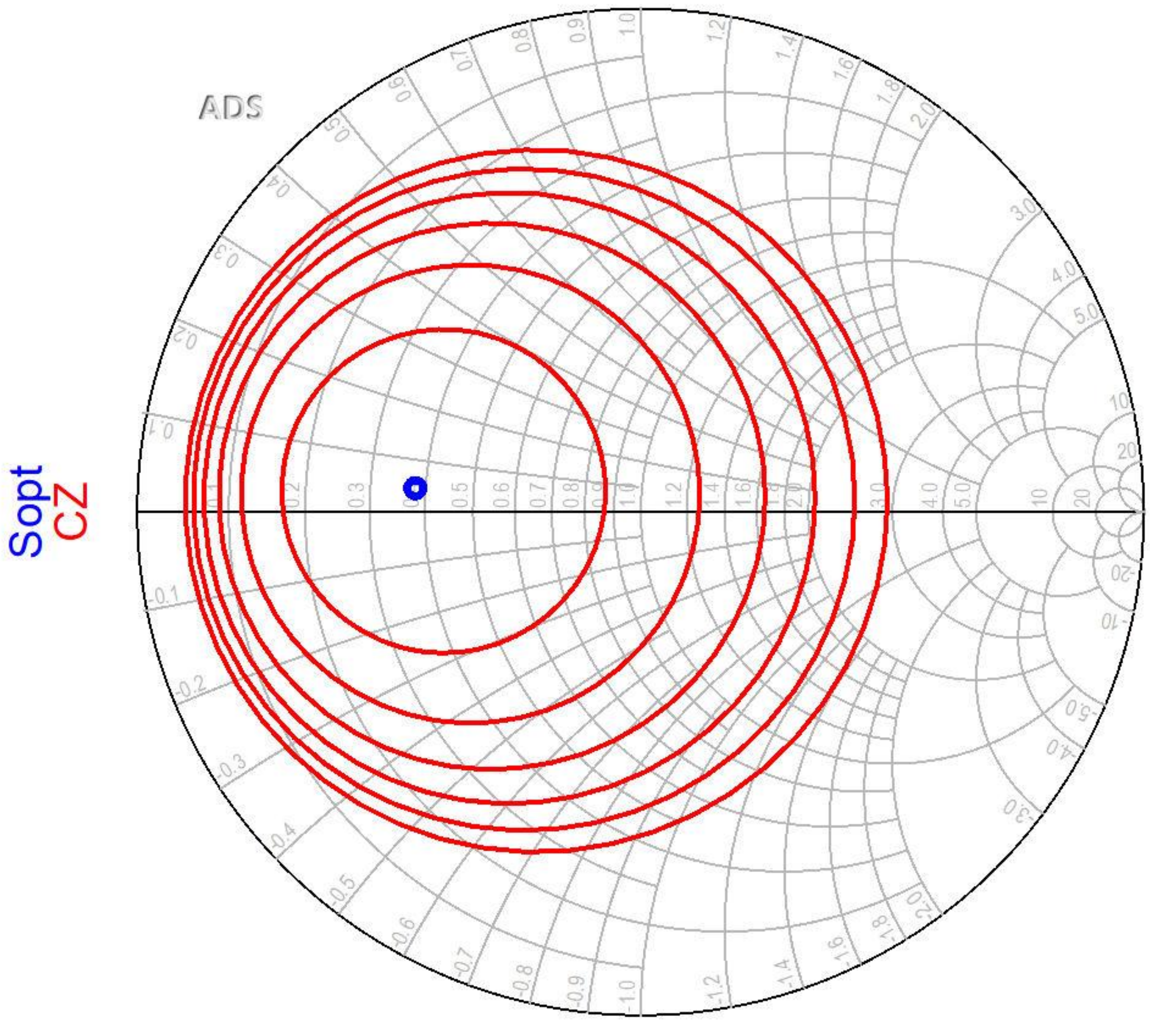
Circles of constant noise figure

$$\left|\Gamma_S - \frac{\Gamma_{opt}}{N+1}\right| = \frac{\sqrt{N \cdot \left(N+1-\left|\Gamma_{opt}\right|^2\right)}}{N+1} \qquad \left|\Gamma_S - C_F\right| = R_F$$

$$C_F = \frac{\Gamma_{opt}}{N+1} \qquad R_F = \frac{\sqrt{N \cdot \left(N+1-\left|\Gamma_{opt}\right|^2\right)}}{N+1}$$

- The locus in the complex plane Γ_S of the points with constant noise figure is a circle
- **Interpretation**: Any reflection coefficient Γ_S which plotted in the complex plane lies **on** the circle drawn for F_{circle} will lead to a noise factor $F = F_{circle}$
  - Any reflection coefficient Γ_S plotted **outside** this circle will lead to a noise factor $F > F_{circle}$
  - Any reflection coefficient Γ_S plotted **inside** this circle will lead to a noise factor $F < F_{circle}$

ADS

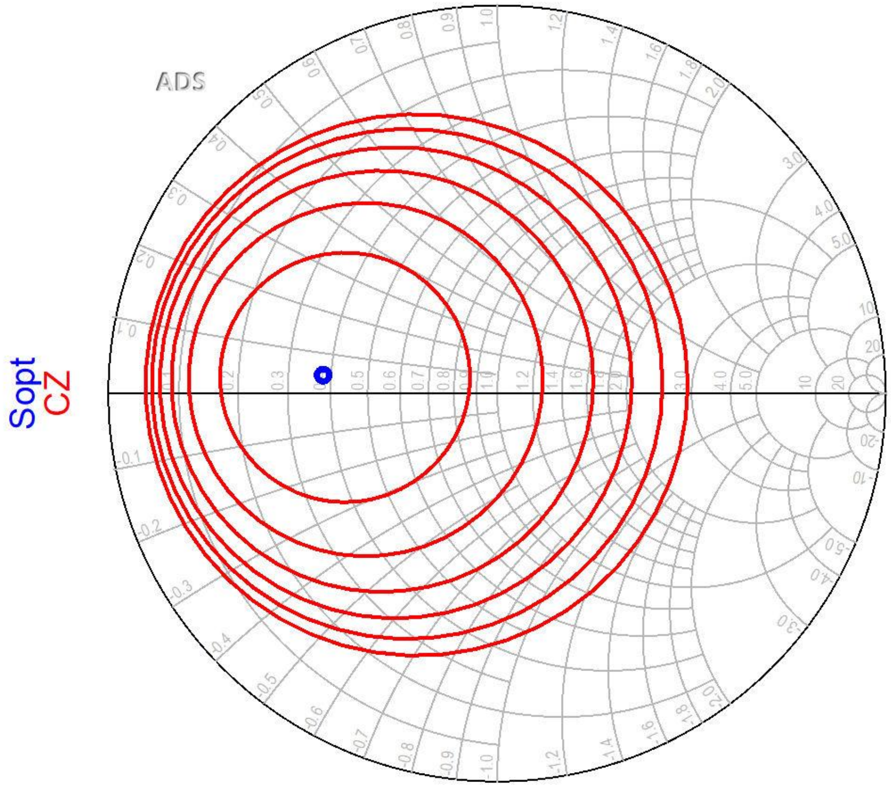


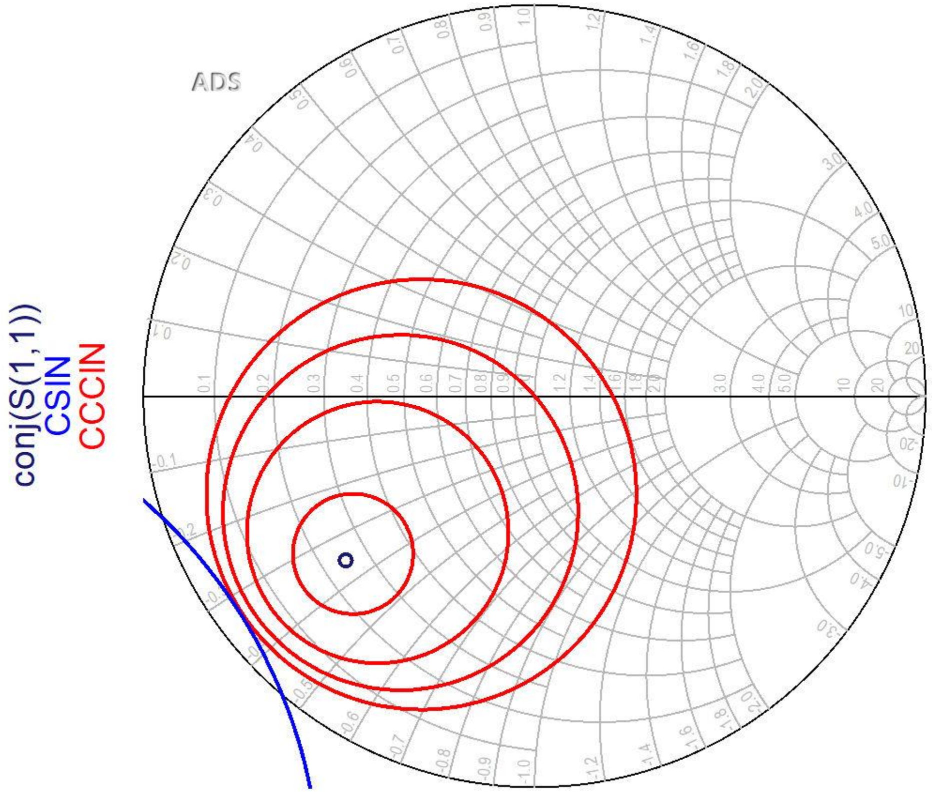
Circles of constant noise figure

- The noise internally generated by the transistor depends **only** by the input matching circuit
- A minimum noise figure is possible (NF_{min} – a datasheet parameter for the transistor)
- If we design a low noise amplifier (**LNA**) the usual design technique is as follows:
  - design of the input matching circuit solely (largely) for noise optimization
  - design of output matching circuit for gain compensation/optimization (if lossy circuits are used the output matching circuit noise can be added but the transistor noise is not influenced)

LNA – Low Noise Amplifier

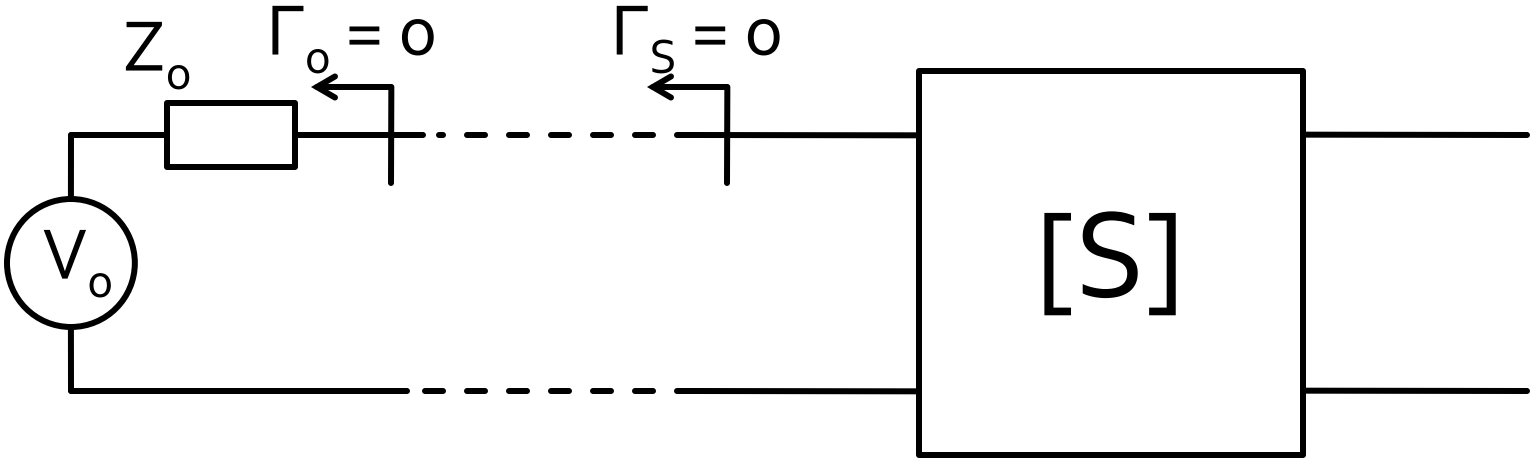
- Usually a transistor suitable for implementing an LNA at a certain frequency will have input gain circles and noise circles in the same area for Γ_S





Matching – 1

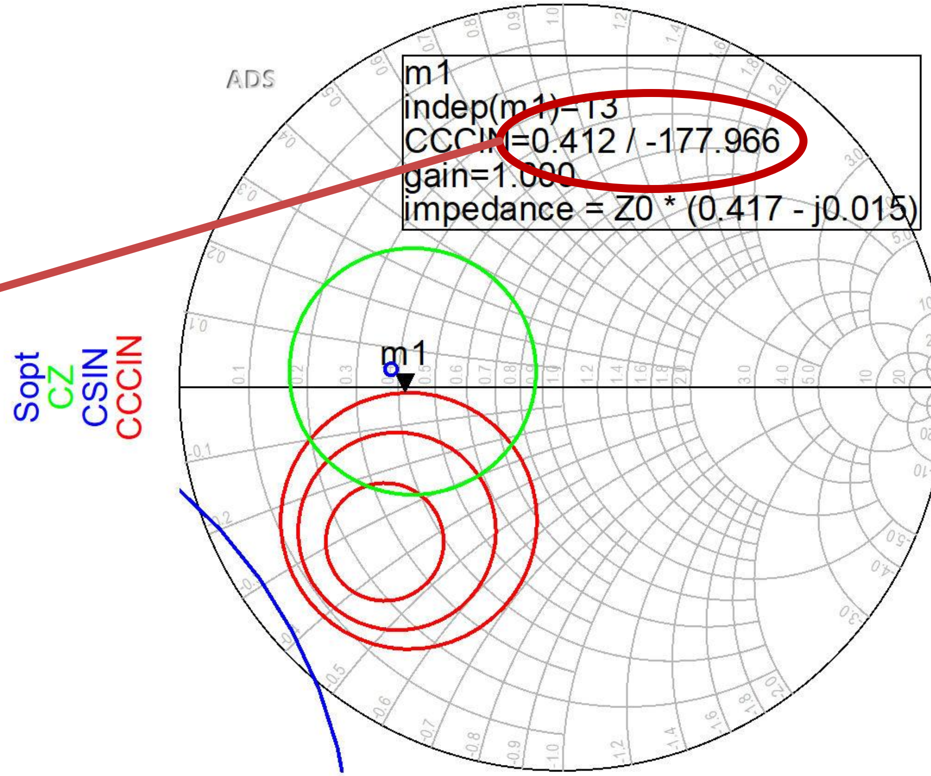
- Connecting the amplifier (transistor) directly to the source with Zo generate a reflection coefficient seen towards the source equal with **0** (complex, $\Gamma_o = 0 + 0 \cdot j$)
  - most of the time this reflection coefficient does not offer optimum noise/gain



Matching – 2

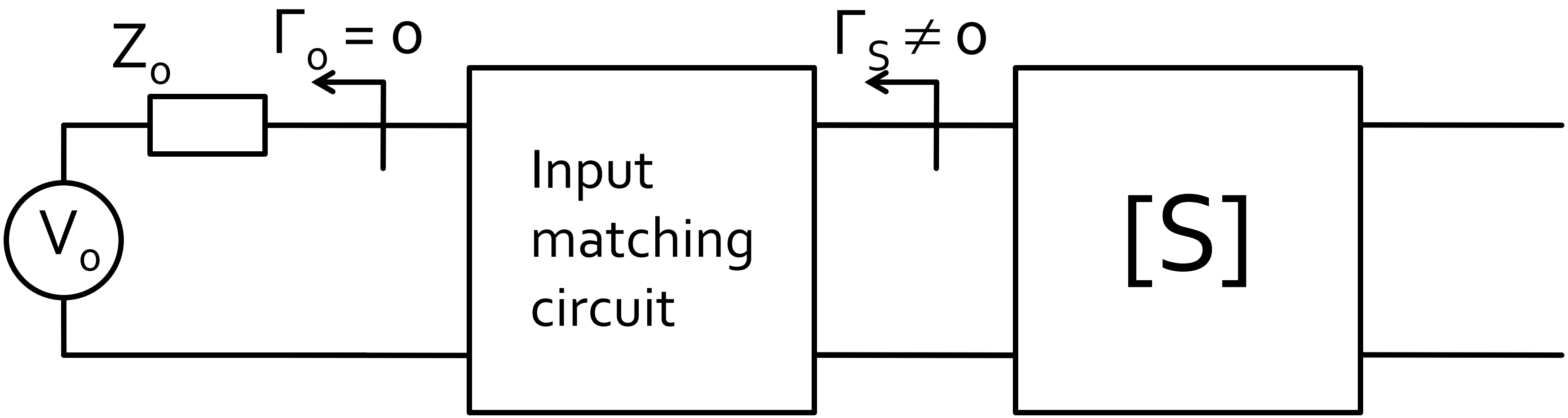
- We plot on the complex plane (Smith Chart) the stability/gain/noise circles (depending on the particular application)
- We choose a point with a suitable position relative to these circles (also application dependent)
- We determine the input reflection coefficient corresponding to this point, Γ_S

$$\Gamma_S = 0.412\angle -177.966°$$



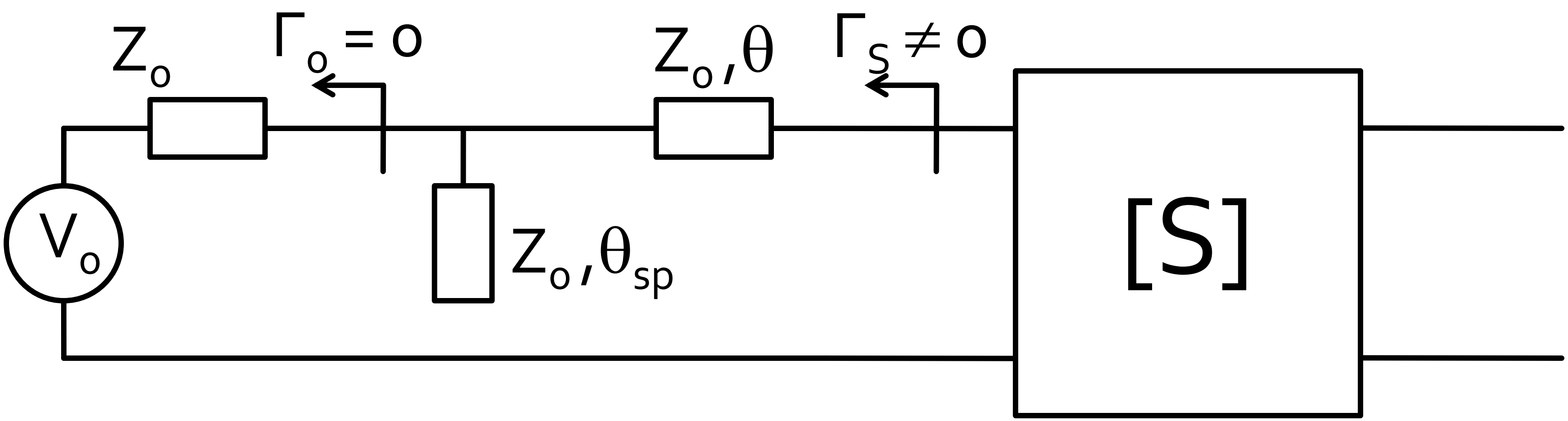
Matching – 3

- We insert the input matching circuits which allows the transistor to see towards the source the previously determined reflection coefficient Γ_S



Matching – 4

- Easiest to design matching section consists in the insertion of (in order from the transistor towards the Z_o source):
  - a series Z_o line, with electrical length **θ**
  - a shunt stub, open-circuited, made from a Z_o line, with electrical length **θ_{sp}**

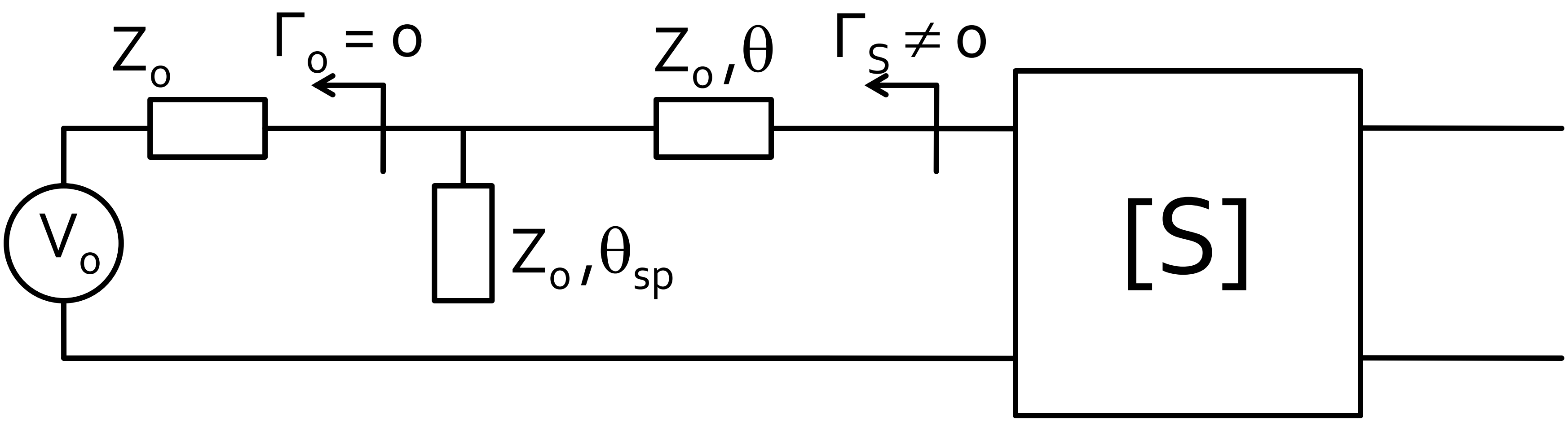


Matching – 5

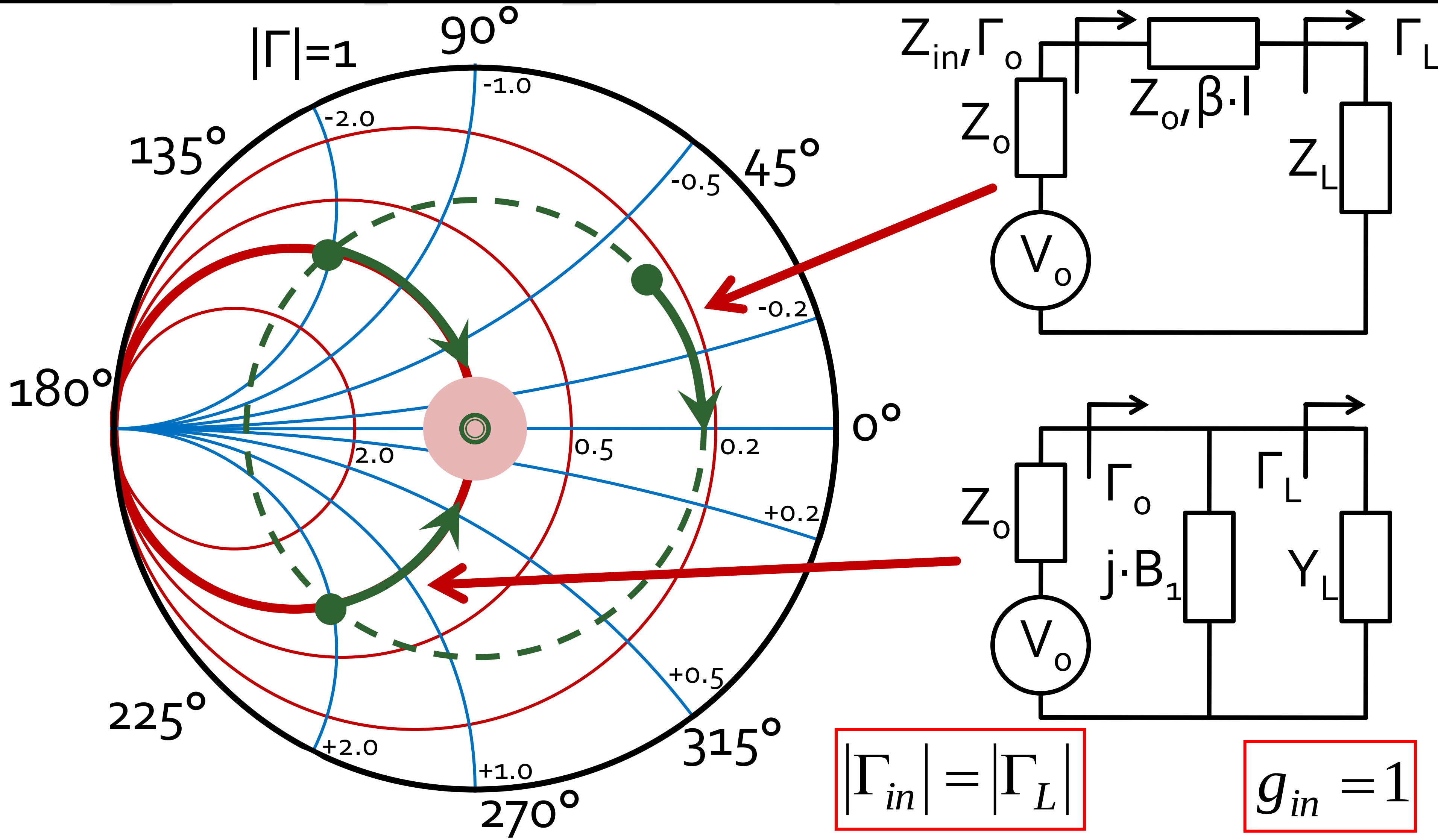
- Computation depends solely on Γ_S (magnitude and phase)

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

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



Shunt stub matching, C6-7



$$|\Gamma_{in}| = |\Gamma_L|$$

$$g_{in} = 1$$

Example, LNA @ 5 GHz

- 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 (**tipic [dB]**)
  - $\Gamma_{opt} = 0.45\angle 174°$
  - $r_n = 0.03$

!ATF-34143
!S-PARAMETERS at Vds=3V Id=20mA. LAST UPDATED 01-29-99

ghz s ma r 50

2.0	0.75	-126	6.306	90	0.088	23	0.26	-120
2.5	0.72	-145	5.438	75	0.095	15	0.25	-140
3.0	0.69	-162	4.762	62	0.102	7	0.23	-156
4.0	0.65	166	3.806	38	0.111	-8	0.22	174
5.0	0.64	139	3.165	16	0.119	-21	0.22	146
6.0	0.65	114	2.706	-5	0.125	-35	0.23	118
7.0	0.66	89	2.326	-27	0.129	-49	0.25	91
8.0	0.69	67	2.017	-47	0.133	-62	0.29	67
9.0	0.72	48	1.758	-66	0.135	-75	0.34	46

!FREQ	Fopt	GAMMA OPT		RN/Zo
!GHZ	dB	MAG	ANG	-
2.0	0.19	0.71	66	0.09
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4.0	0.42	0.51	138	0.03
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7.0	0.79	0.42	-118	0.10
8.0	0.92	0.45	-88	0.18
9.0	1.04	0.51	-63	0.30
10.0	1.16	0.61	-43	0.46

Example, LNA @ 5 GHz

- Low Noise Amplifier
- At the input matching a compromise is required between:
  - noise (~~input~~ constant noise figure circles)
  - gain (input constant gain circles)
  - stability (input stability circle)
- At the output matching noise **is not influenced**. A compromise is required between :
  - gain (output constant gain circles)
  - stability (output stability circle)

Exemplu, LNA @ 5 GHz

$$U = \frac{|S_{12}|\cdot|S_{21}|\cdot|S_{11}|\cdot|S_{22}|}{\left(1-|S_{11}|^2\right)\cdot\left(1-|S_{22}|^2\right)} = 0.094$$

$$-0.783\,dB < G_T[dB] - G_{TU}[dB] < 0.861\,dB$$

$$G_{TU\,max} = \frac{1}{1-|S_{11}|^2}\cdot|S_{21}|^2\cdot\frac{1}{1-|S_{22}|^2} = 17.83$$

$$G_{TU\,max}[dB] = 12.511\,dB$$

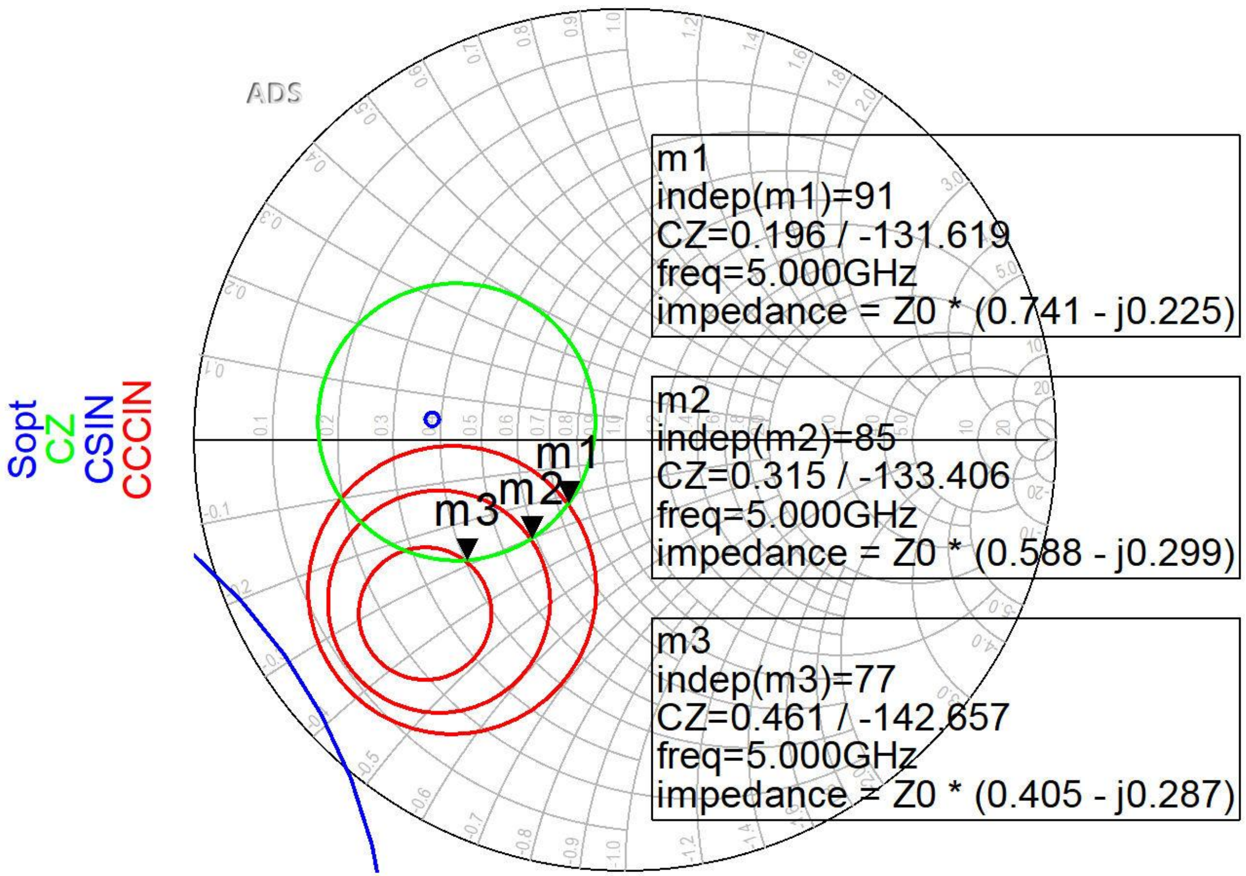
$$G_0 = |S_{21}|^2 = 10.017 = 10.007\,dB$$

$$G_{S\,max} = \frac{1}{1-|S_{11}|^2} = 1.694 = 2.289\,dB$$

$$G_{L\,max} = \frac{1}{1-|S_{22}|^2} = 1.051 = 0.215\,dB$$

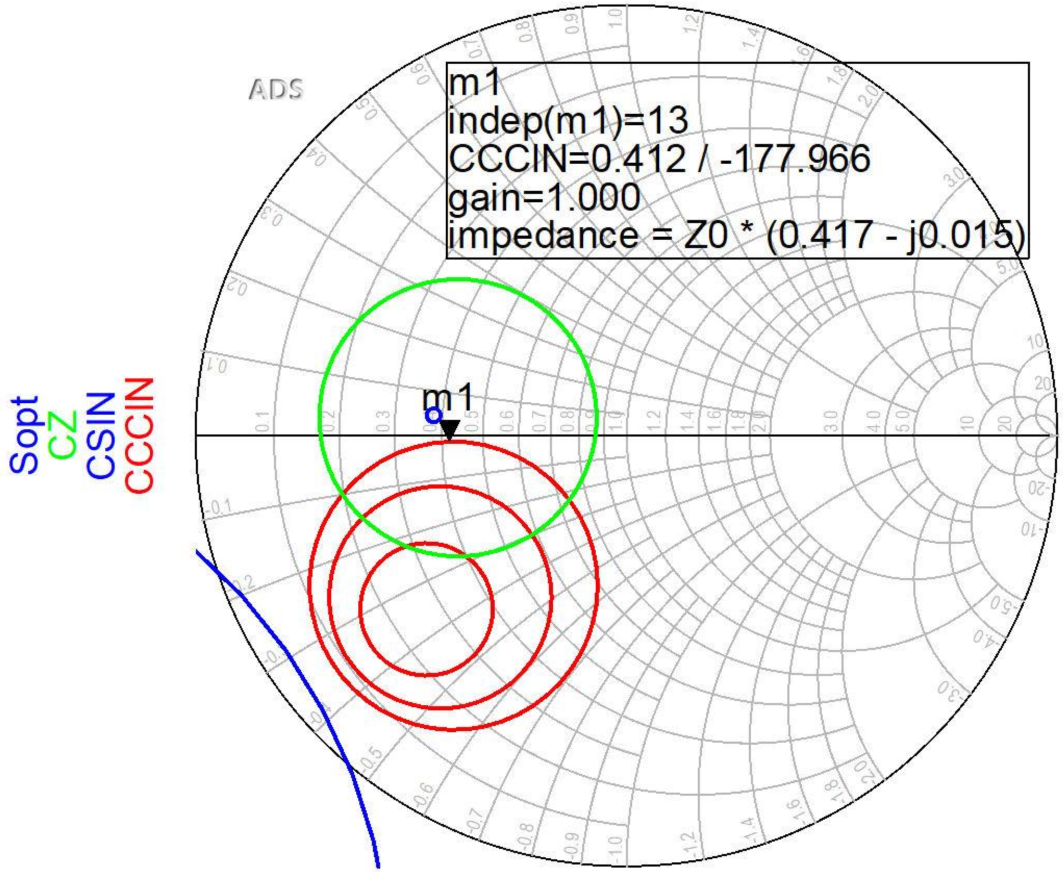
- In this particular case G_{Lmax} = 0.21 dB, the transistor could be used directly connected to the 50Ω load
- The absence of the output matching circuit **is not** recommended. While the attainable power gain is low, it's absence eliminates the possibility to use it to compensate an improper gain generated by the noise optimization of the input matching circuit

Input matching circuit



- For the input matching circuit
  - noise circle CZ: 0.75dB
  - input constant gain circles CCCIN: 1dB, 1.5dB, 2 dB
- We choose (small Q → wide bandwidth) position m1

Input matching circuit



- 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

Input matching circuit

- Position m1 in complex plane (Smith Chart)

$$\Gamma_S = 0.412\angle -178°$$

$$|\Gamma_S| = 0.412; \quad \varphi = -178°$$

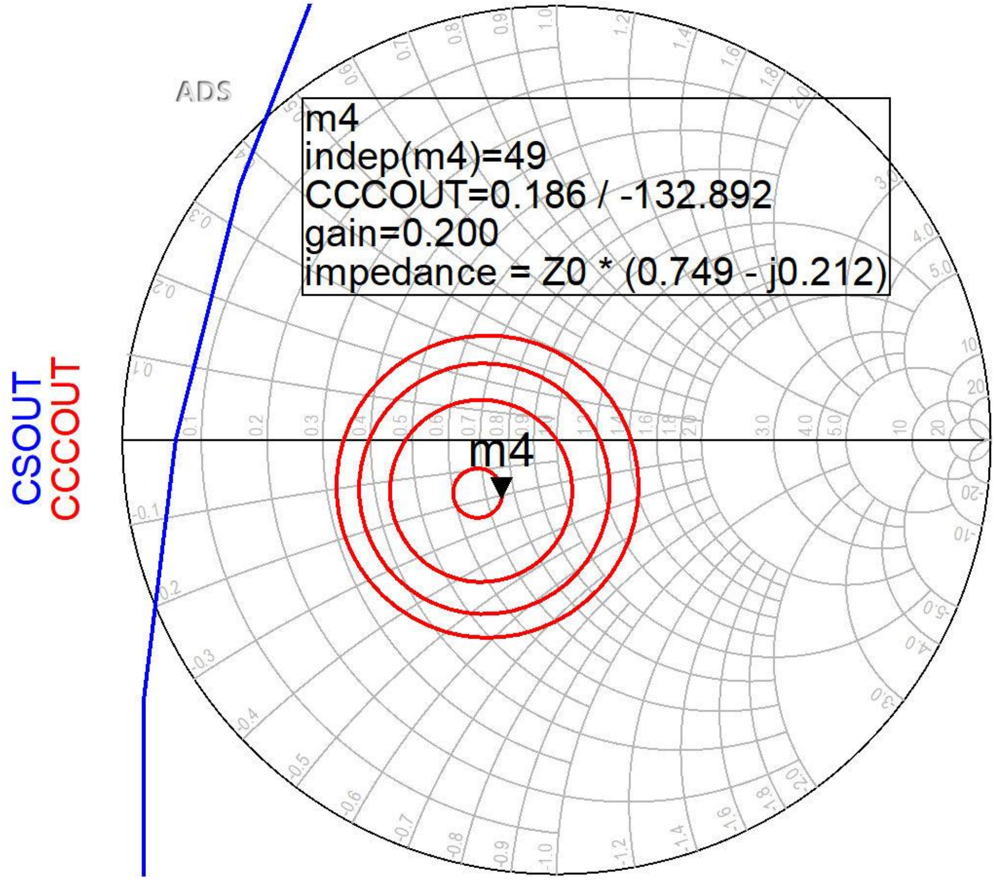
$$\cos(\varphi + 2\theta) = -|\Gamma_S|$$

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

$$\cos(\varphi + 2\theta) = -0.412 \Rightarrow \quad (\varphi + 2\theta) = \pm 114.33°$$

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

Output matching circuit



- 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)

Output matching circuit

- Position m4 in complex plane (Smith Chart)

$$\Gamma_L = 0.186\angle -132.9°$$

$$|\Gamma_L| = 0.186; \quad \varphi = -132.9°$$

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

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

$$\cos(\varphi + 2\theta) = -0.186 \Rightarrow \quad (\varphi + 2\theta) = \pm 100.72°$$

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

LNA

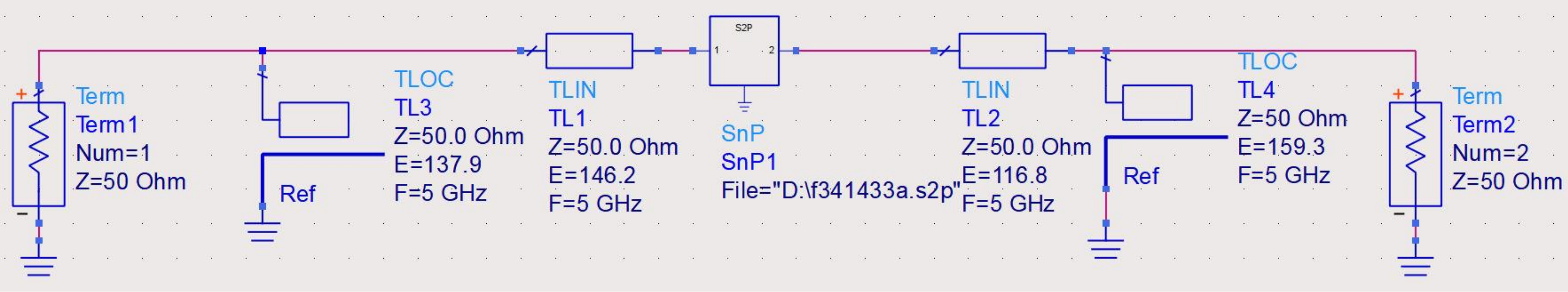
- We estimate a gain (in unilateral assumption, ±0.9 dB)

$$G_T[dB] = G_S[dB] + G_0[dB] + G_L[dB]$$

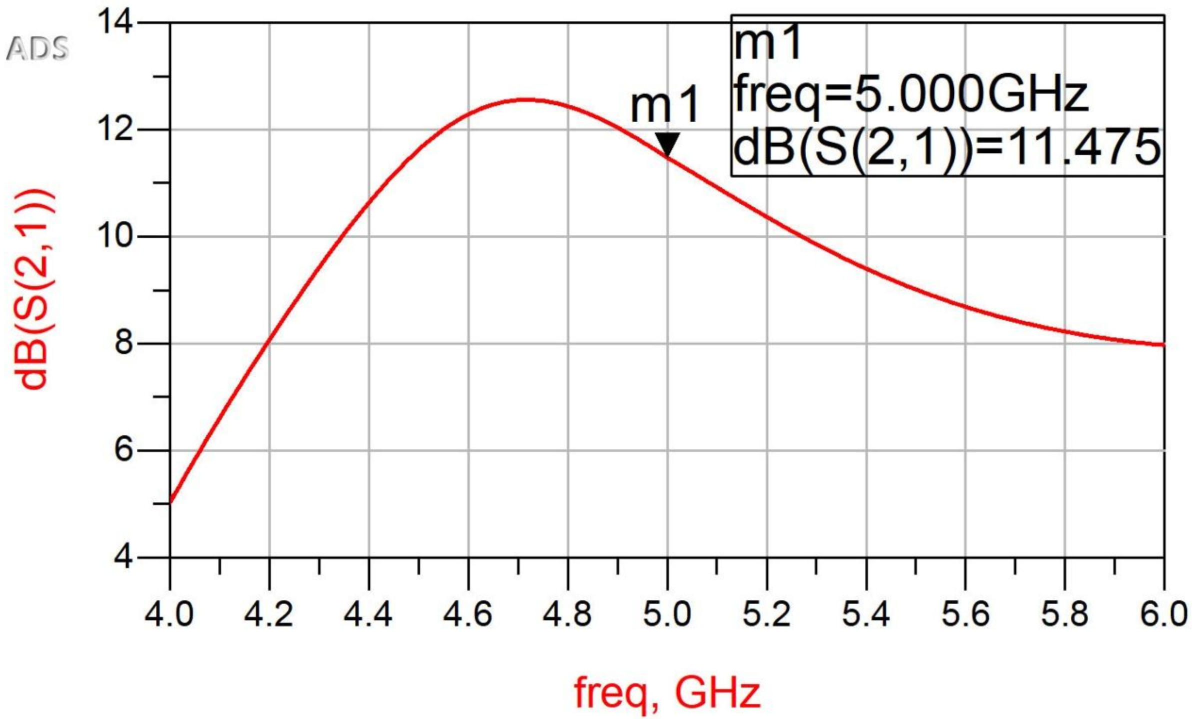
$$G_T[dB] = 1\,dB + 10\,dB + 0.2\,dB = 11.2\,dB$$

- We estimate a noise factor well bellow 0.75dB (quite close to the minimum ~0.6 dB)

ADS

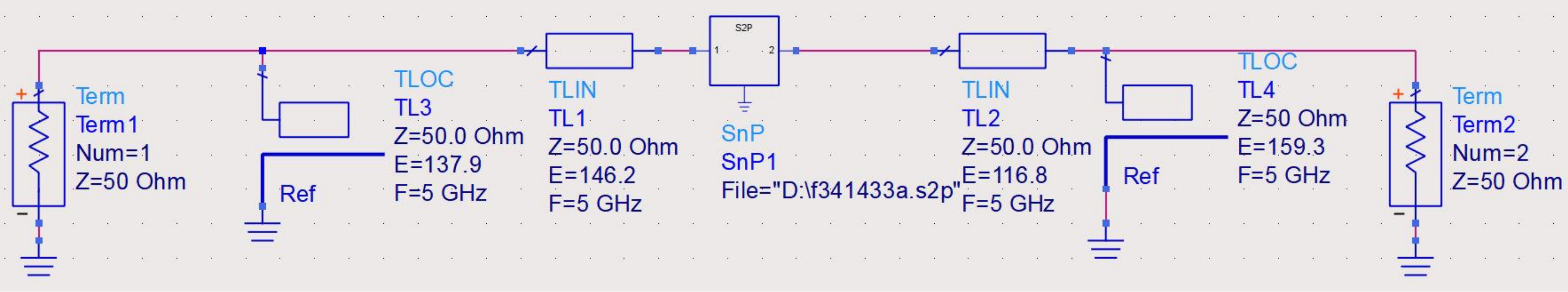


Term
Term1
Num=1
Z=50 Ohm
Ref
TLOC
TL3
Z=50.0 Ohm
E=137.9
F=5 GHz
TLIN
TL1
Z=50.0 Ohm
E=146.2
F=5 GHz
S2P
SnP
SnP1
File="D:\f341433a.s2p"
TLIN
TL2
Z=50.0 Ohm
E=116.8
F=5 GHz
Ref
TLOC
TL4
Z=50 Ohm
E=159.3
F=5 GHz
Term
Term2
Num=2
Z=50 Ohm

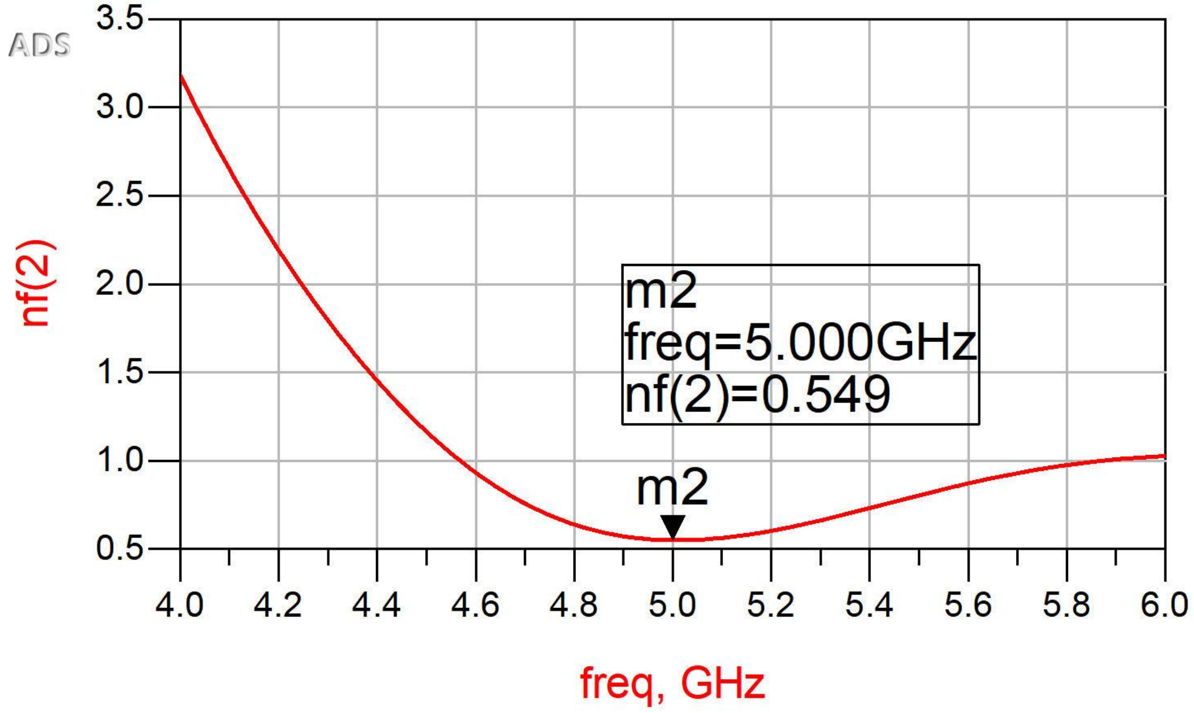


ADS
m1
freq=5.000GHz
dB(S(2,1))=11.475
dB(S(2,1))
freq, GHz

ADS



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
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)
m2
freq=5.000GHz
nf(2)=0.549
m2
freq, GHz

Contact

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