Lecture 13
2025/2026

Microwave Devices and Circuits for Radiocommunications

2025/2026

- 2C/1L, **MDCR**
- **Attendance at minimum 7 sessions (course or laboratory)**
- Lectures- **associate professor Radu Damian**
  - Tuesday **12-14**, **P2**
  - E – 50% final grade
  - **problems** + (2p atten. lect.) + (3 tests) + (bonus activity)
    - first test L1: 24.02.2026 (t2 and t3 not announced, lecture)
    - 3att.=+0.5p
  - **all materials/equipments authorized**

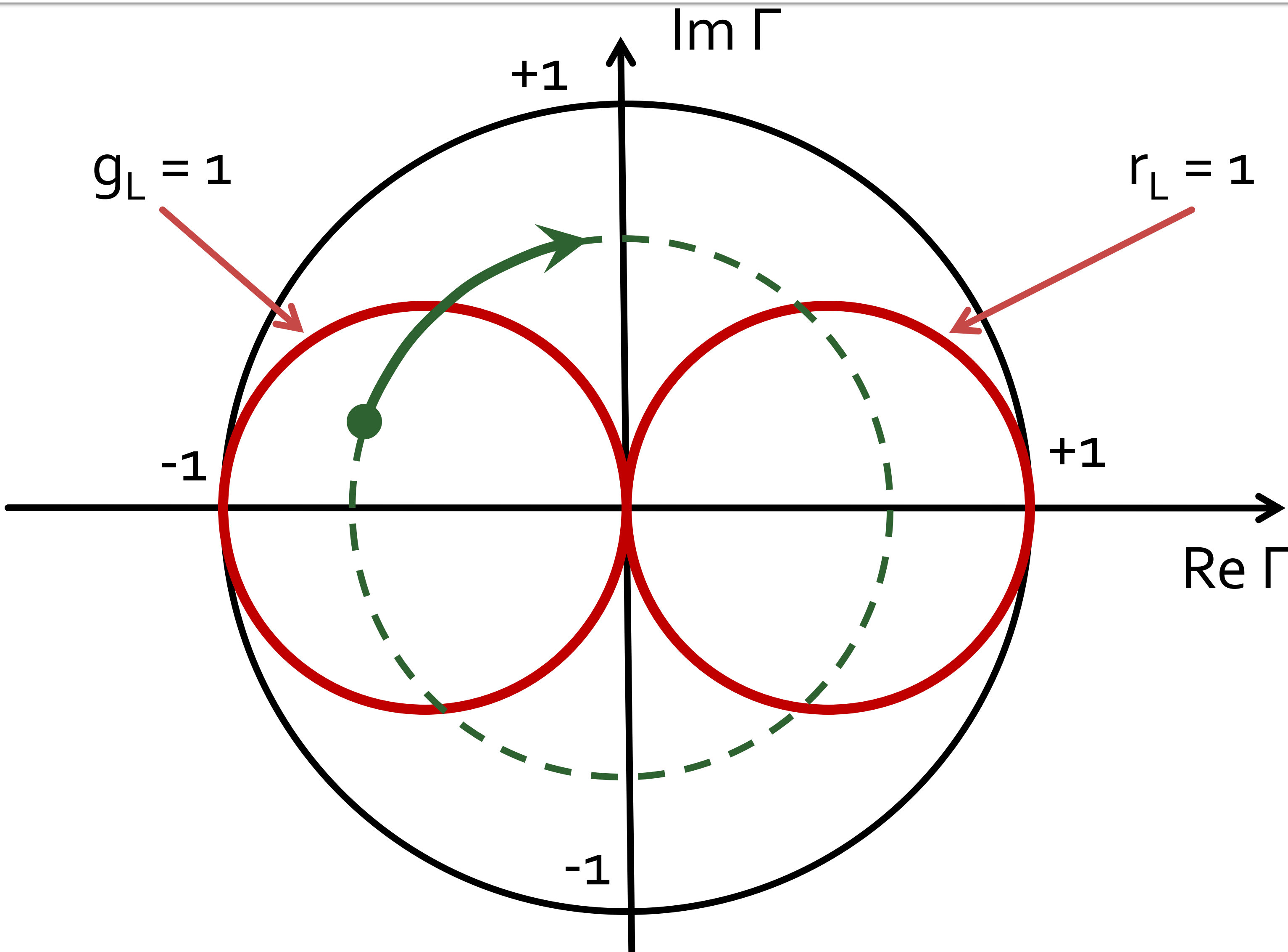
2025/2026

- Laboratory – **associate professor Radu Damian**
  - Monday 14-16, II.13 / (even weeks)
  - L – 25% final grade
    - ADS, 4 sessions
    - Attendance + **personal results**
  - P – 25% final grade
    - ADS, 3 sessions (-1? 24.02.2026)
    - personal homework

Impedance Matching

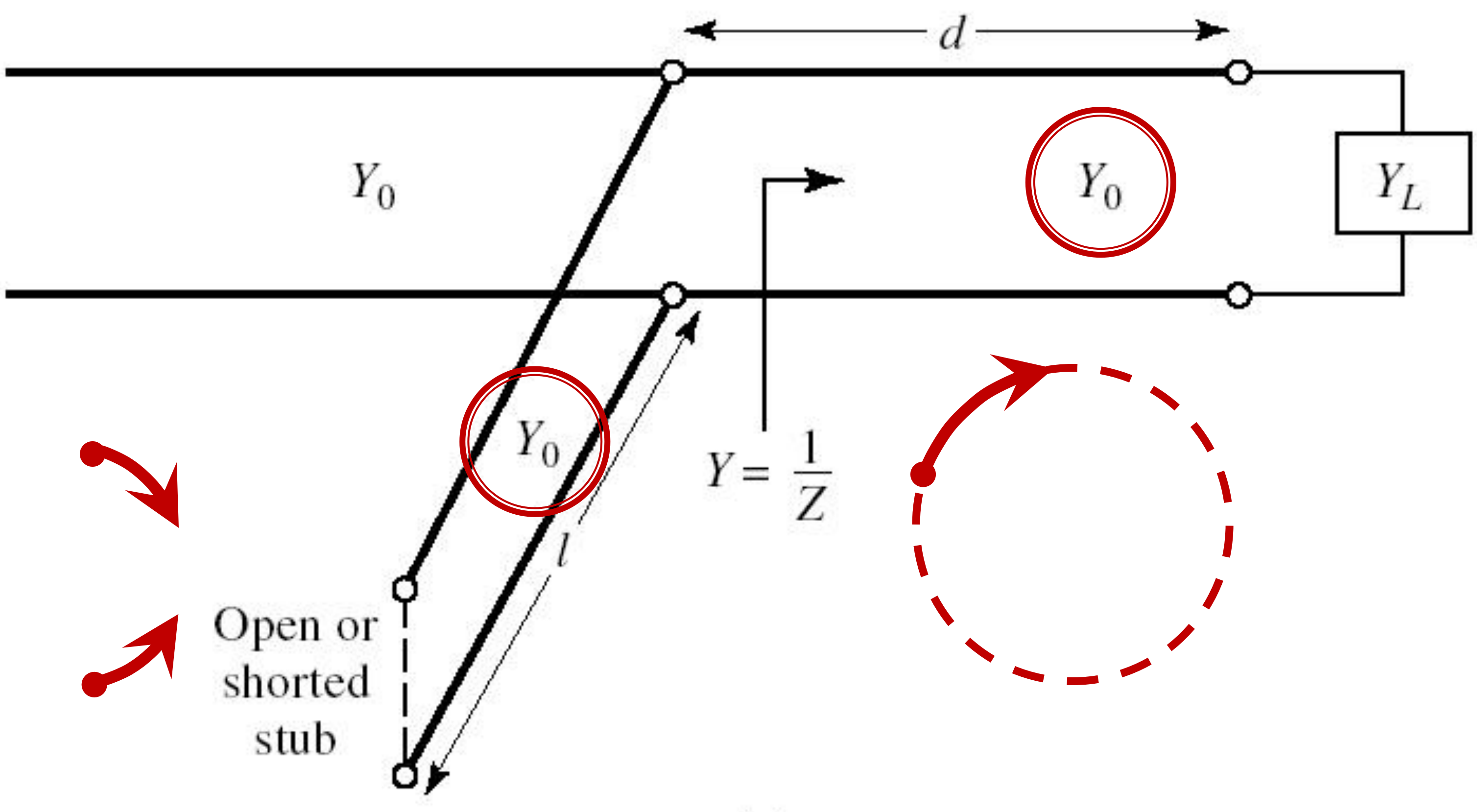
Impedance Matching with Stubs

Smith chart, r=1 and g=1



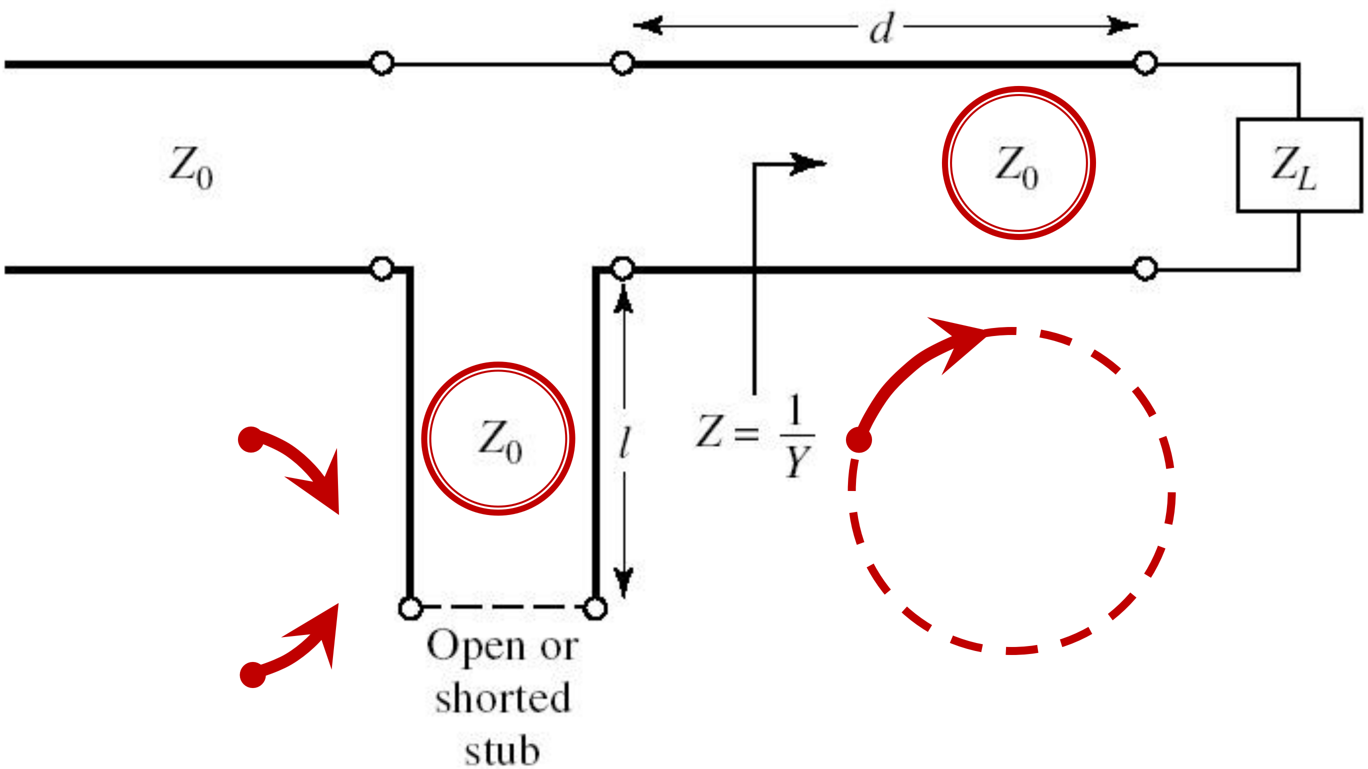
Single stub tuning

- Shunt Stub



Single stub tuning

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

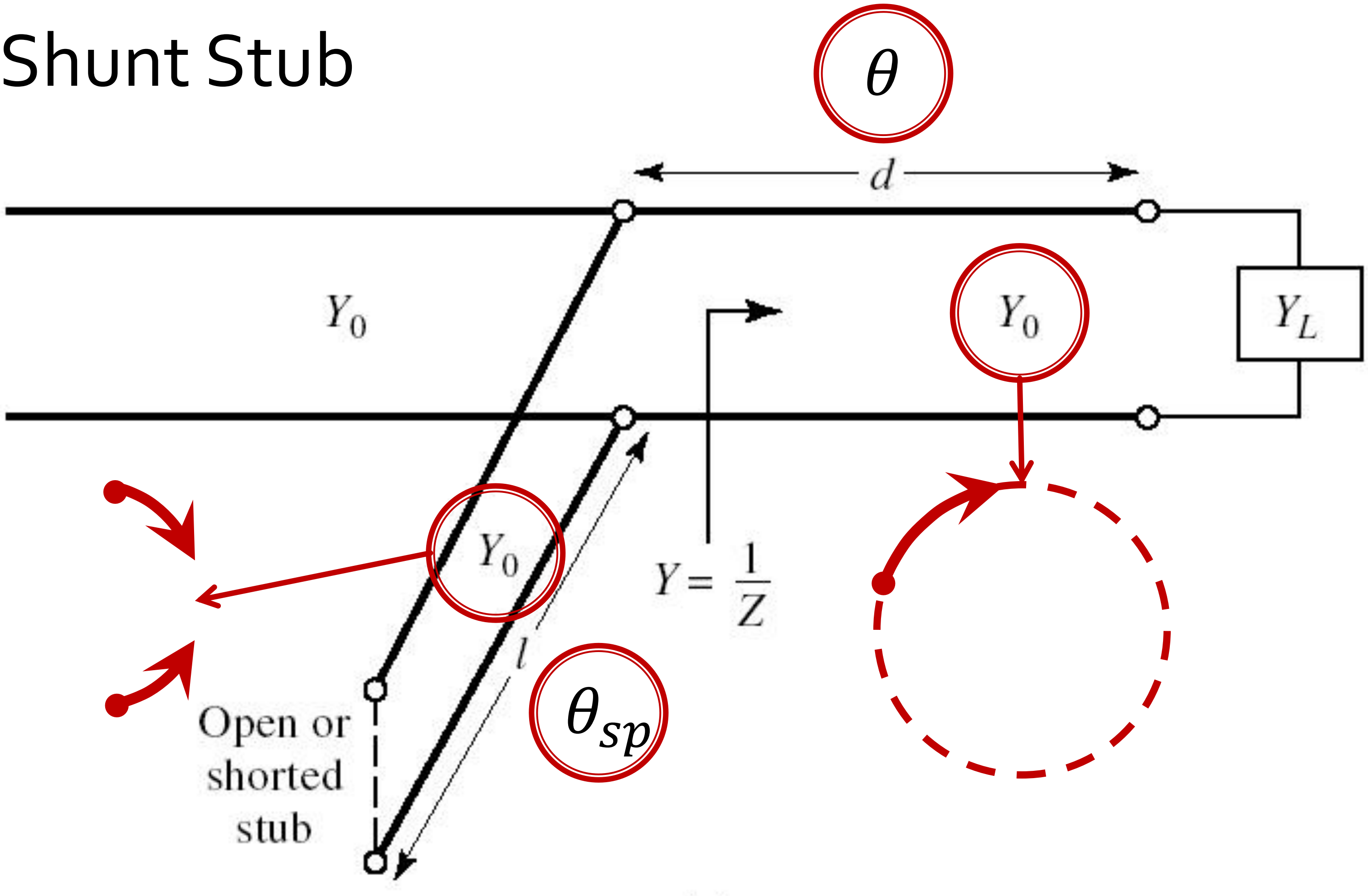


Analytical solutions

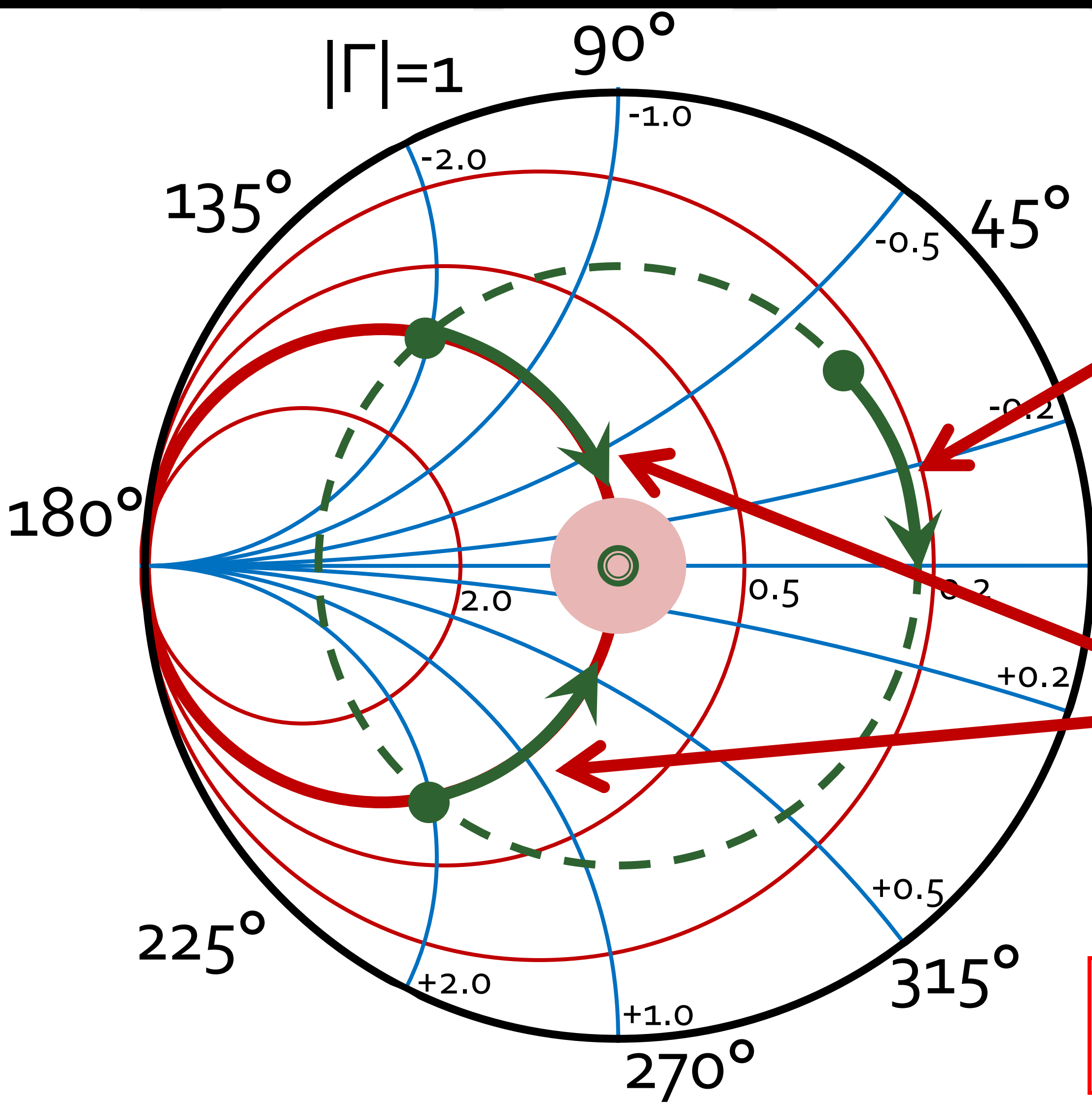
Exam / Project

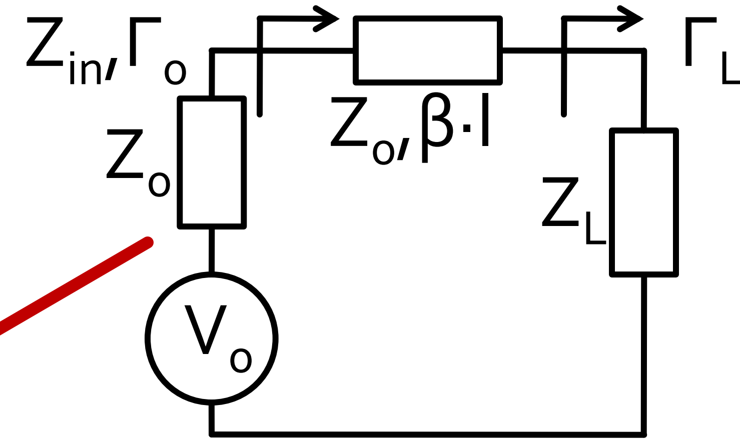
Case 1, Shunt Stub

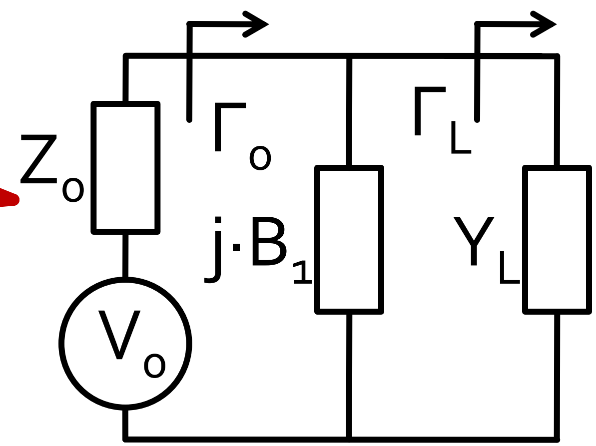
- Shunt Stub



Matching, series line + shunt susceptance







$$\boxed{\left|\Gamma_{in}\right| = \left|\Gamma_L\right|}$$

$$\boxed{g_{in} = 1}$$

Analytical solution, usage

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

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

$$\Gamma_S = 0.593\angle 46.85^\circ$$

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

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

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

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

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

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

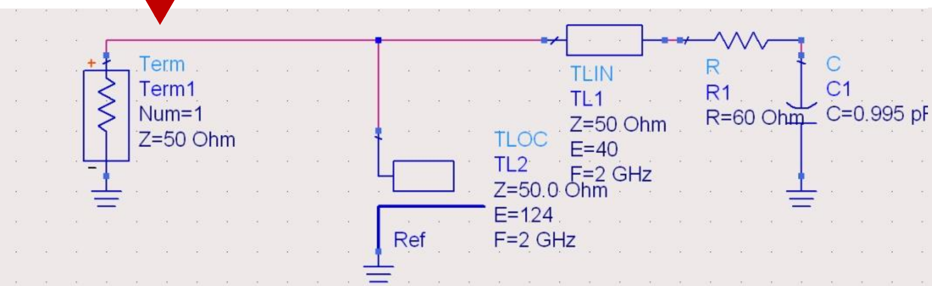
Analytical solution, usage

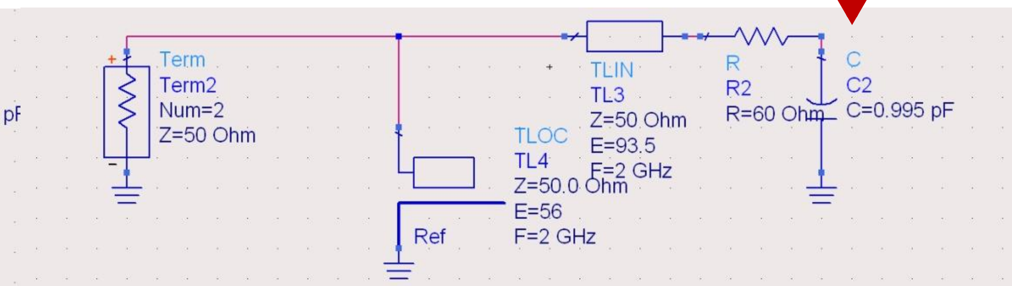
$$(\varphi+2\theta)=\begin{cases}+126.35^\circ\\-126.35^\circ\end{cases}\quad \theta=\begin{cases}39.7^\circ\\93.4^\circ\end{cases}\quad \mathrm{Im}\left[y_S(\theta)\right]=\begin{cases}-1.472\\+1.472\end{cases}\quad \theta_{sp}=\begin{cases}-55.8^\circ+180^\circ=124.2^\circ\\+55.8^\circ\end{cases}$$

- We choose **one** of the two possible solutions
- The **sign** (+/-) chosen for the **series line** equation imposes the **sign** used for the **shunt stub** equation

$$l_1=\frac{39.7^\circ}{360^\circ}\cdot\lambda=0.110\cdot\lambda \qquad l_1=\frac{93.4^\circ}{360^\circ}\cdot\lambda=0.259\cdot\lambda$$

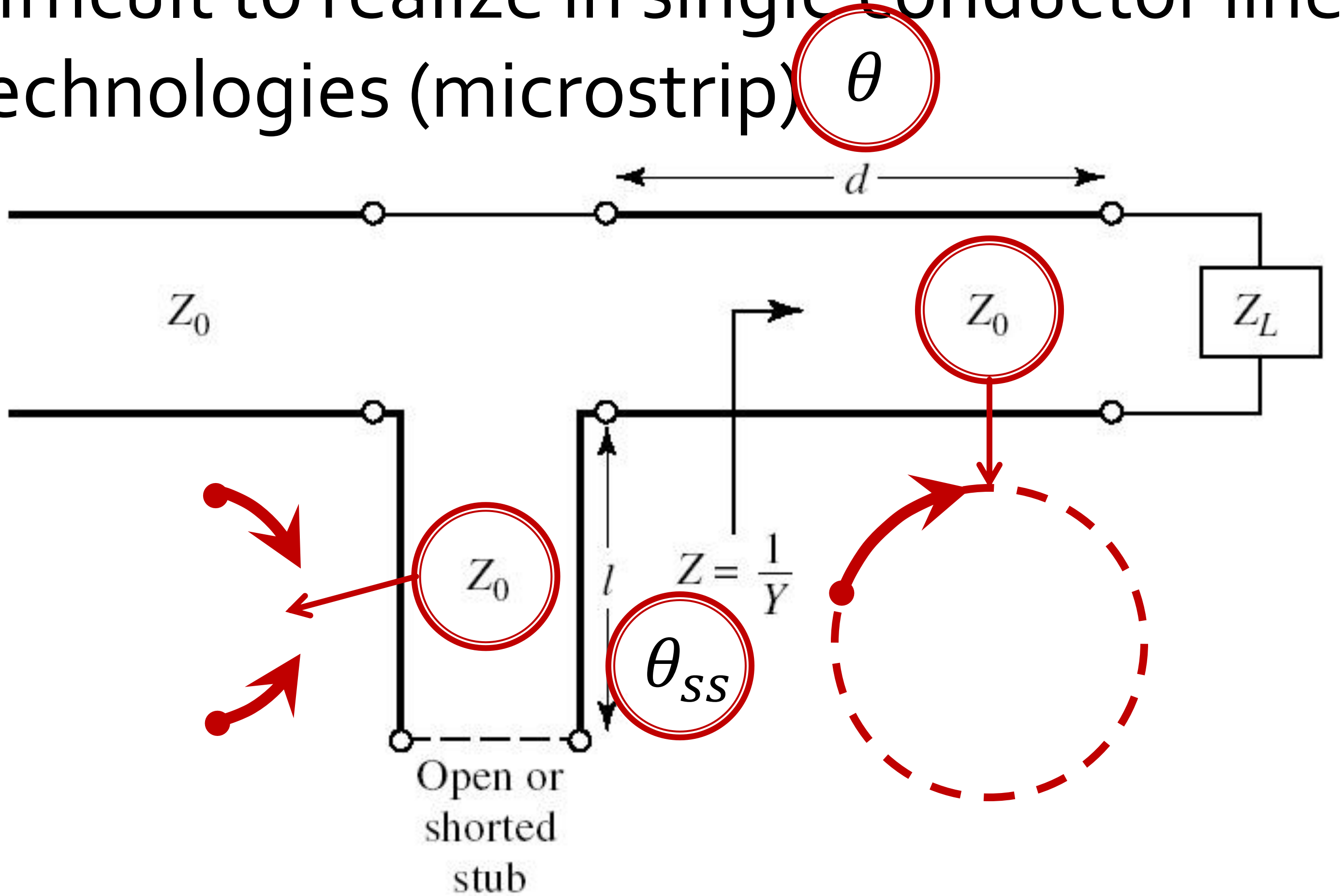
$$l_2=\frac{124.2^\circ}{360^\circ}\cdot\lambda=0.345\cdot\lambda \qquad l_2=\frac{55.8^\circ}{360^\circ}\cdot\lambda=0.155\cdot\lambda$$



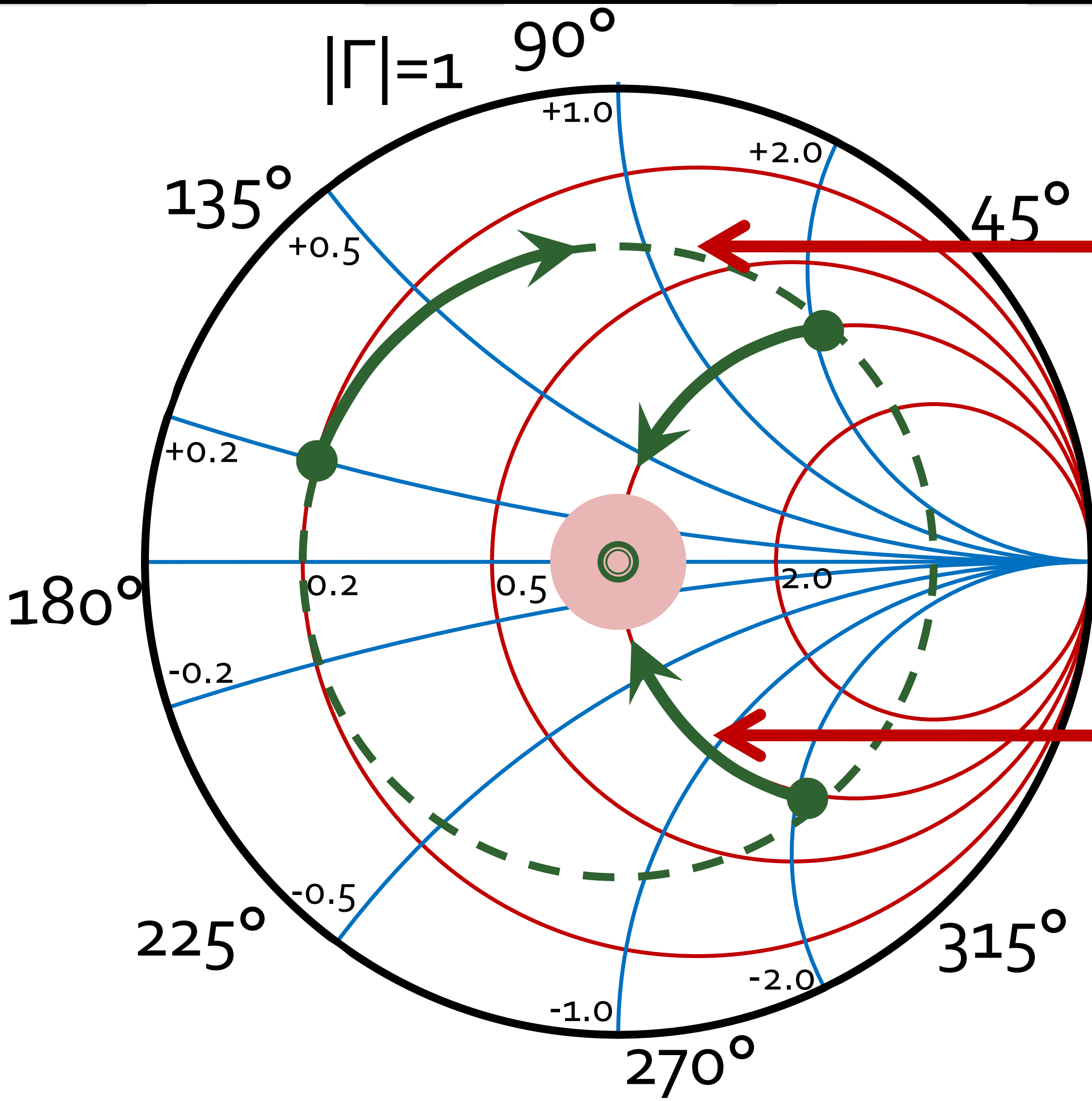


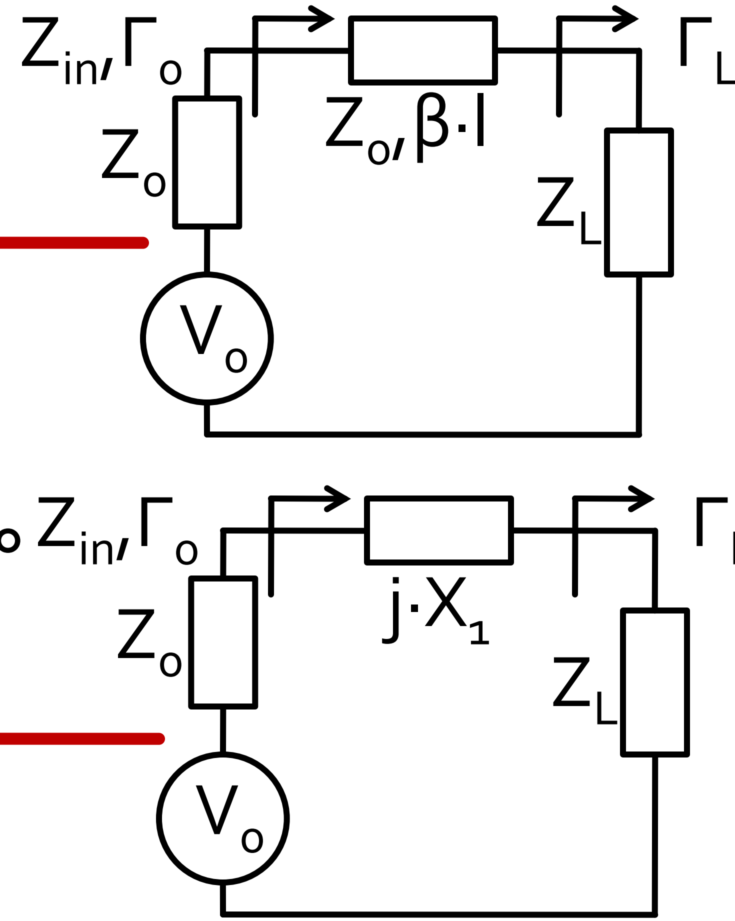
Case 2, Series Stub

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



Matching, series line + series reactance





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

$$r_{in} = 1$$

Analytical solution, usage

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

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

$$\Gamma_S = 0.555\angle -29.92°$$

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

- 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**

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

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

Analytical solution, usage

$$(\varphi+2\theta)=\begin{cases}+56.28^\circ\\-56.28^\circ\end{cases}\quad \theta=\begin{cases}43.1^\circ\\166.8^\circ\end{cases}\quad \mathrm{Im}[z_S(\theta)]=\begin{cases}+1.335\\-1.335\end{cases}\quad \theta_{ss}=\begin{cases}-36.8^\circ+180^\circ=143.2^\circ\\+36.8^\circ\end{cases}$$

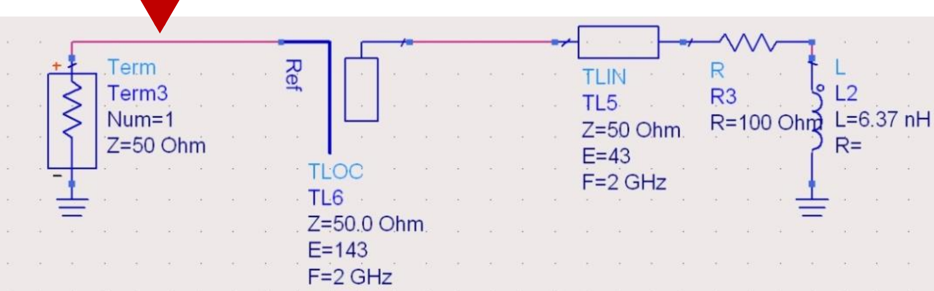
- We choose **one** of the two possible solutions
- The **sign** (+/-) chosen for the **series line** equation imposes the **sign** used for the **series stub** equation

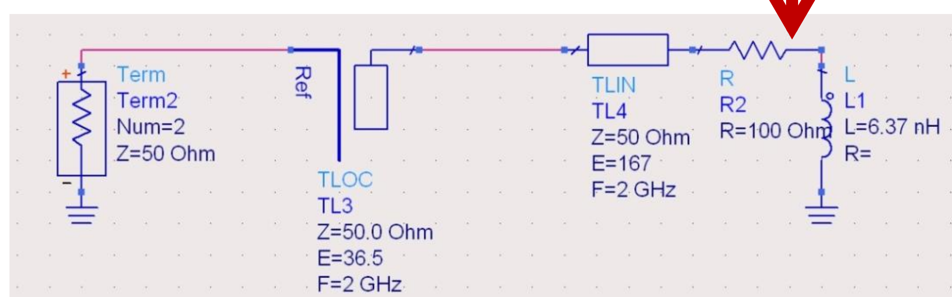
$$l_1=\frac{43.1^\circ}{360^\circ}\cdot\lambda=0.120\cdot\lambda$$

$$l_2=\frac{143.2^\circ}{360^\circ}\cdot\lambda=0.398\cdot\lambda$$

$$l_1=\frac{166.8^\circ}{360^\circ}\cdot\lambda=0.463\cdot\lambda$$

$$l_2=\frac{36.8^\circ}{360^\circ}\cdot\lambda=0.102\cdot\lambda$$





Stub, observations

- adding or subtracting **180°** (λ/2) doesn't change the result (full rotation around the Smith Chart)

$$E = \beta \cdot l = \pi = 180^\circ \qquad l = k \cdot \frac{\lambda}{2}, \forall k \in \mathbf{N}$$

  - if the lines/stubs result with **negative** "length"/ "electrical length" we add λ/2 / 180° to obtain physically realizable lines
- adding or subtracting **90°** (λ/4) change the stub impedance:

$$Z_{in,sc} = j \cdot Z_0 \cdot \tan \beta \cdot l \quad \Leftrightarrow \quad Z_{in,g} = -j \cdot Z_0 \cdot \cot \beta \cdot l$$

  - for the stub we can add or subtract 90° (λ/4) while in the same time changing **open-circuit** ⇔ **short-circuit**

Microwave Filters

Insertion loss method

- We control the power loss ratio/attenuation introduced by the filter:
  - in the passband (pass all frequencies)
  - in the stopband (reject all frequencies)

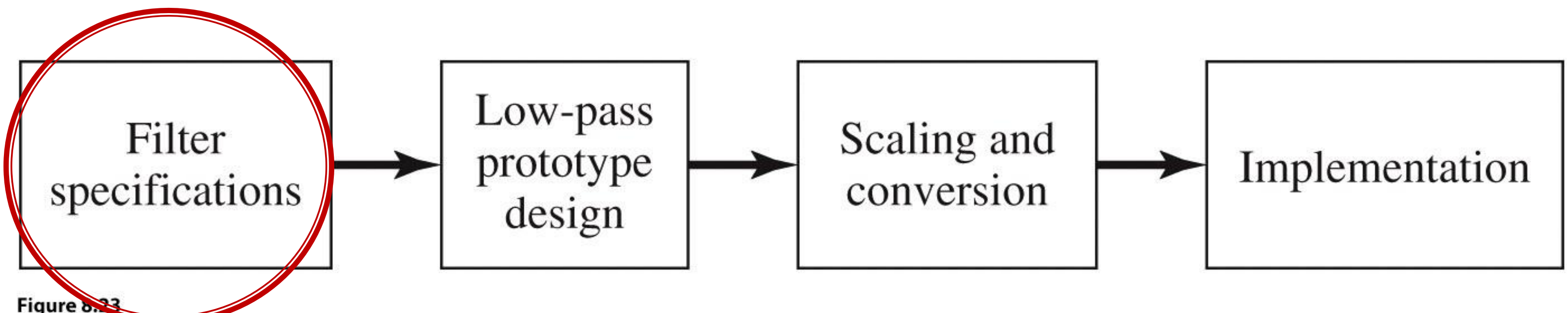
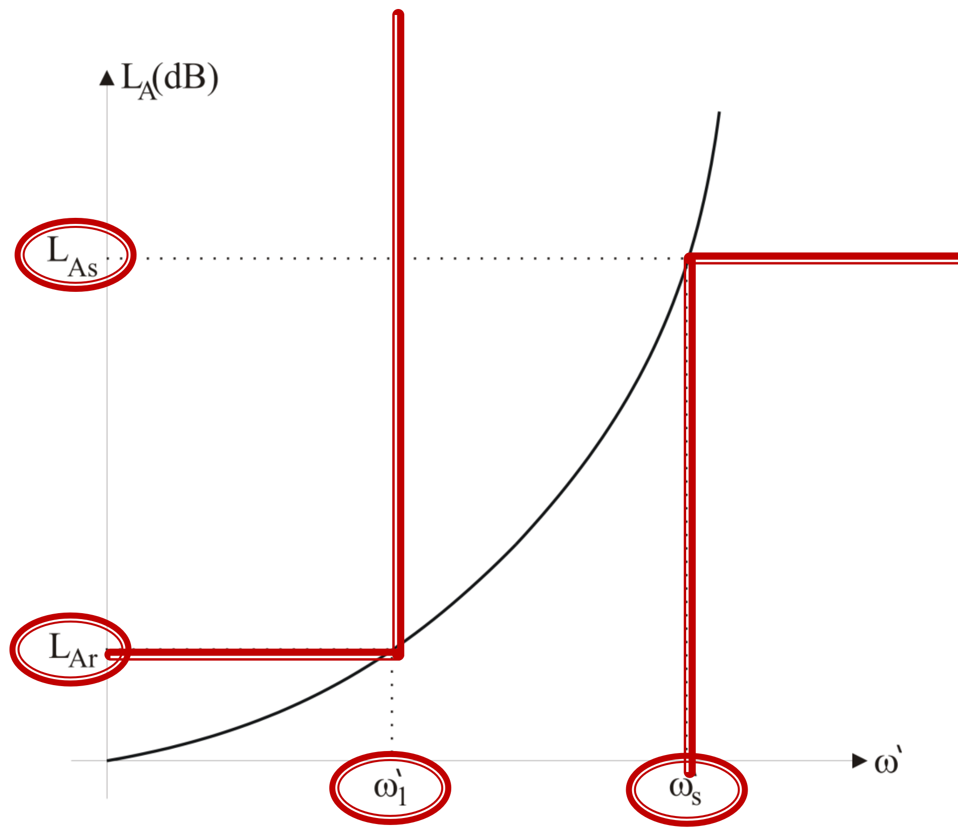


Figure 8.23

Filter specifications

- Attenuation
  - in passband
  - in stopband
  - most often in **dB**
- Frequency range
  - passband
  - stopband
  - cutoff frequency ω_1' usually normalized (**= 1**)



Insertion loss method

- We choose the right polynomials to design an **low-pass** filter (prototype)
- The low-pass prototype are then converted to the desired other types of filters
  - low-pass, high-pass, bandpass, or bandstop

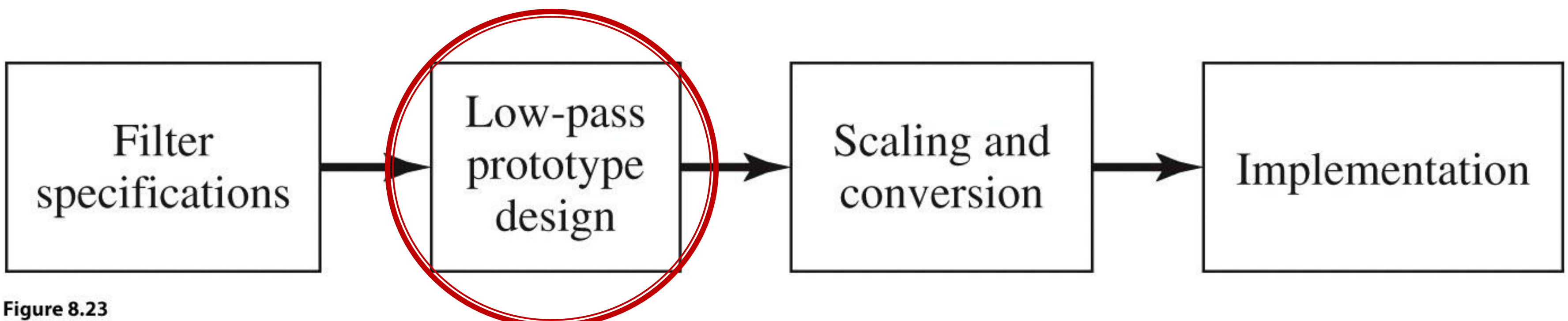


Figure 8.23

Maximally Flat/Equal ripple LPF Prototype

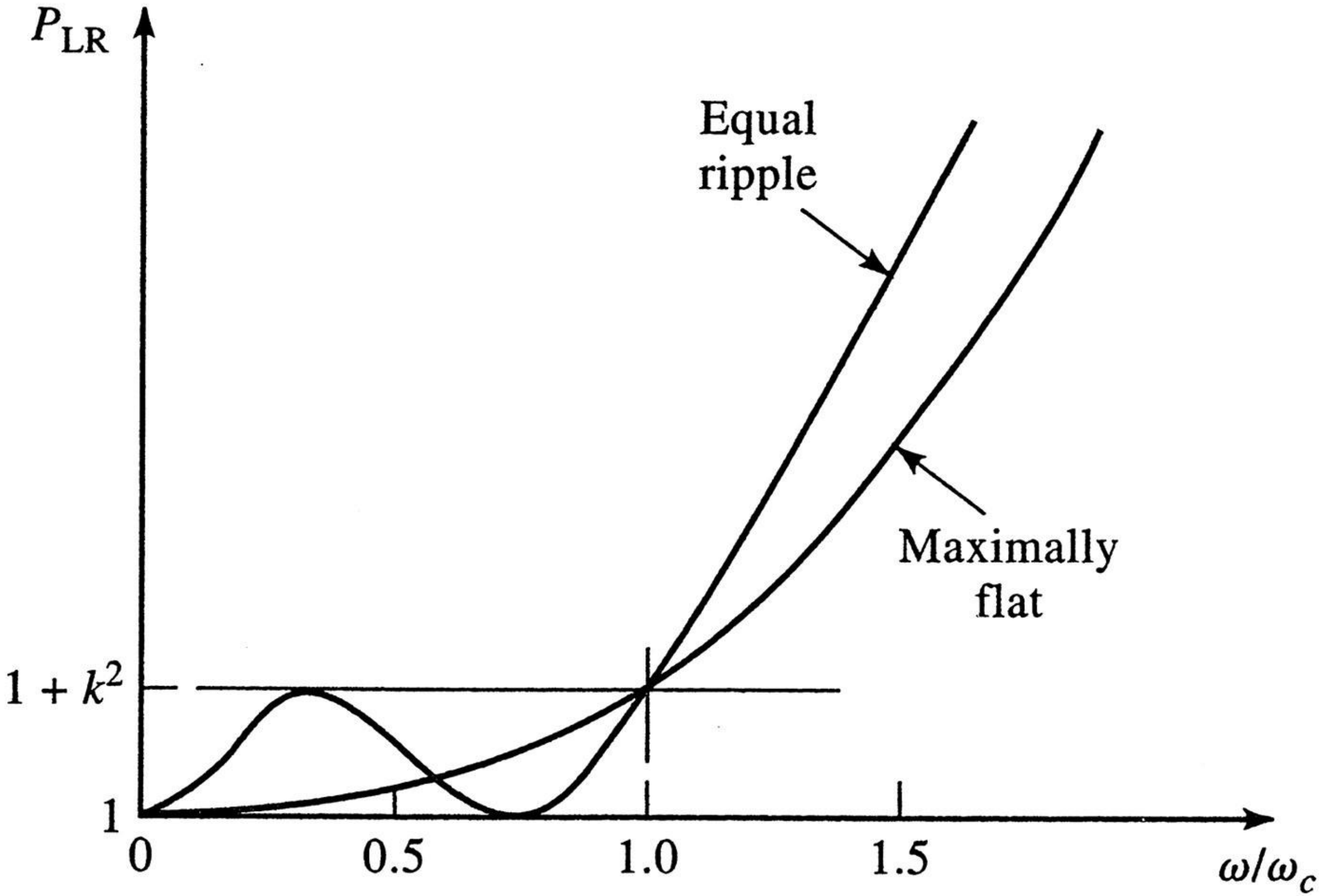
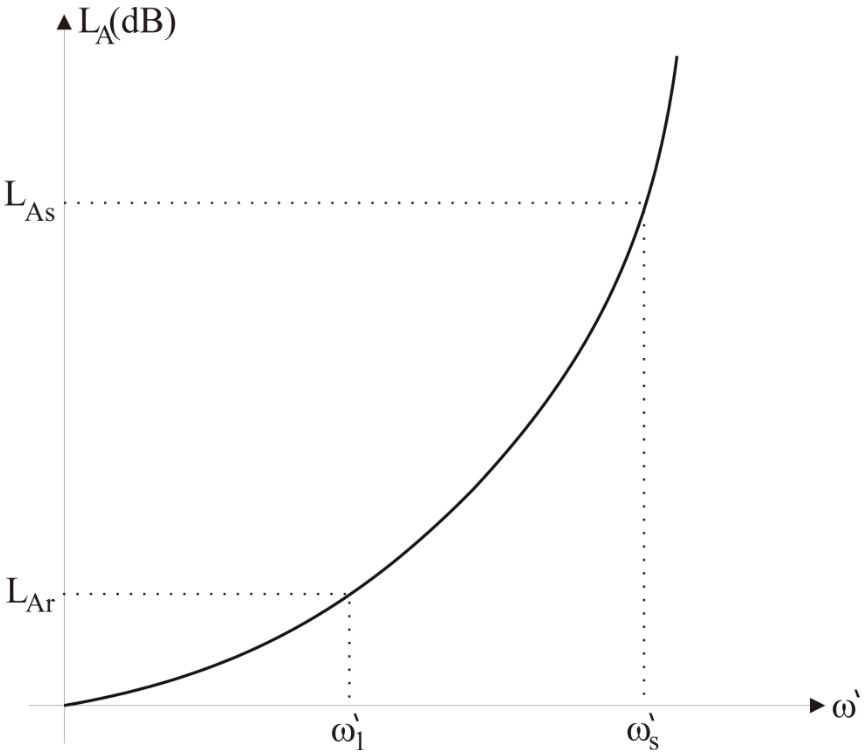


Figure 8.21

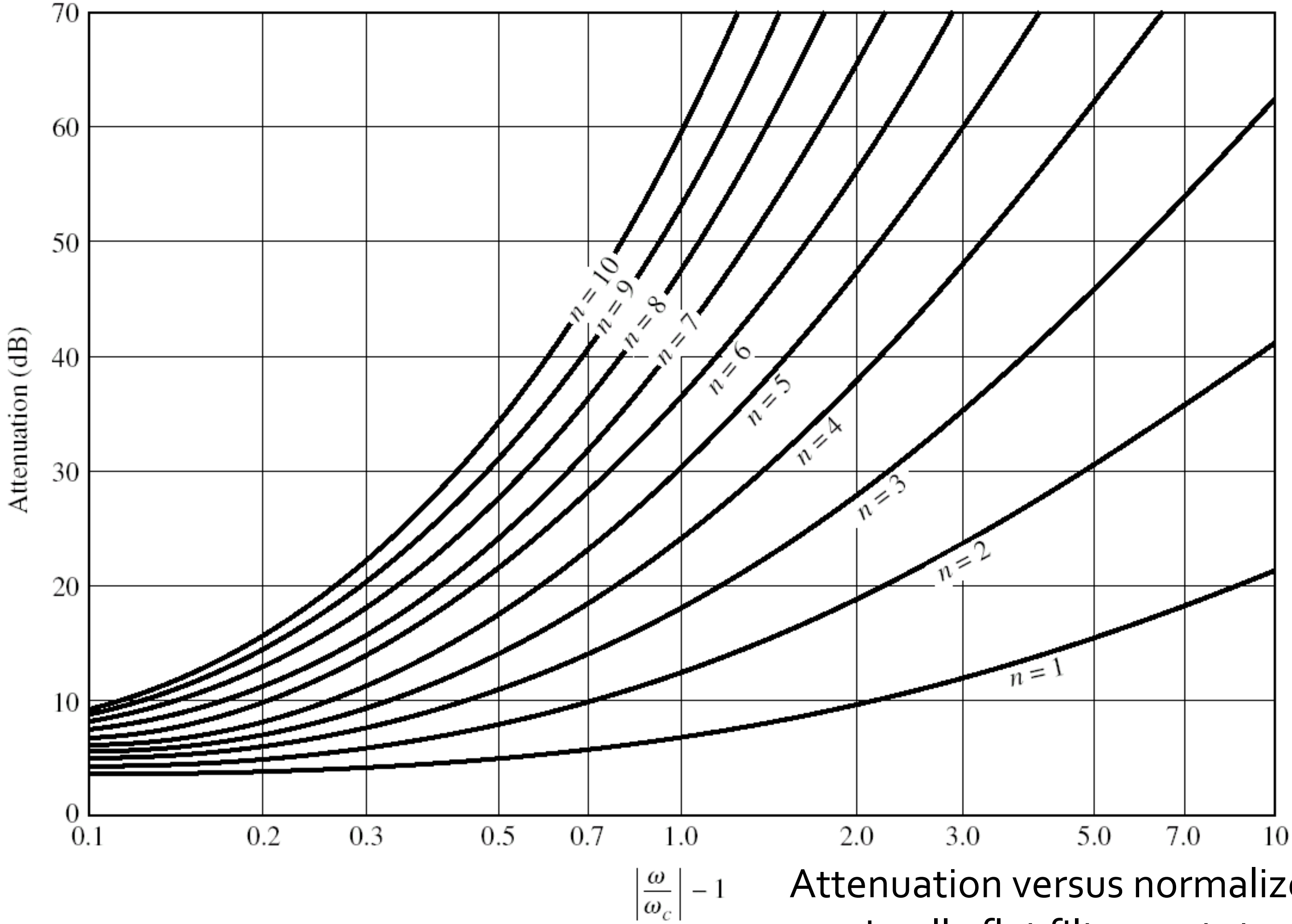
Order (N) of the Maximally Flat filter

$$n \geq \frac{\log\left(\dfrac{10^{\frac{L_{As}}{10}}-1}{10^{\frac{L_{Ar}}{10}}-1}\right)}{2\cdot\log\dfrac{\omega'_s}{\omega'_1}}$$

- !attenuations in **dB**

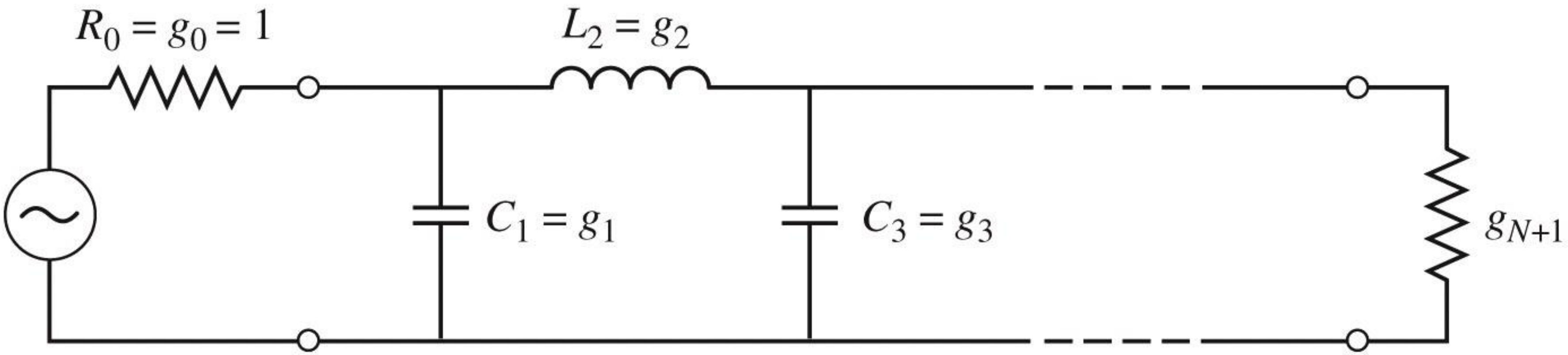


Maximally flat filter prototypes

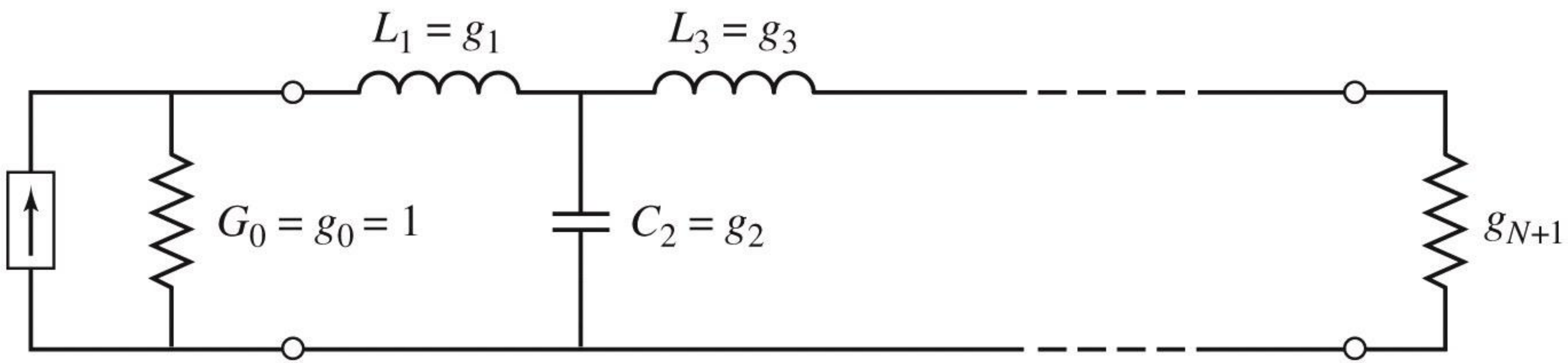


Attenuation versus normalized frequency for maximally flat filter prototypes

Prototype Filters



(a)



(b)

Prototype Filters

- Prototype filters are:
  - Low-Pass Filters (**LPF**)
  - cutoff frequency $\boldsymbol{\omega_0 = 1}$ **rad/s** (f_0 = 0.159 Hz)
  - connected to a source with $\mathbf{R = 1\Omega}$
- The number of reactive elements (L/C) is the order of the filter (N)
- Reactive elements are alternated: series L / shunt C
- There two prototypes with the same response, a prototype beginning with a shunt C element, and a prototype beginning with a series L element

Prototype Filters

- We define filter parameters g_i, i=0,N+1
- g_i are the element values in the prototype filter

$$g_0 = \begin{cases} \text{generator resistance } R'_0 \; if \;\; g_1 = C'_1 \\ \text{generator conductance } G'_0 \; if \;\; g_1 = L'_1 \end{cases}$$

$$g_k \Big|_{k=\overline{1,N}} = \begin{cases} \text{inductance for series inductors} \\ \text{capacitance for shunt capacitors} \end{cases}$$

$$g_{N+1} = \begin{cases} \text{load resistance } R'_{N+1} \; if \;\; g_N = C'_N \\ \text{load conductance } G'_{N+1} \; if \;\; g_N = L'_N \end{cases}$$

Maximally Flat LPF Prototype

- Formulas for filter parameters

$$g_0 = 1$$

$$g_k = 2 \cdot \sin\left[\frac{(2 \cdot k - 1) \cdot \pi}{2 \cdot N}\right] \quad , \quad k = 1, N$$

$$g_{N+1} = 1$$

Maximally Flat LPF Prototype

TABLE 8.3 Element Values for Maximally Flat Low-Pass Filter Prototypes ($g_0 = 1$, $\omega_c = 1$, $N = 1$ to 10)

N	g_1	g_2	g_3	g_4	g_5	g_6	g_7	g_8	g_9	g_{10}	g_{11}
1	2.0000	1.0000									
2	1.4142	1.4142	1.0000								
3	1.0000	2.0000	1.0000	1.0000							
4	0.7654	1.8478	1.8478	0.7654	1.0000						
5	0.6180	1.6180	2.0000	1.6180	0.6180	1.0000					
6	0.5176	1.4142	1.9318	1.9318	1.4142	0.5176	1.0000				
7	0.4450	1.2470	1.8019	2.0000	1.8019	1.2470	0.4450	1.0000			
8	0.3902	1.1111	1.6629	1.9615	1.9615	1.6629	1.1111	0.3902	1.0000		
9	0.3473	1.0000	1.5321	1.8794	2.0000	1.8794	1.5321	1.0000	0.3473	1.0000	
10	0.3129	0.9080	1.4142	1.7820	1.9754	1.9754	1.7820	1.4142	0.9080	0.3129	1.0000

Source: Reprinted from G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance-Matching Networks, and Coupling Structures*, Artech House, Dedham, Mass., 1980, with permission.

Table 8.3

TABLE 8.4 Element Values for Equal-Ripple Low-Pass Filter Prototypes ($g_0 = 1$, $\omega_c = 1$, $N = 1$ to 10, 0.5 dB and 3.0 dB ripple)

N	g_1	g_2	g_3	g_4	g_5	g_6	g_7	g_8	g_9	g_{10}	g_{11}
0.5 dB Ripple											
1	0.6986	1.0000									
2	1.4029	0.7071	1.9841								
3	1.5963	1.0967	1.5963	1.0000							
4	1.6703	1.1926	2.3661	0.8419	1.9841						
5	1.7058	1.2296	2.5408	1.2296	1.7058	1.0000					
6	1.7254	1.2479	2.6064	1.3137	2.4758	0.8696	1.9841				
7	1.7372	1.2583	2.6381	1.3444	2.6381	1.2583	1.7372	1.0000			
8	1.7451	1.2647	2.6564	1.3590	2.6964	1.3389	2.5093	0.8796	1.9841		
9	1.7504	1.2690	2.6678	1.3673	2.7239	1.3673	2.6678	1.2690	1.7504	1.0000	
10	1.7543	1.2721	2.6754	1.3725	2.7392	1.3806	2.7231	1.3485	2.5239	0.8842	1.9841
3.0 dB Ripple											
N	g_1	g_2	g_3	g_4	g_5	g_6	g_7	g_8	g_9	g_{10}	g_{11}
1	1.9953	1.0000									
2	3.1013	0.5339	5.8095								
3	3.3487	0.7117	3.3487	1.0000							
4	3.4389	0.7483	4.3471	0.5920	5.8095						
5	3.4817	0.7618	4.5381	0.7618	3.4817	1.0000					
6	3.5045	0.7685	4.6061	0.7929	4.4641	0.6033	5.8095				
7	3.5182	0.7723	4.6386	0.8039	4.6386	0.7723	3.5182	1.0000			
8	3.5277	0.7745	4.6575	0.8089	4.6990	0.8018	4.4990	0.6073	5.8095		
9	3.5340	0.7760	4.6692	0.8118	4.7272	0.8118	4.6692	0.7760	3.5340	1.0000	
10	3.5384	0.7771	4.6768	0.8136	4.7425	0.8164	4.7260	0.8051	4.5142	0.6091	5.8095

Source: Reprinted from G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance-Matching Networks, and Coupling Structures*, Artech House, Dedham, Mass.,1980, with permission.

Table 8.4

- For even N order of the filter (N = 2, 4, 6, 8 ...) equal-ripple filters **must** closed by a load impedance **$g_{N+1} \neq 1$**
- If the application doesn't allow this, supplemental impedance matching is required (quarter-wave transformer, binomial ...) to **$g_L = 1$**

Impedance and Frequency Scaling

- LPF Prototype is only used as an intermediate step
  - Low-Pass Filter (LPF)
  - cutoff frequency $\omega_o = 1$ rad/s ($f_o = 0.159$ Hz)
  - connected to a source with $R = 1\Omega$

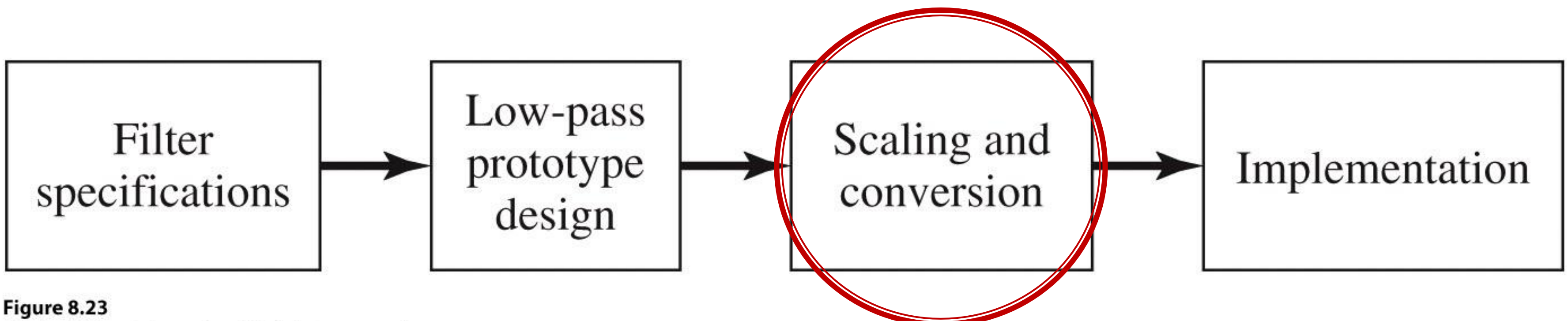


Figure 8.23

Impedance Scaling

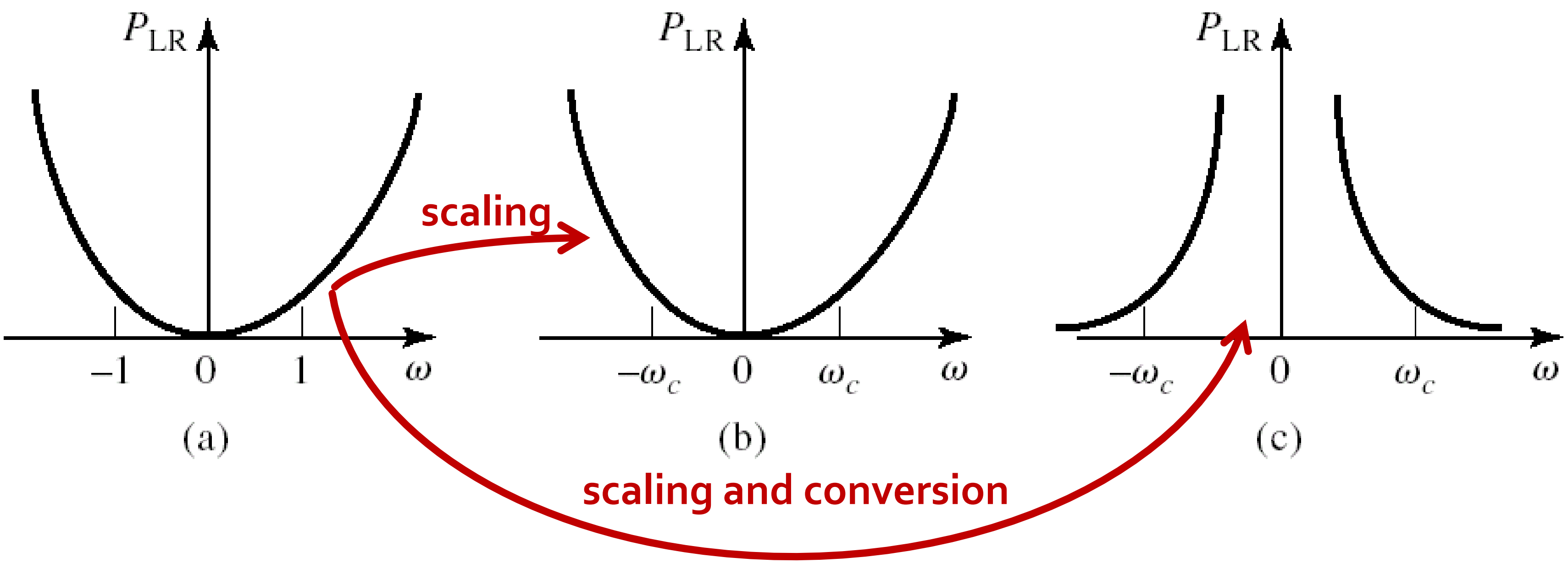
- To design a filter which will work with a source resistance of R_o we multiplying all the impedances of the prototype design by R_o (" ' " denotes scaled values)

$$R_s' = R_0 \cdot \left(R_s = 1\right) \qquad R_L' = R_0 \cdot R_L$$

$$L' = R_0 \cdot L \qquad C' = \frac{C}{R_0}$$

Frequency Scaling

- changing the cutoff frequency – (fig. b)
- changing the type (for example LPF → HPF – fig. c) requires also conversion



Summary of Prototype Filter Transformations

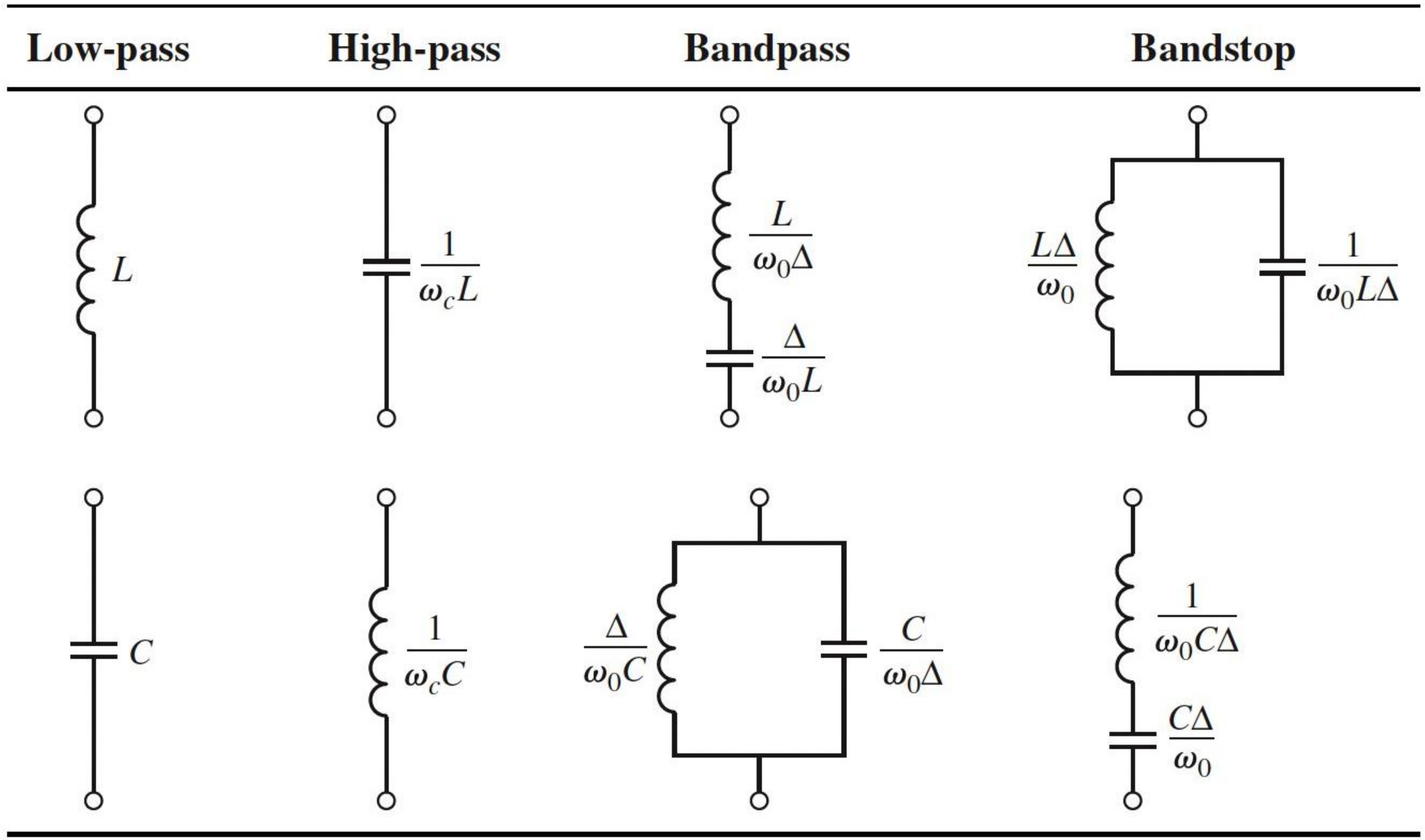


Table 8.6

Microwave Filters Implementation

- The lumped-element (L, C) filter design generally works well **only** at low frequencies (RF):
  - lumped-element inductors and capacitors are generally available only for a limited range of values, and can be difficult to implement at microwave frequencies
  - difficulty to obtain the (very low) required tolerance for elements

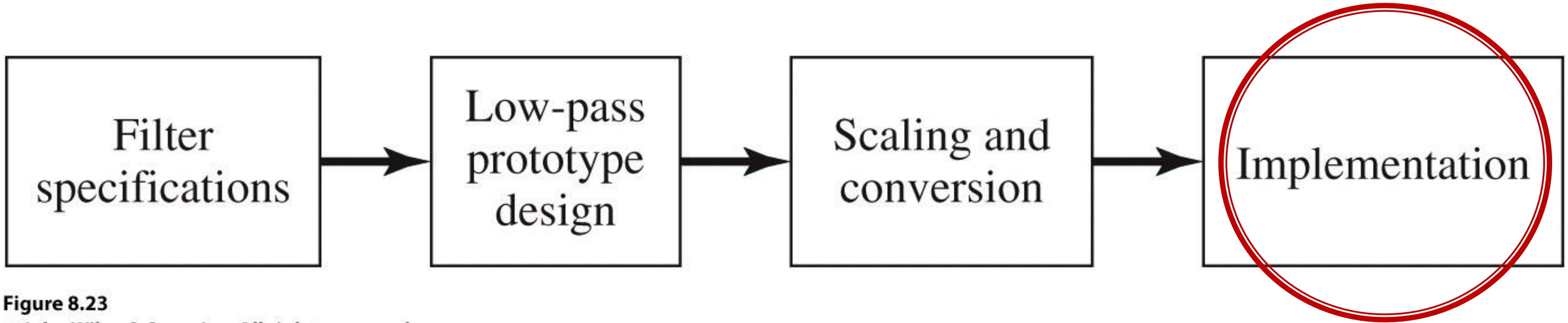
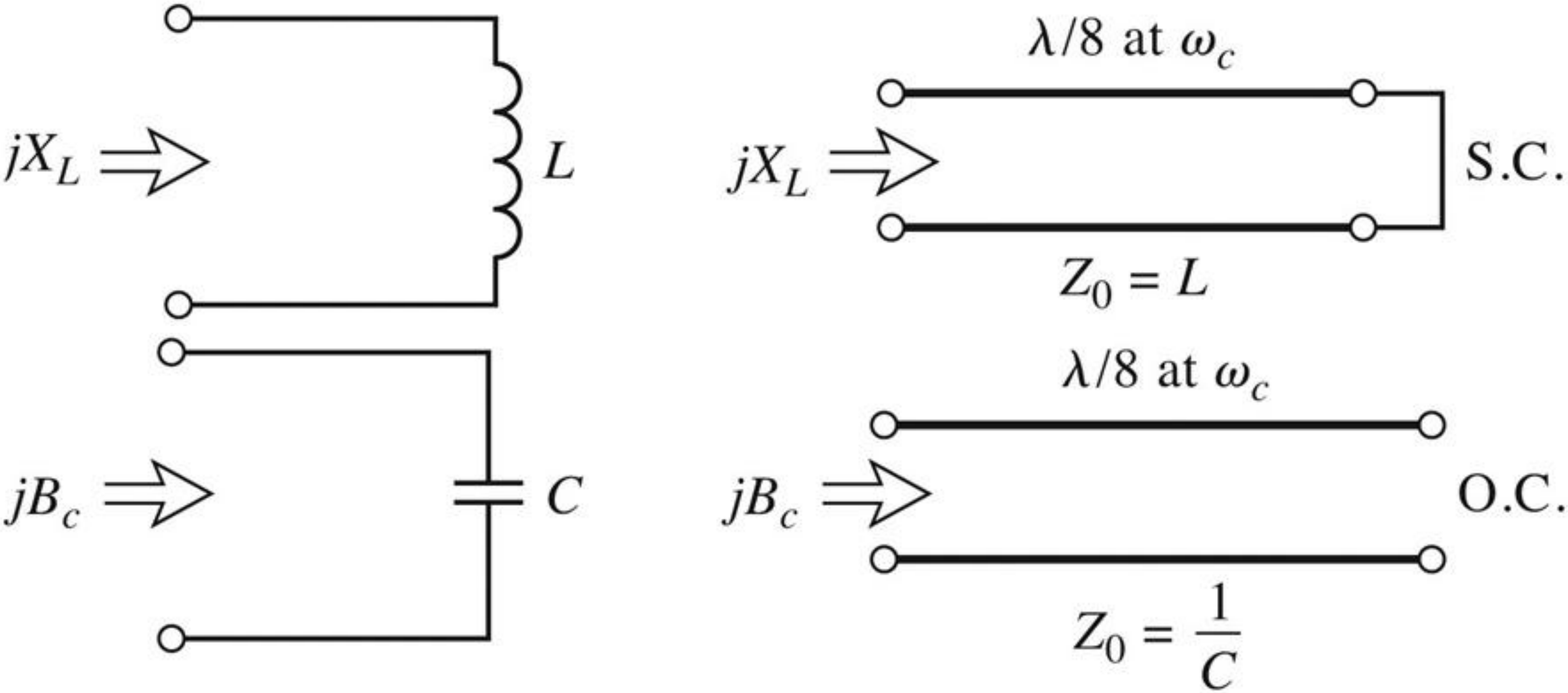


Figure 8.23

Richards' Transformation

- allows implementation of the inductors and capacitors with lines **after** the transformation of the LPF prototype to the required type (LPF/HPF/BPF/BSF)



Impedance and Admittance Inverters

- For cases where Richards + Kuroda do not offer practical solutions we use circuits called **impedance and admittance inverters**

$$Z_{in} = \frac{K^2}{Z_L} \qquad\qquad Y_{in} = \frac{J^2}{Y_L}$$

Impedance inverters　　　　Admittance inverters

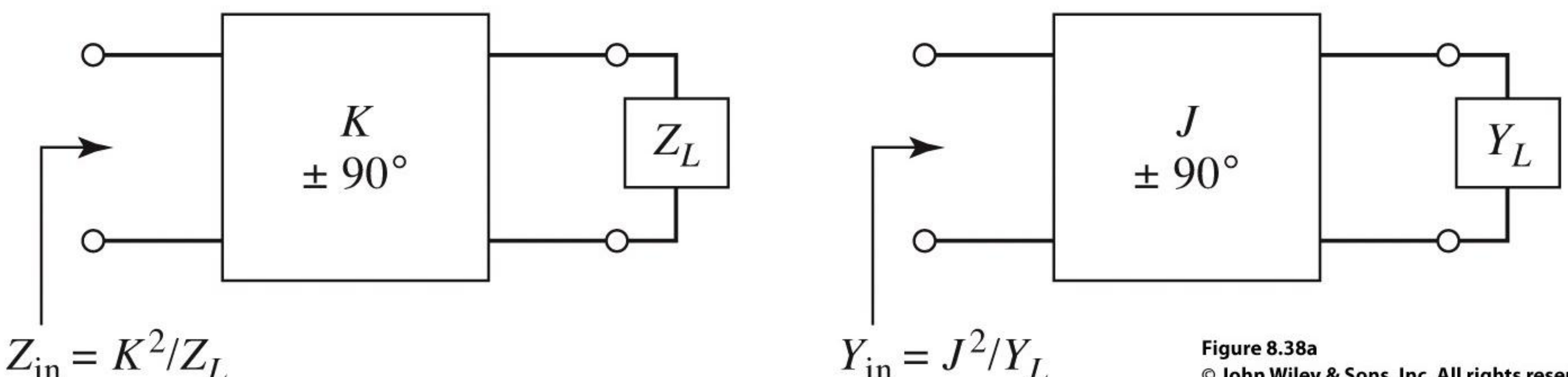
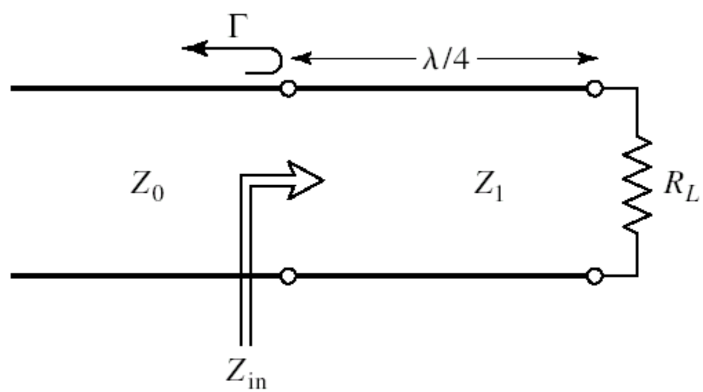


Figure 8.38a

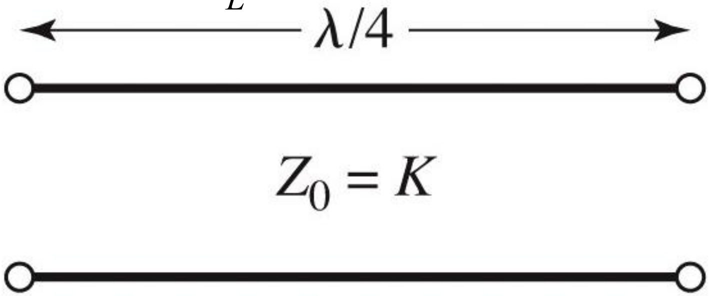
Impedance and Admittance Inverters

- The simplest example of impedance and admittance inverter is the **quarter-wave transformer** (L3)



$$Z_{in} = \frac{Z_1^2}{R_L}$$

$$Z_{in} = \frac{K^2}{Z_L}$$



$$Y_{in} = \frac{J^2}{Y_L}$$

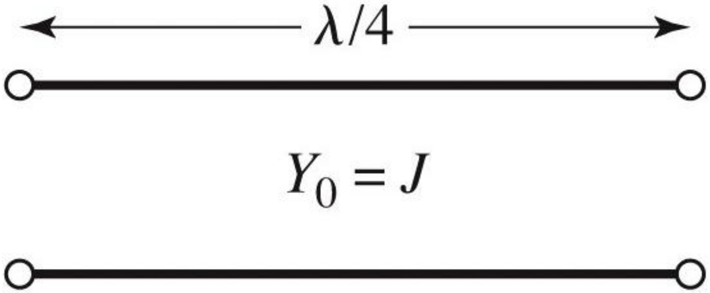
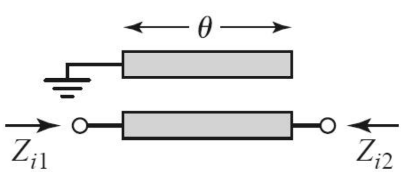
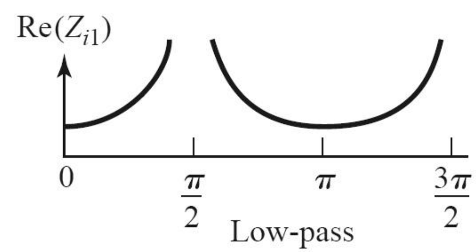
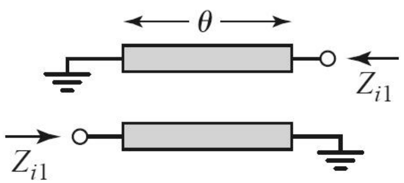
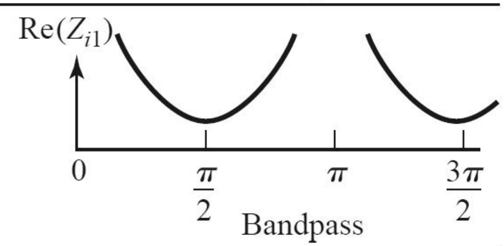
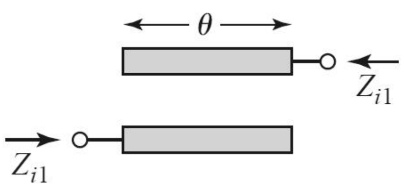
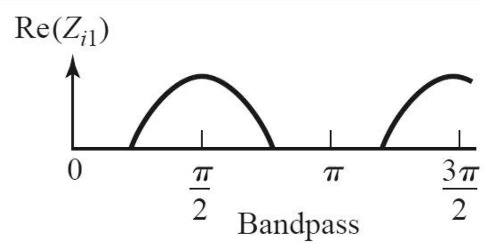
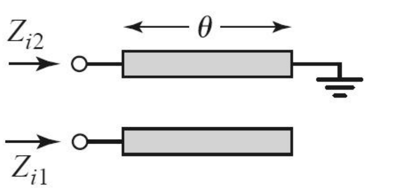
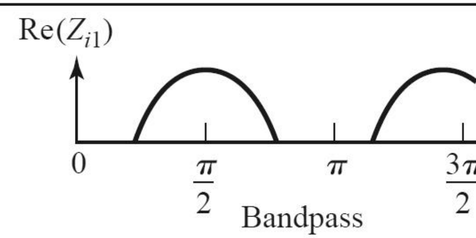


Figure 8.38b

Coupled Line Filters

Circuit	Image Impedance	Response
	$Z_{i1} = \dfrac{2Z_{0e}Z_{0o}\cos\theta}{\sqrt{(Z_{0e}+Z_{0o})^2\cos^2\theta-(Z_{0e}-Z_{0o})^2}}$ $Z_{i2} = \dfrac{Z_{0e}Z_{0o}}{Z_{i1}}$	
	$Z_{i1} = \dfrac{2Z_{0e}Z_{0o}\sin\theta}{\sqrt{(Z_{0e}-Z_{0o})^2-(Z_{0e}+Z_{0o})^2\cos^2\theta}}$	
	$Z_{i1} = \dfrac{\sqrt{(Z_{0e}-Z_{0o})^2-(Z_{0e}+Z_{0o})^2\cos^2\theta}}{2\sin\theta}$	
	$Z_{i1} = \dfrac{\sqrt{Z_{0e}Z_{0o}}\sqrt{(Z_{0e}-Z_{0o})^2-(Z_{0e}+Z_{0o})^2\cos^2\theta}}{(Z_{0e}+Z_{0o})\sin\theta}$ $Z_{i2} = \dfrac{Z_{0e}Z_{0o}}{Z_{i1}}$	

Coupled Line Filters

- Bandpass filter with resonance at θ=π/2 (*l*=λ/4)

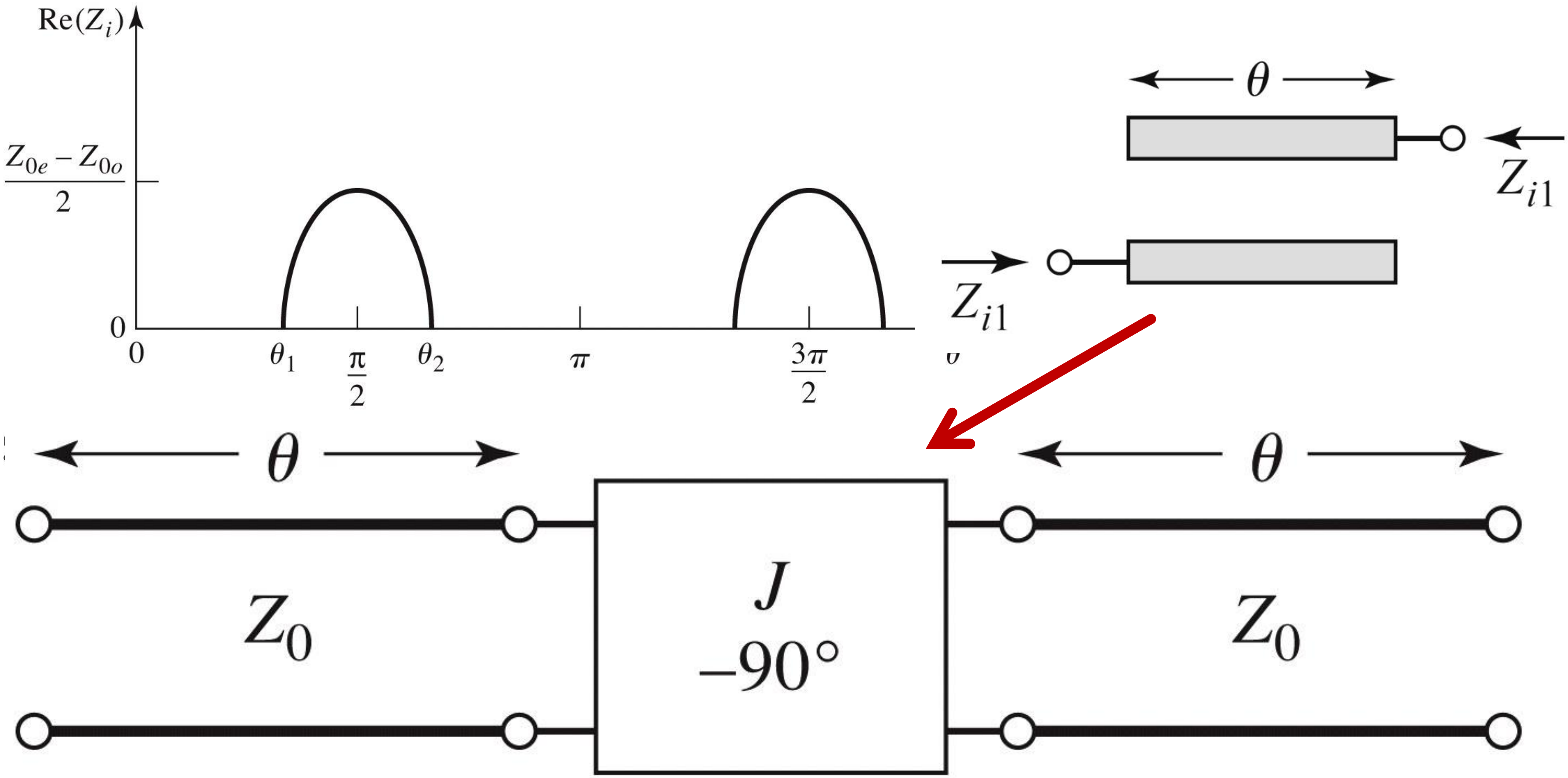
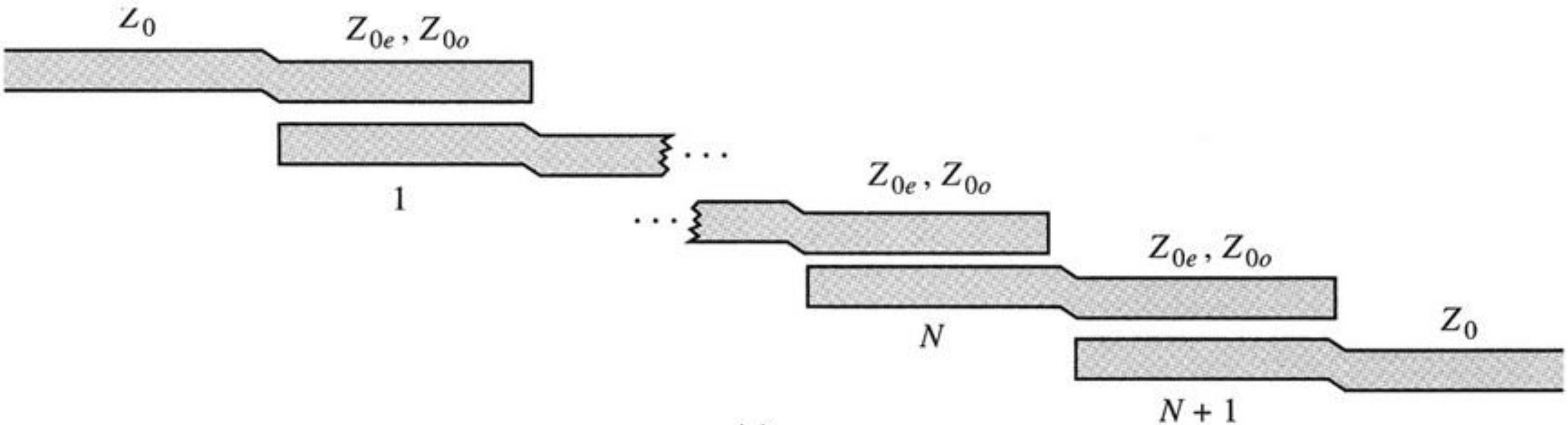


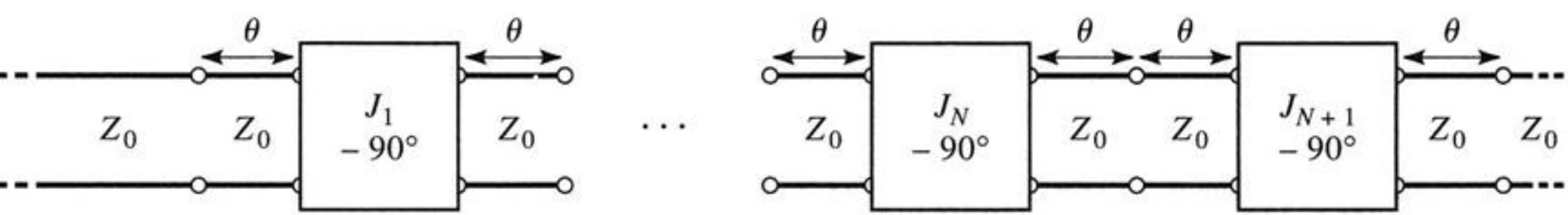
Figure 8.44

Coupled Line Filters

- We get a **N^th^** order filter with **N+1** parallel coupled line section



(a)



Example

- Similar to a project assignment
- Follows the amplifier designed as in L10
- **4**th order bandpass filter, fo = 5GHz, fractional bandwidth of the passband 10 %
- 0.5dB equal-ripple table for g_n followed by filter design formulas

n	g	ZoJn	Zoe	Zoo
1	1.6703	0.306664	70.04	39.37
2	1.1926	0.111295	56.18	45.05
3	2.3661	0.09351	55.11	45.76
4	0.8419	0.111294	56.18	45.05
5	1.9841	0.306653	70.03	39.37

Continue

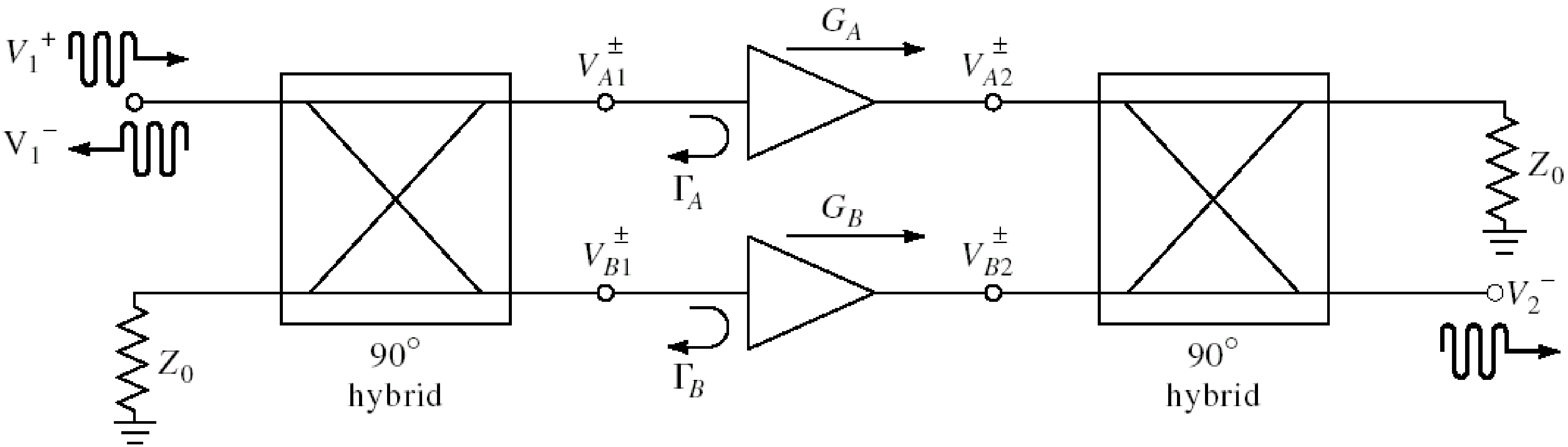
Microwave Amplifiers

Broadband amplifiers

Broadband/Wideband amplifiers

- Achieved by some design techniques (only at the expense of gain, complexity)
  1. Compensated matching networks
  2. Resistive matching networks
  3. Negative feedback
  4. Balanced amplifiers
  5. Distributed amplifiers
  6. Differential amplifiers

Balanced amplifiers



- two identical amplifiers with two hybrid couplers 3 dB / 90° to cancel input and output reflections

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

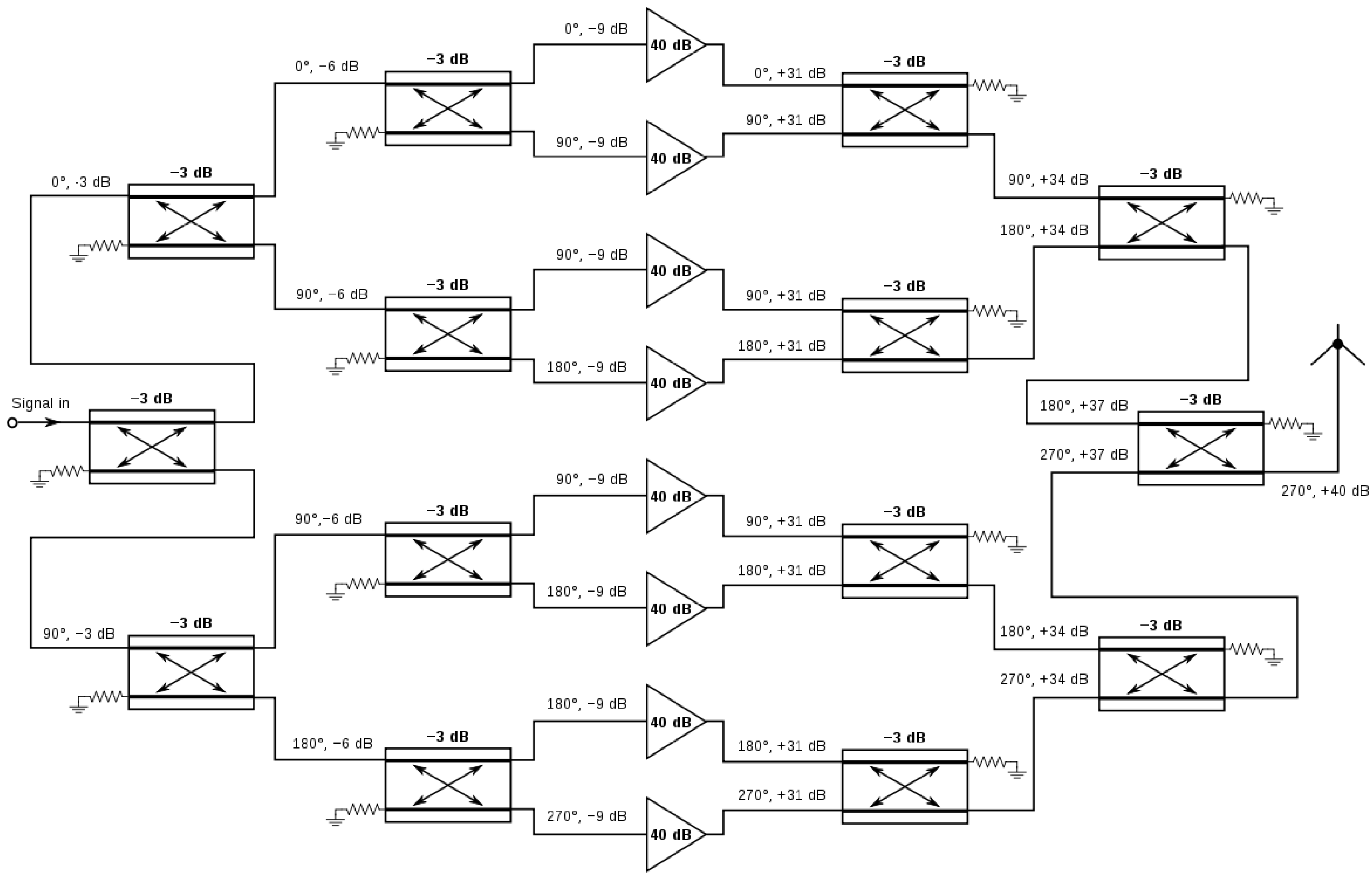
$$S_{11} = \frac{1}{2} \cdot (\Gamma_A - \Gamma_B)$$

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

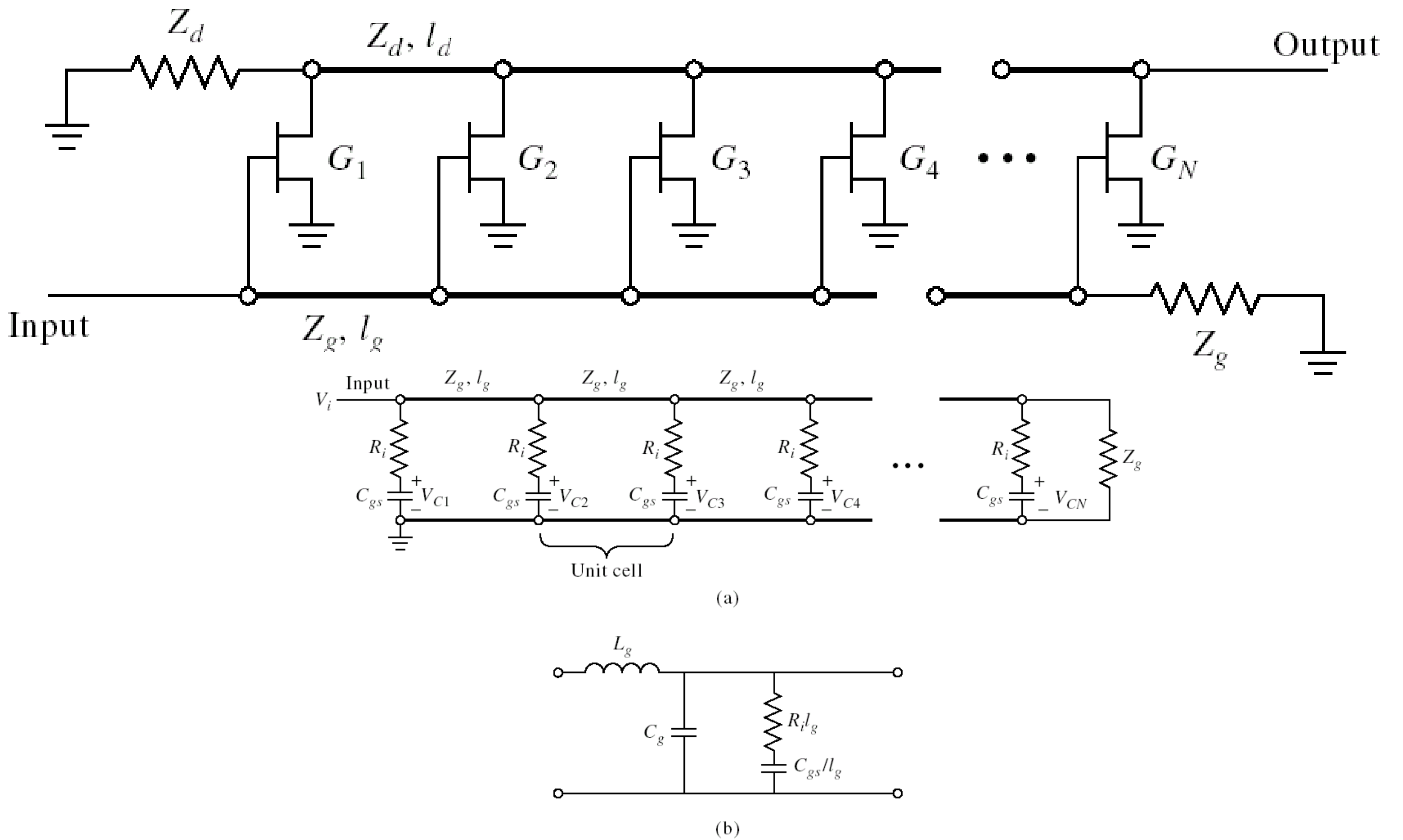
$$S_{21}\big|_{A=B} = -j \cdot G$$

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

Balanced amplifiers



Distributed amplifiers



Distributed amplifiers

- the phase delays on the gate (input) and drain (output) lines are synchronized

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

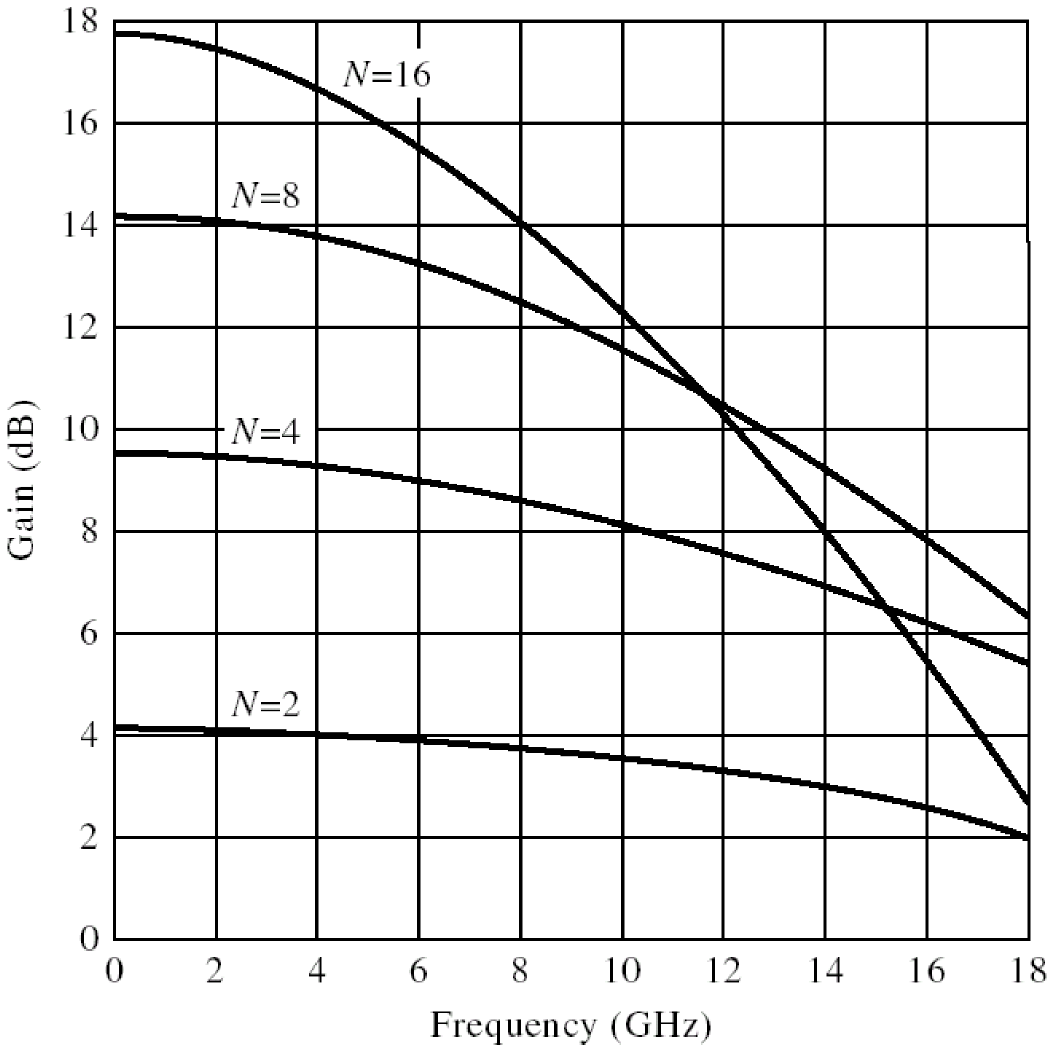
- Power gain

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

- Lossless power gain

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

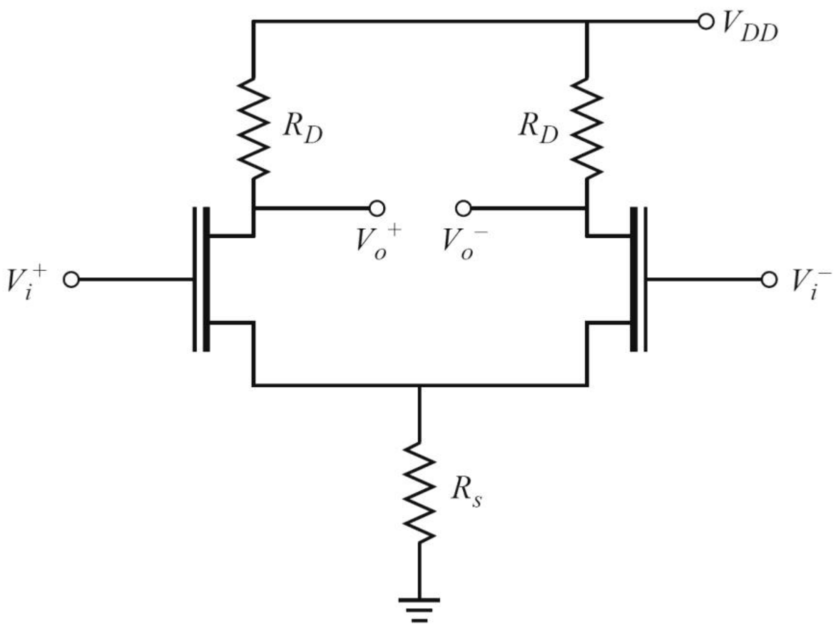
Distributed amplifiers



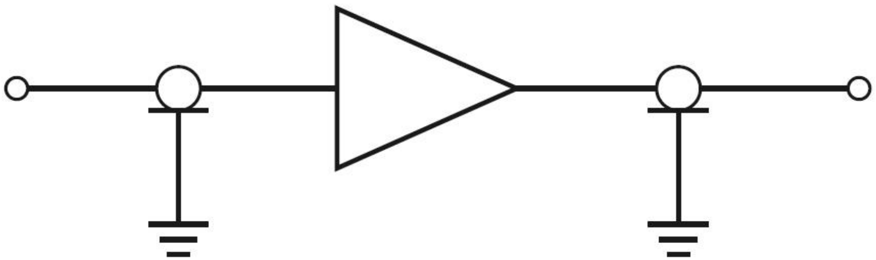
$$N_{opt} = \frac{\ln(\alpha_g \cdot l_g) - \ln(\alpha_d \cdot l_d)}{\alpha_g \cdot l_g - \alpha_d \cdot l_d}$$

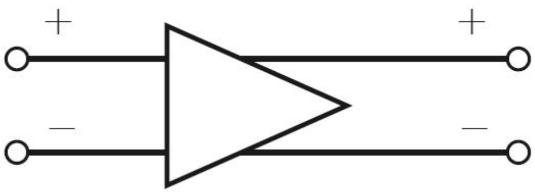
Differential amplifiers

- In differential mode the input capacitances of the two transistors are connected in series
- Unity gain frequency is doubled

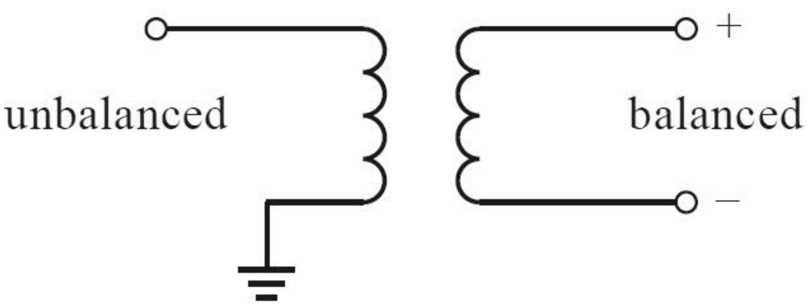


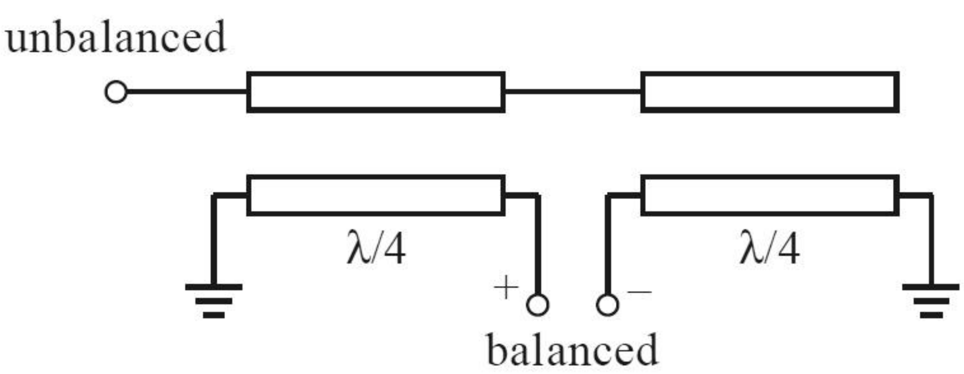
Differential amplifiers





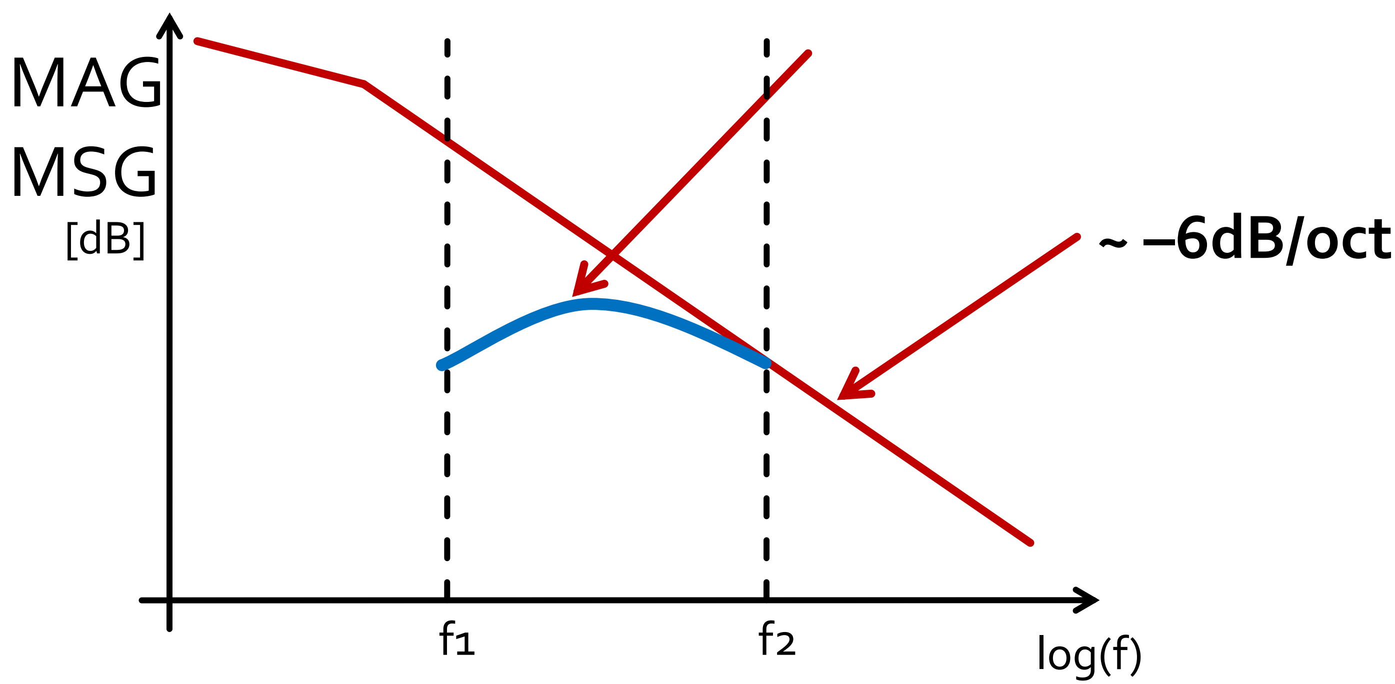
- We use circuits to transition from an unbalanced signal to a balanced signal (or vice versa)
  - hybrid couplers 3dB / 180°
  - "balun" (balanced - unbalanced)





Compensated matching networks

- Control the design of the matching networks at more (at least 2) frequencies and impose the same gain

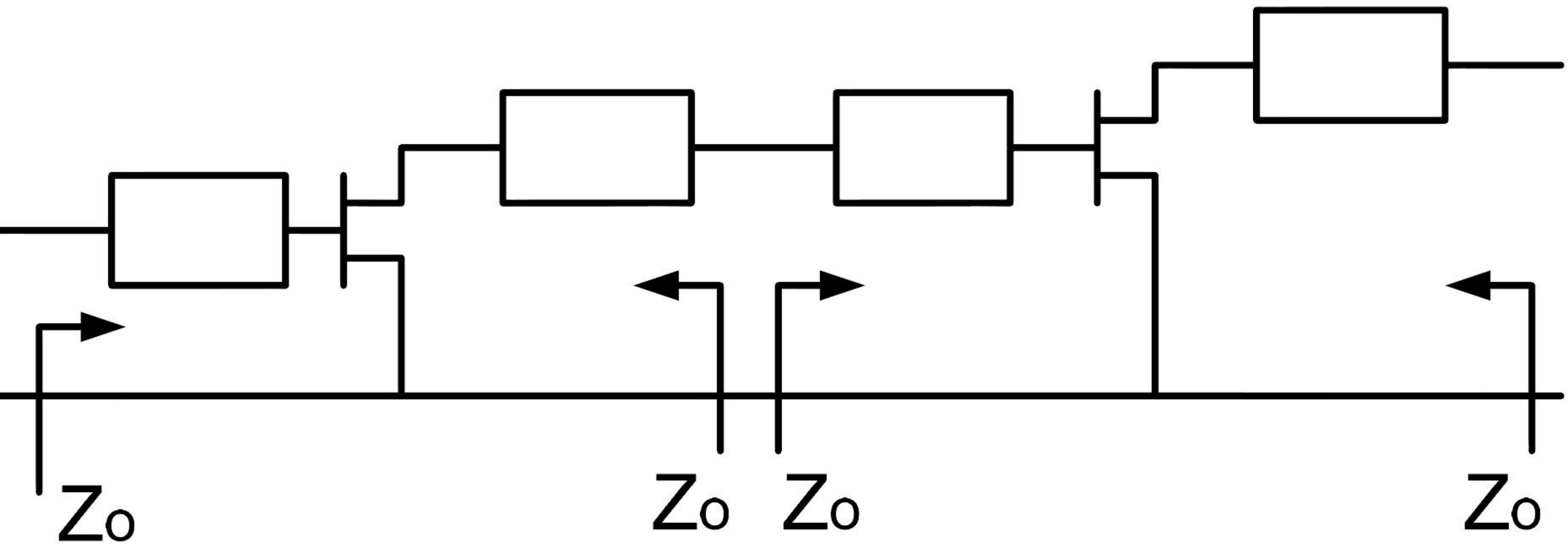


Microwave Amplifiers

Multistage Amplifier Design

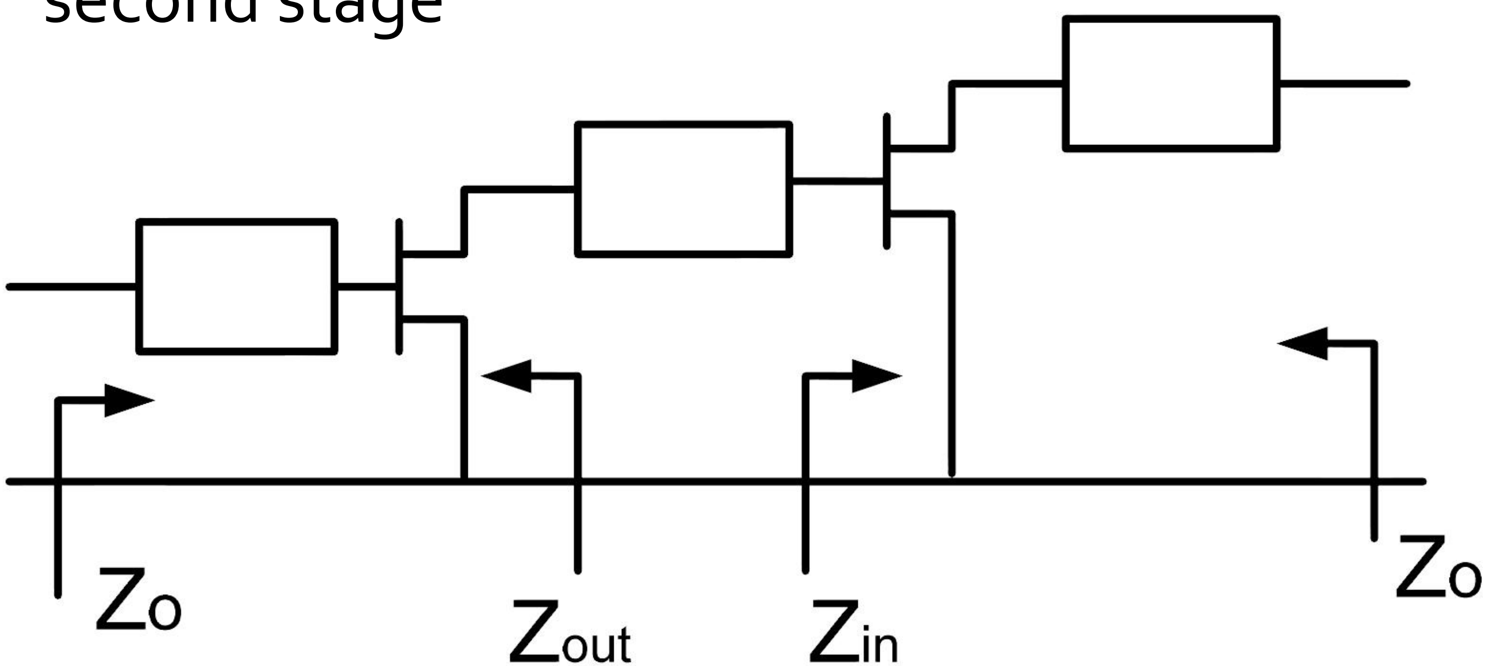
Multistage amplifiers

- 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

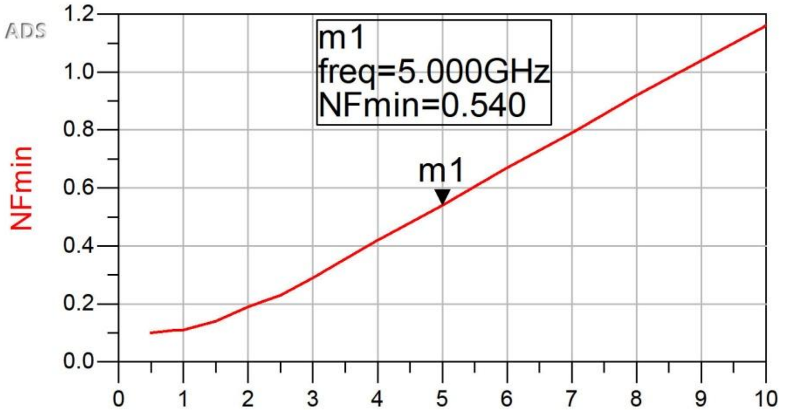


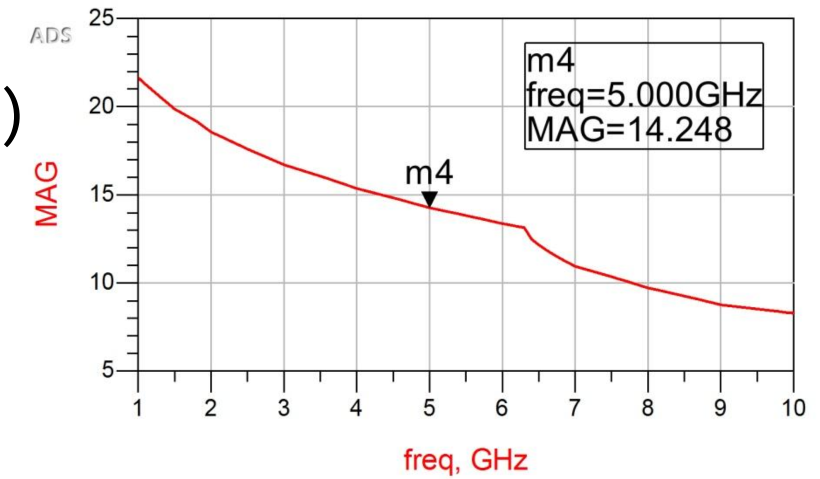
Example multistage LNA

- Similar to the project assignment
- 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] !**)
  - $\Gamma_{opt} = 0.45\angle 174°$
  - $r_n = 0.03$





Example, LNA @ 5 GHz

- ATF-34143 **at Vds=3V Id=20mA**.
- @5GHz
  - $S_{11} = 0.64\angle 139°$
  - $S_{12} = 0.119\angle -21°$
  - $S_{21} = 3.165\angle 16°$
  - $S_{22} = 0.22\angle 146°$
  - $F_{min} = 0.54$ (**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
```

Multistage amplifiers

- If we need more power gain than only one transistor can supply
  - design target 20dB
  - MAG @5GHz = 14.248 dB < 20dB
- We use Friis formula to separate the target:
  - Power gain
  - Noise
- on 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
- Separation of the design parameters on the 2 amplification 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 !**

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

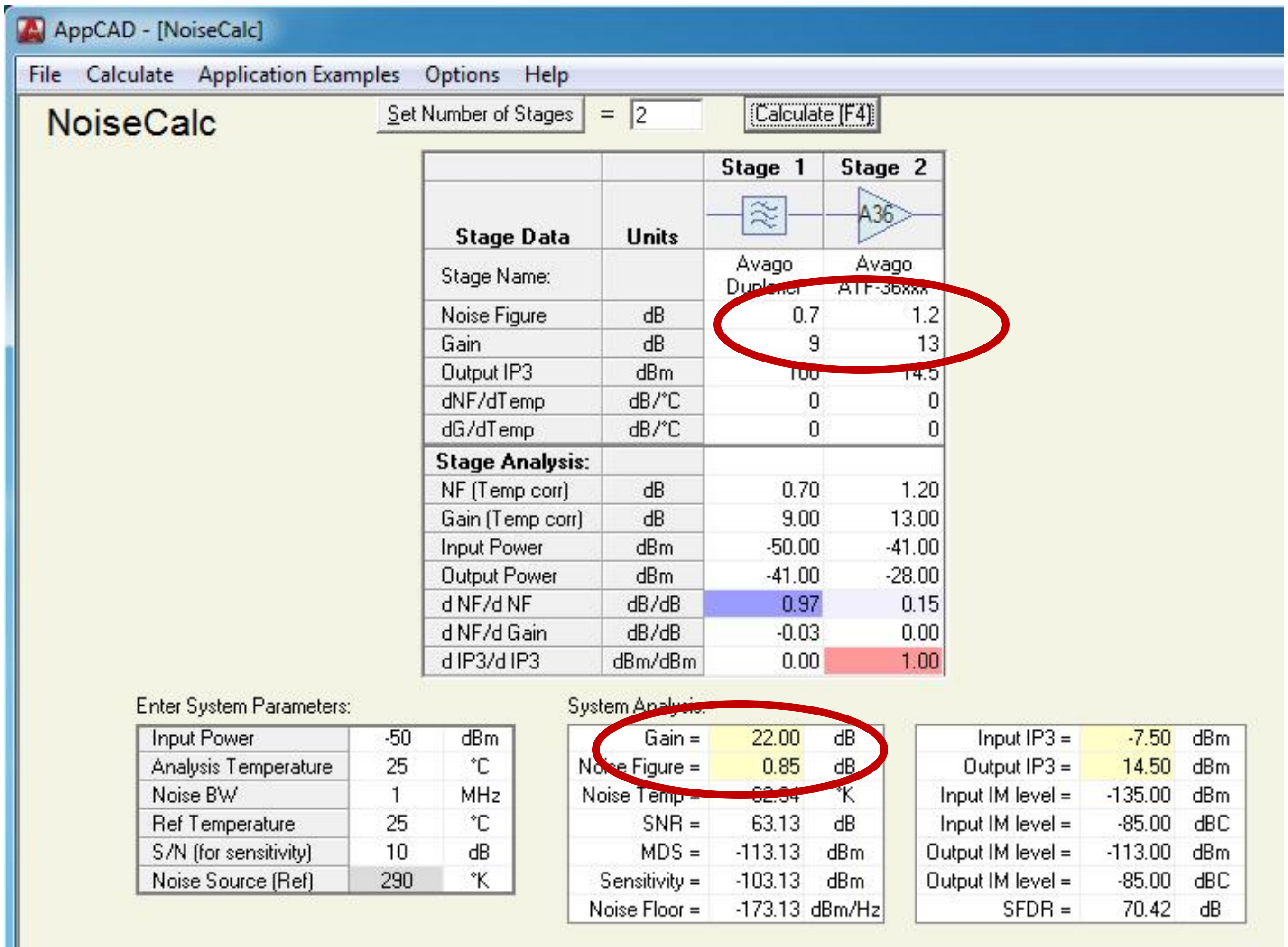
$$F_2 = 10^{\frac{F_2[dB]}{10}} = 10^{0.12} = 1.318 \qquad G_2 = 10^{\frac{G_2[dB]}{10}} = 10^{1.3} = 19.953$$

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

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

Friis Formula (noise)

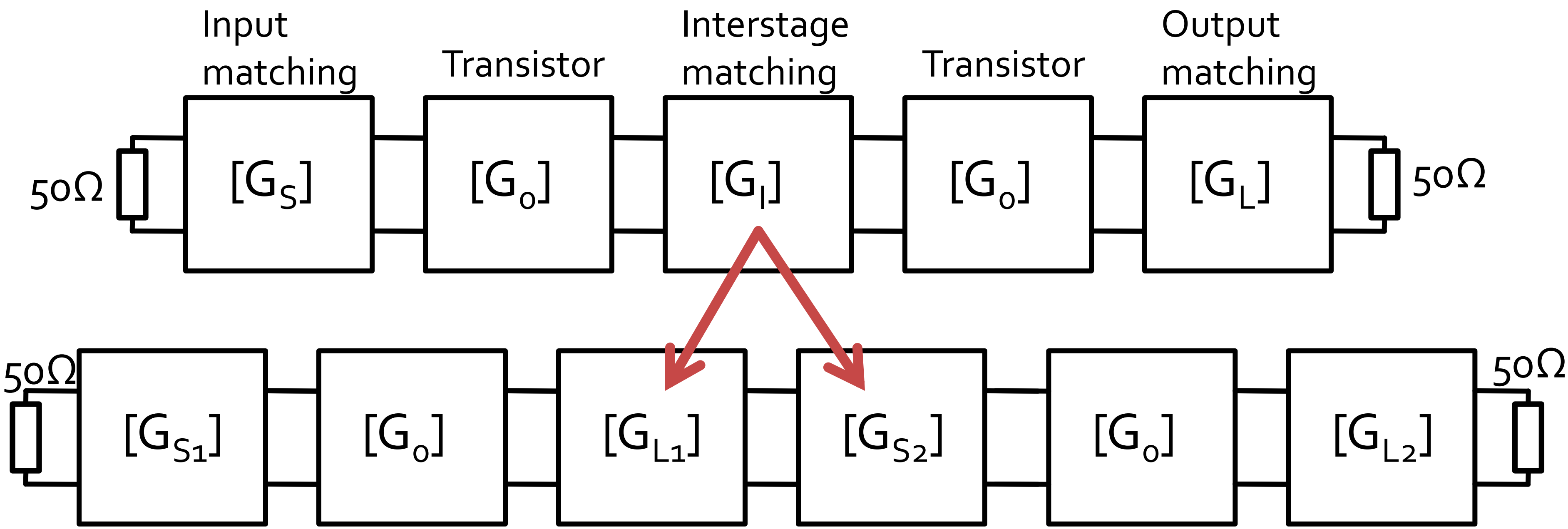
- Avago/Broadcom AppCAD



Multistage amplifier design

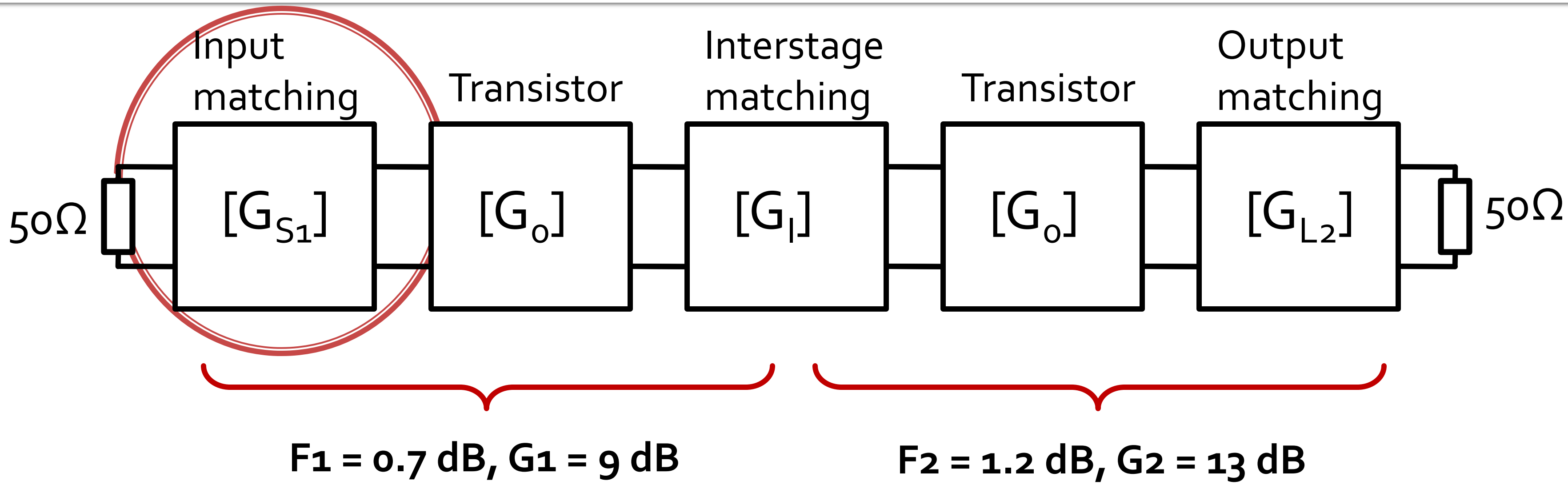
- Separation of the design parameters on the 2 amplification stages (Estimated!)
  - input stage: $F_1 = 0.7$ dB, $G_1 = 9$ dB
  - output stage: $F_2 = 1.2$ dB, $G_2 = 13$ dB
  - total: $F = 0.85$ dB, $G = 22$ dB
- Meets design specifications (with design margin)
- We can reuse some of the results in the single stage LNA design (Lecture 10)
  - input matching can be used for the input of the first stage – very low noise, good enough power gain
  - output matching was designed for maximum gain, can be used for the output of the second stage
  - input and output matching were designed for 50Ω source and load, similar to current conditions

Multistage amplifier design



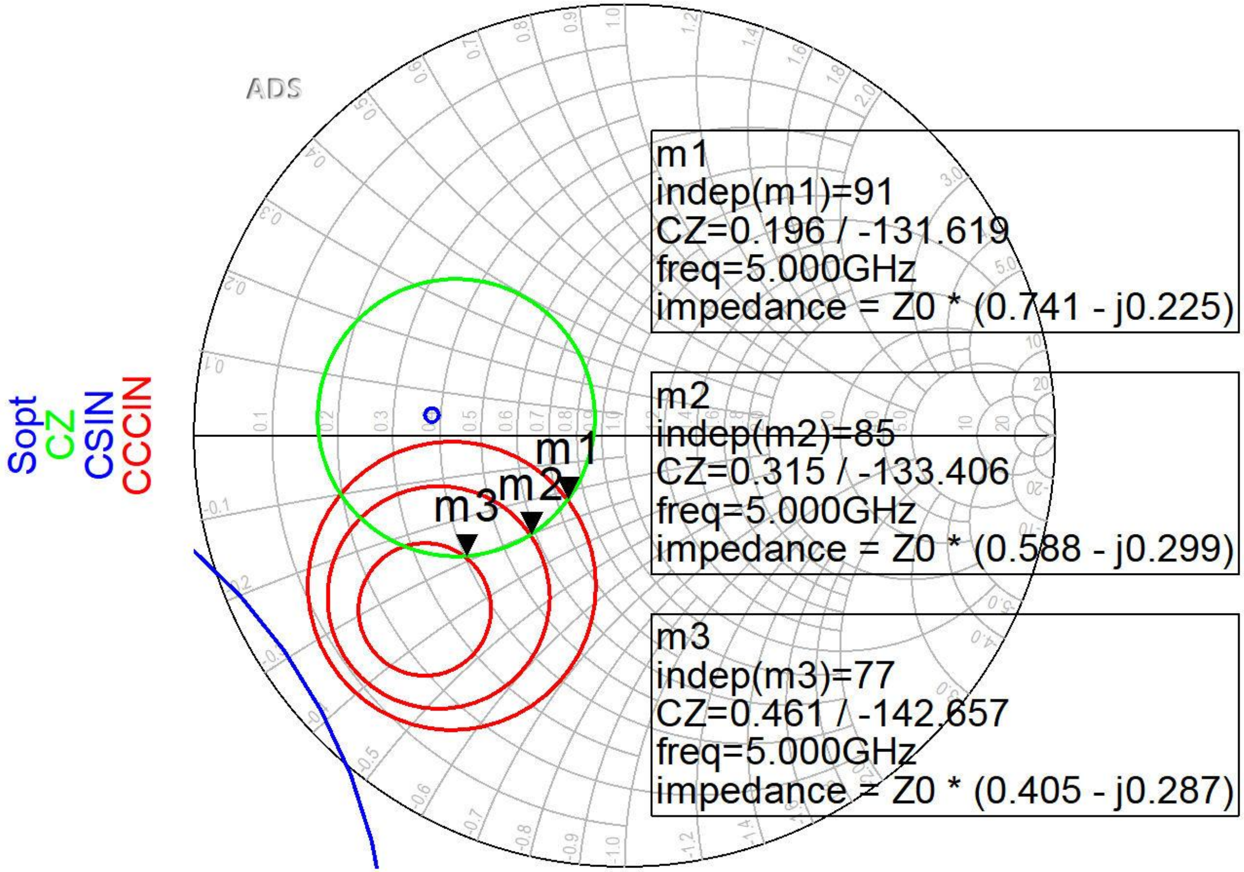
- Gain computation
  - Interstage matching can supplement the gain for both amplifier stages
  - The design for input and output matching must be achieved on a single transistor schematic (recommended: easier)

Input matching stage 1 (S1)



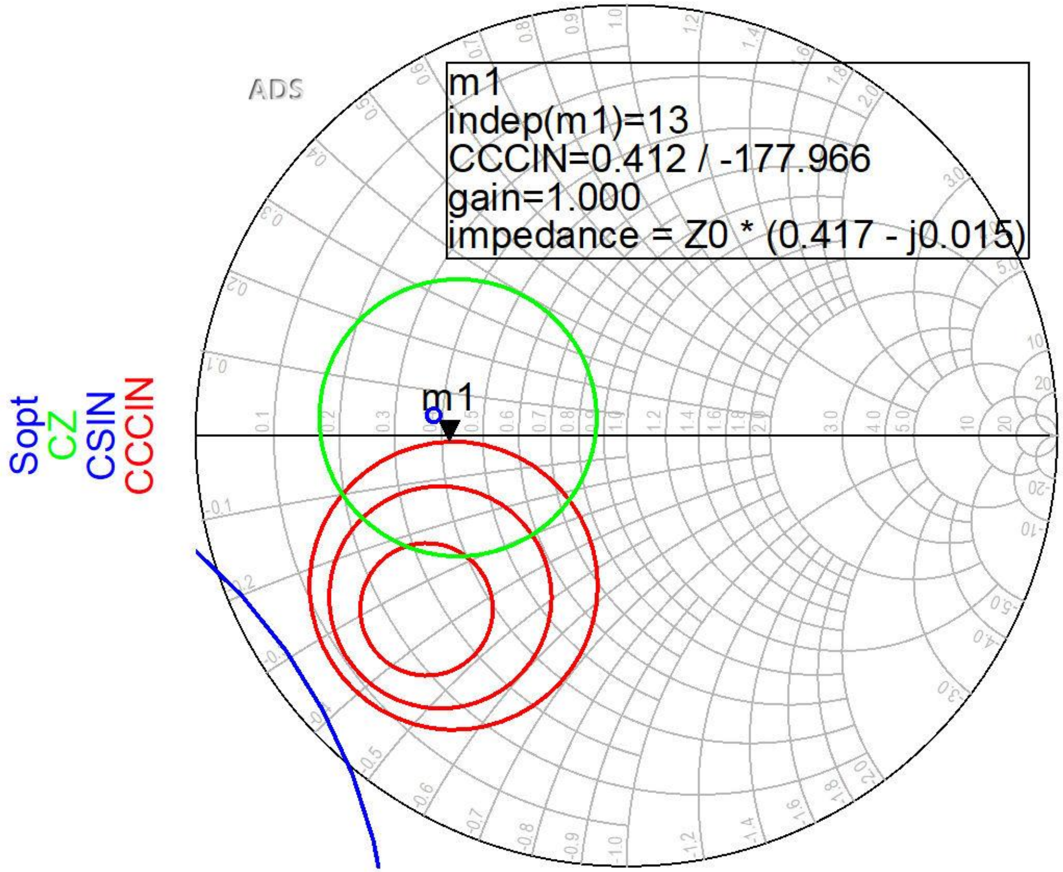
- We favor optimization for **noise** (low/minimum)
- Also considered
  - Power gain (can be lower, but not too much)
  - Bandwidth (through Q, quality factor)
  - Stability

Input matching stage 1 (S1)



- 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 stage 1 (S1)



- 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 favor **better** (smaller) **NF**

Input matching stage 1 (S1)

- G_{S_1}: Position m1 in complex plane, **1dB**

$$\Gamma_S = 0.412\angle -178^\circ \qquad |\Gamma_S| = 0.412; \quad \varphi = -178^\circ$$

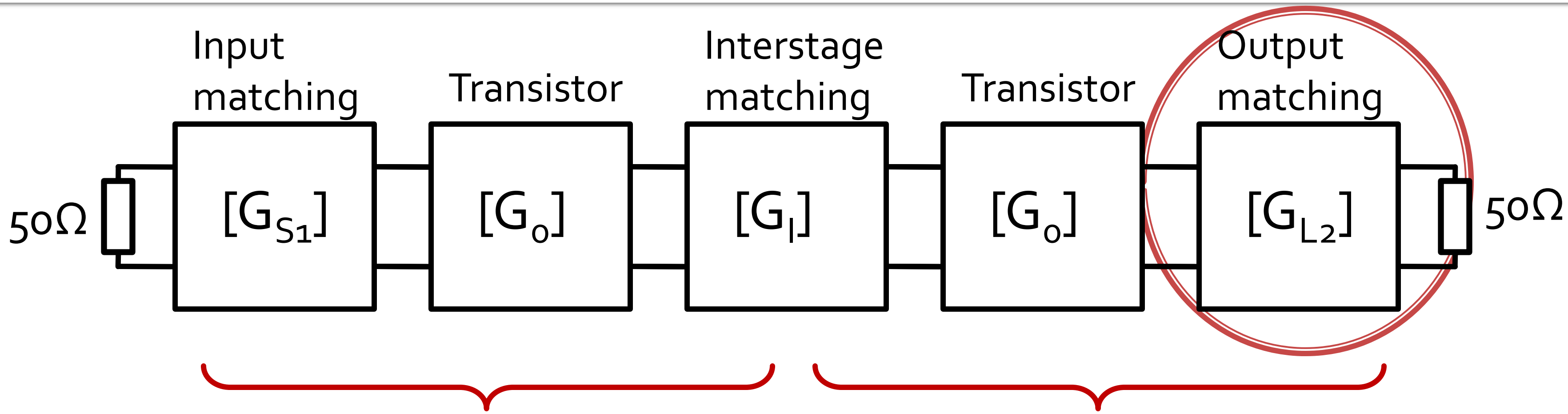
$$\cos(\varphi + 2\theta) = -|\Gamma_S| \qquad \mathrm{Im}[y_S(\theta)] = \frac{\mp 2\cdot|\Gamma_S|}{\sqrt{1-|\Gamma_S|^2}}$$

$$\cos(\varphi + 2\theta) = -0.412 \Rightarrow \qquad (\varphi + 2\theta) = \pm 114.33^\circ$$

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

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

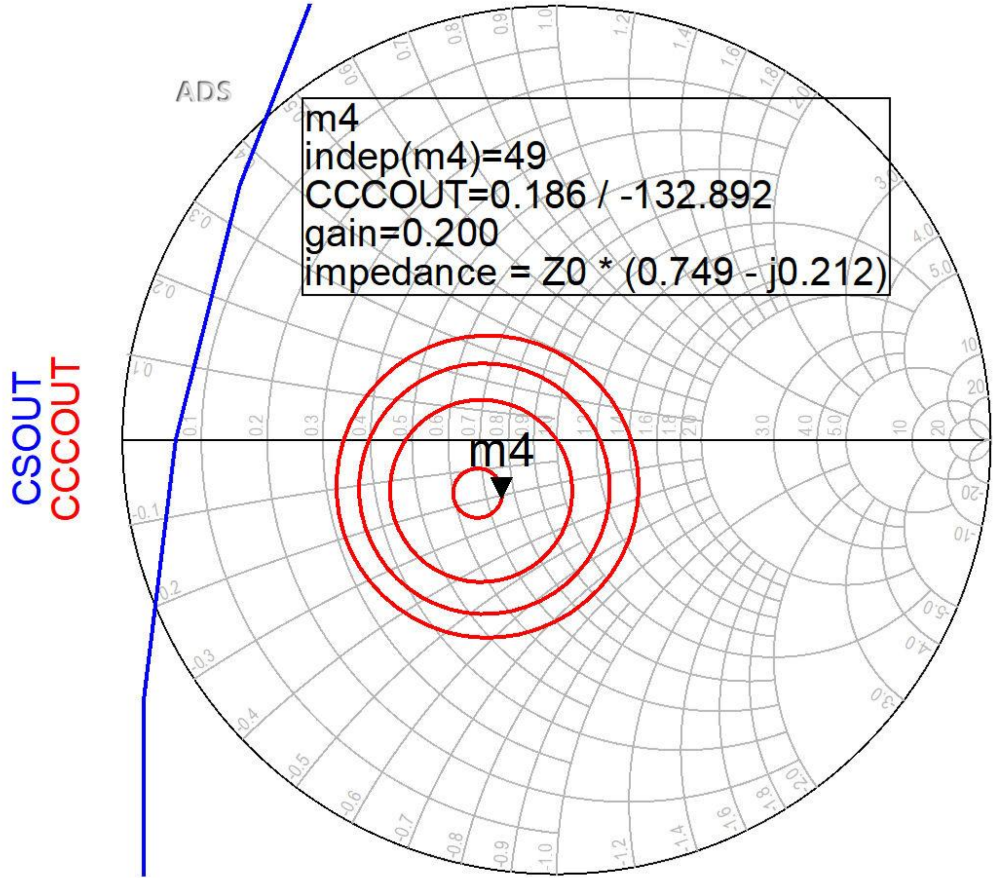
Output matching stage 2 (L2)



F1 = 0.7 dB, G1 = 9 dB **F2 = 1.2 dB, G2 = 13 dB**

- We favor optimization for **gain** (high/maximum)
- Also considered
  - Bandwidth (through Q, quality factor)
  - Stability
- noise is **not** an issue, output matching doesn't influence noise factor

Output matching stage 2 (L2)



- 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 stage 2 (L2)

- **G_{L_2}**: Position m4 in complex plane, **0.2dB**

$$\Gamma_L = 0.186\angle -132.9^\circ \qquad |\Gamma_L| = 0.186; \quad \varphi = -132.9^\circ$$

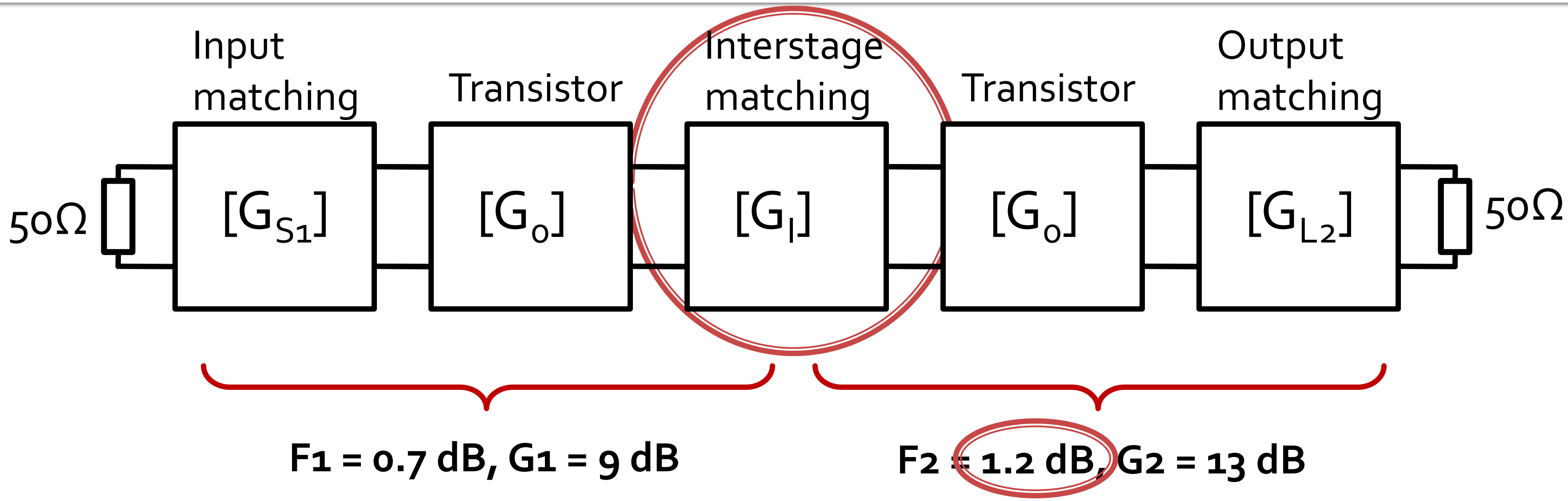
$$\cos(\varphi + 2\theta) = -|\Gamma_L| \qquad \mathrm{Im}[y_L(\theta)] = \frac{-2\cdot|\Gamma_L|}{\sqrt{1-|\Gamma_L|^2}} = -0.379$$

$$\cos(\varphi + 2\theta) = -0.186 \Rightarrow \qquad (\varphi + 2\theta) = \pm 100.72^\circ$$

$$\theta_{sp} = \tan^{-1}(\mathrm{Im}[y_L(\theta)]) = \tan^{-1}\left(\frac{\mp 2\cdot|\Gamma_L|}{\sqrt{1-|\Gamma_L|^2}}\right)$$

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

Interstage matching (I)



- We take into account **gain** (high) but also **noise**
- Also considered
  - Bandwidth (through Q, quality factor)
  - Stability
- We influence the noise factor of the second stage, the noise must be considered but with less restrictive conditions (Friis shows that higher noise is acceptable).

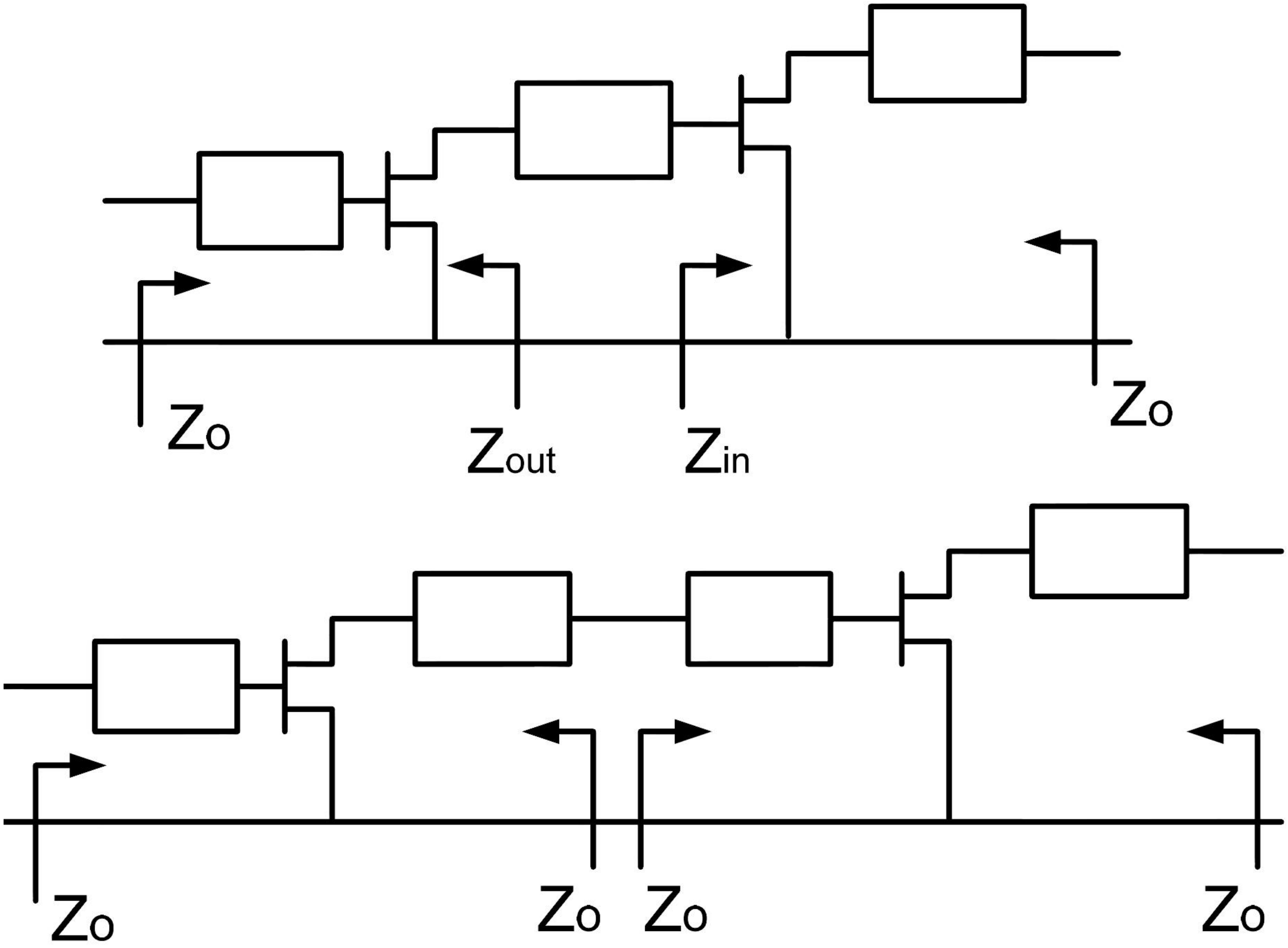
Multistage amplifier

- Power gain

$$G_T[dB] = G_{S1}[dB] + G_0[dB] + G_I[dB] + G_0[dB] + G_{L2}[dB]$$
$$G_0 = |S_{21}|^2 = 10.017 = 10.007\ dB$$
$$G_T[dB] = 1\ dB + 10\ dB + G_I[dB] + 10\ dB + 0.2\ dB$$
$$G_T[dB] = 21.2\ dB + G_I[dB]$$

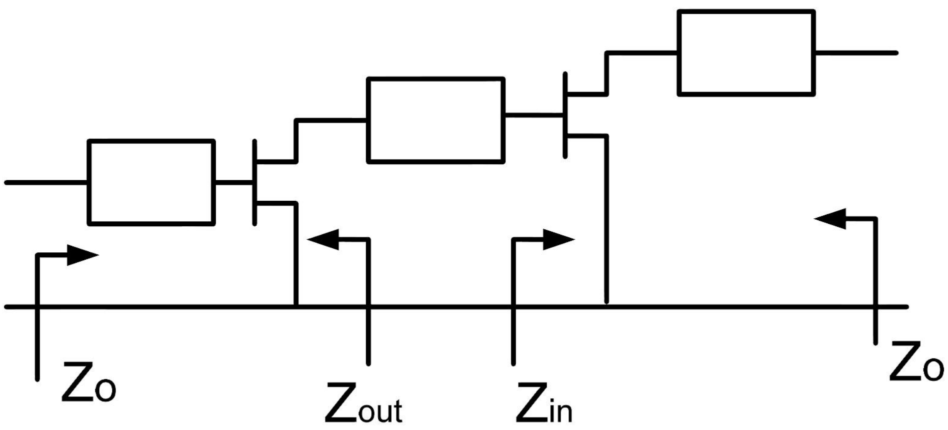
- Interstage match design must provide at least 0.8dB gain to meet specifications, by better match for the output of the first transistor and for the input of the second transistor

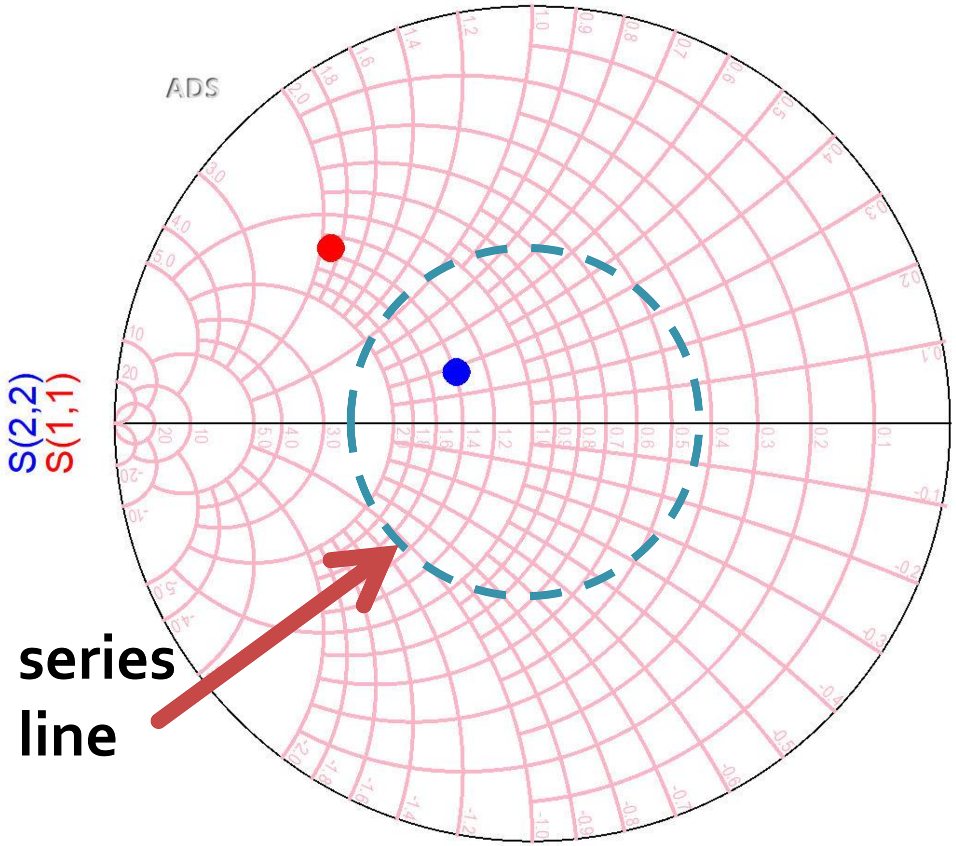
Interstage matching 1/2



Interstage matching 1

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



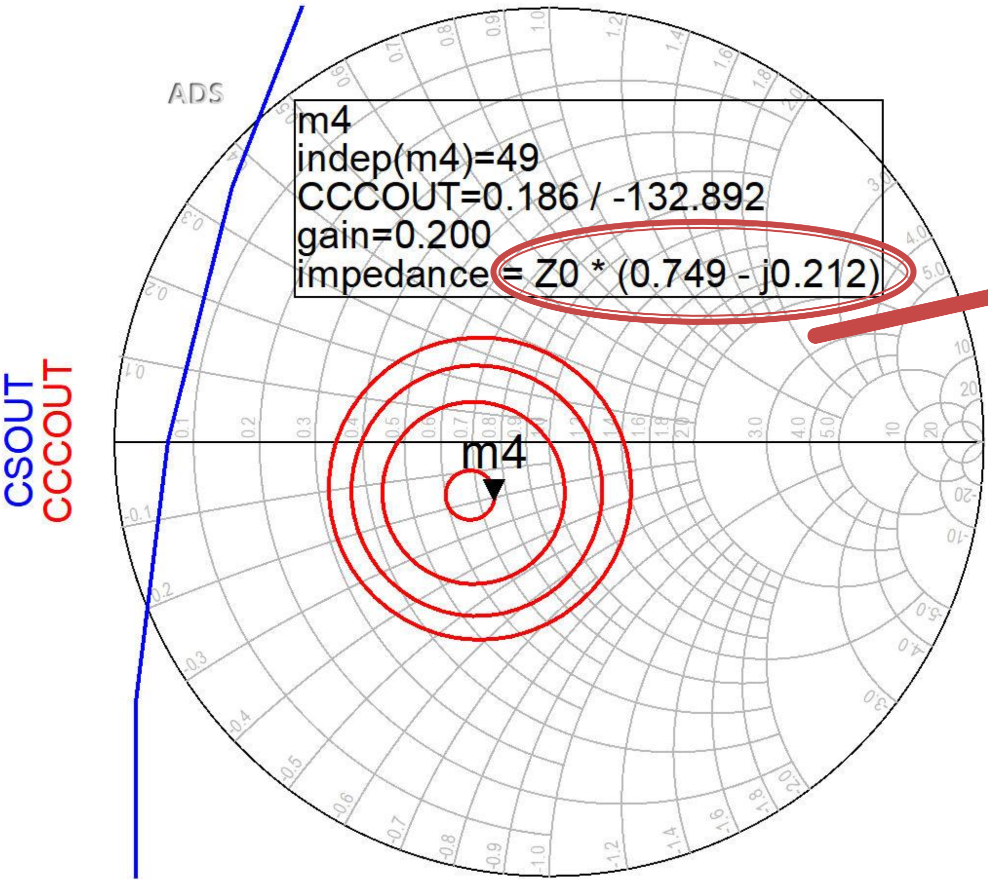


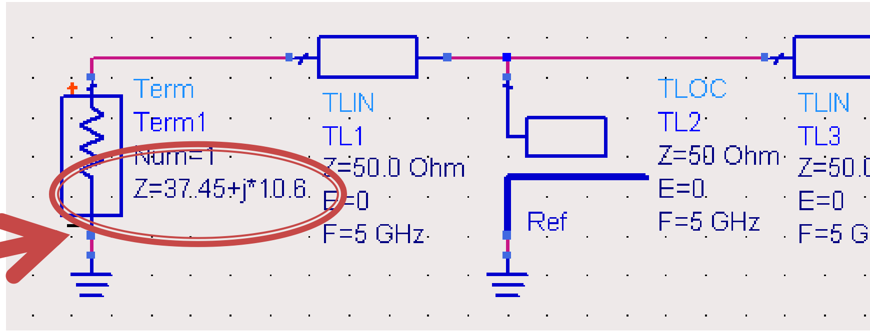
Interstage matching 1

- Can be designed in two ways:
  - 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 offer control over the noise introduced by the second stage

Interstage matching 1

- Starting point – complex conjugate





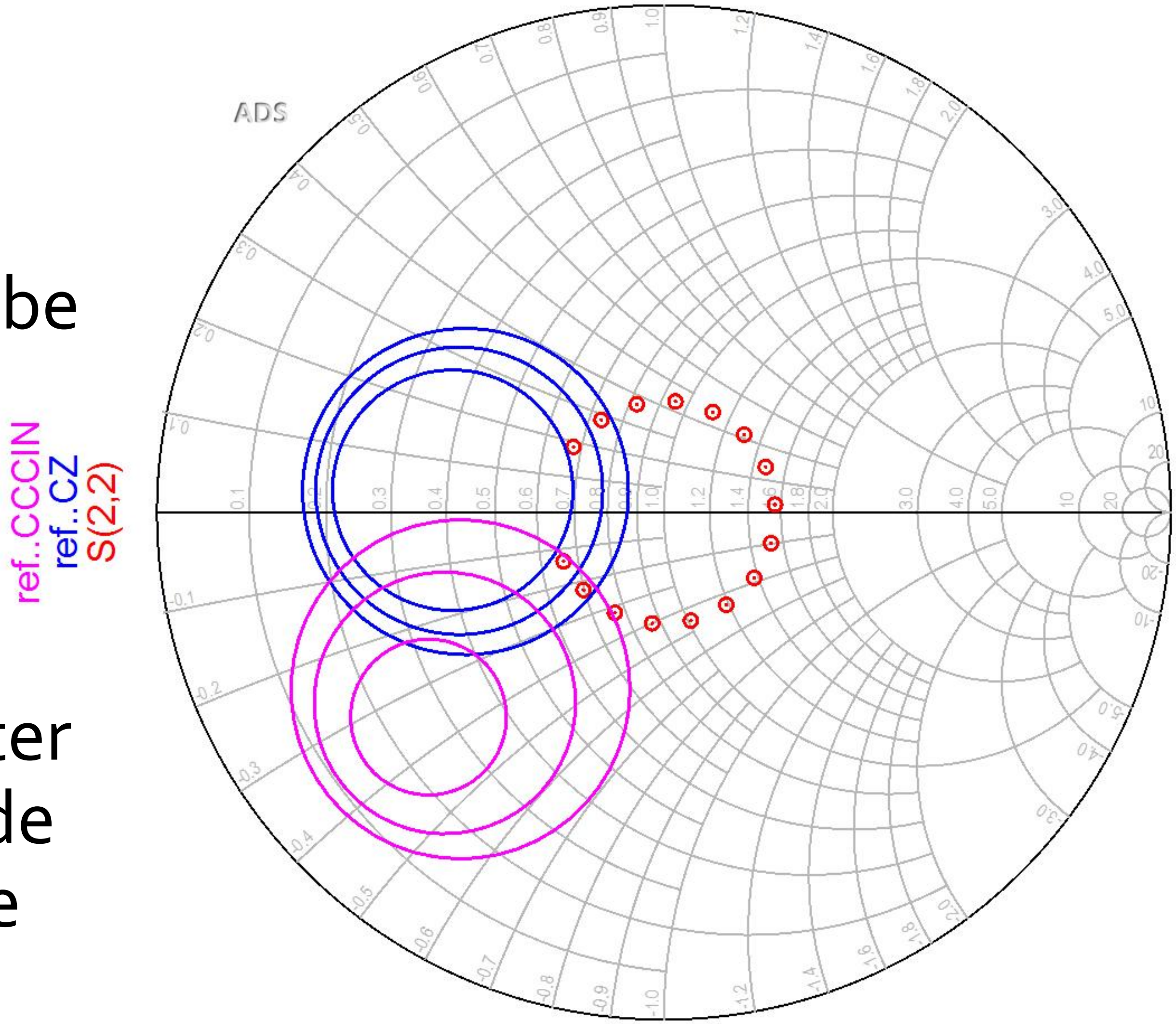
$$Z = 50\Omega \cdot (0.749 - j \cdot 0.212)$$

$$Z = 37.45\Omega - j \cdot 10.6\Omega$$

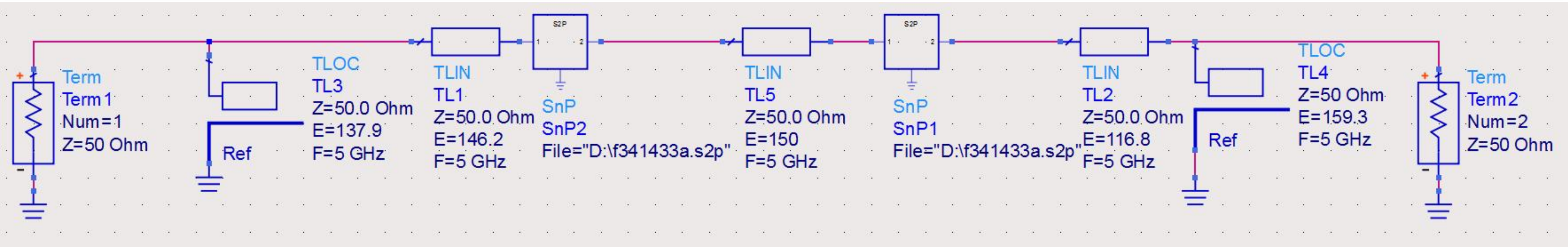
$$Z^* = 37.45\Omega + j \cdot 10.6\Omega$$

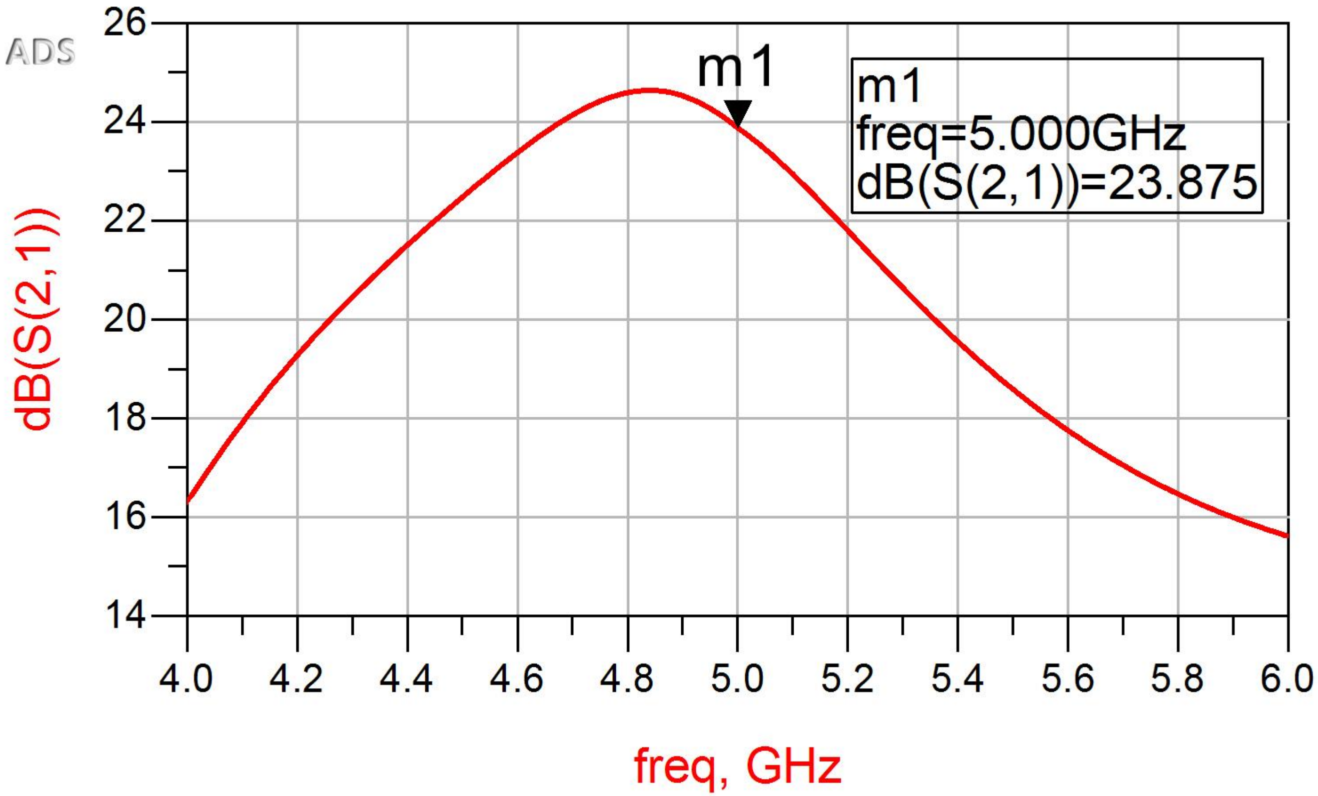
Interstage matching 1

- A **single** transmission line allows reaching a point that cannot be optimized
  - $G_{L1} = 0.2\text{dB}$
  - $G_{S2} = 1\text{ dB}$
  - $F_2 = 0.7\text{ dB}$
- Only one parameter is available for wide band performance tuning

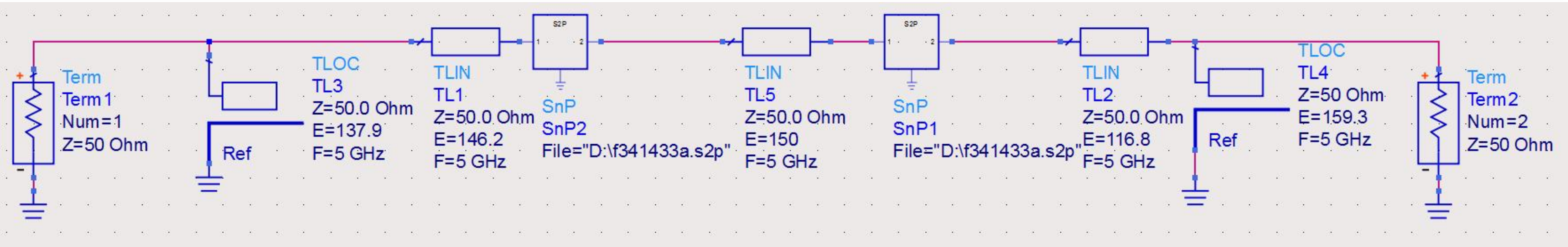


ADS

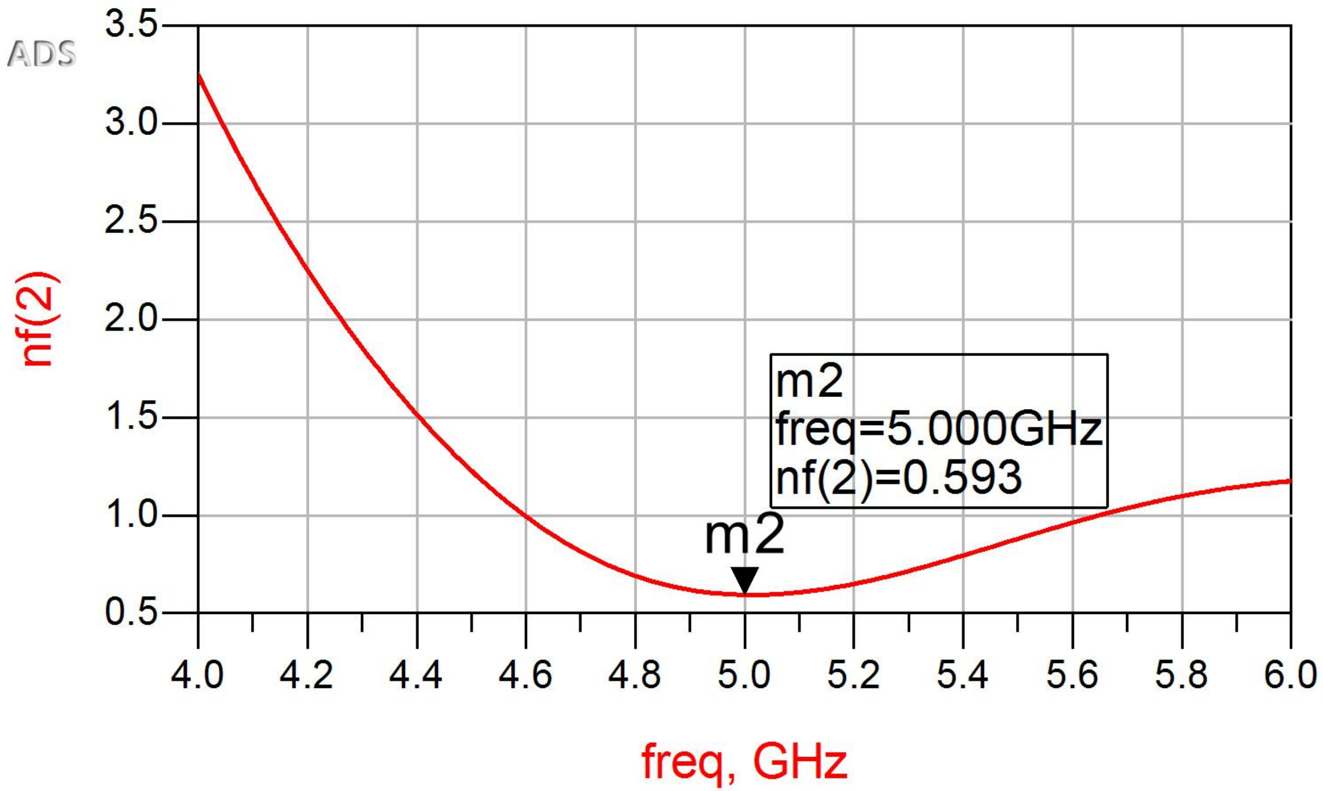

Term
Term1
Num=1
Z=50 Ohm
Ref
TLOC
TL3
Z=50.0 Ohm
E=137.9
F=5 GHz
TLIN
TL1
Z=50.0 Ohm
E=146.2
F=5 GHz
SnP
SnP2
File="D:\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=150
F=5 GHz
SnP
SnP1
File="D:\f341433a.s2p"
TLIN
TL2
Z=50.0 Ohm
E=116.8
F=5 GHz
Ref
TLOC
TL4
Z=50 Ohm
E=159.3
F=5 GHz
Term
Term2
Num=2
Z=50 Ohm


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



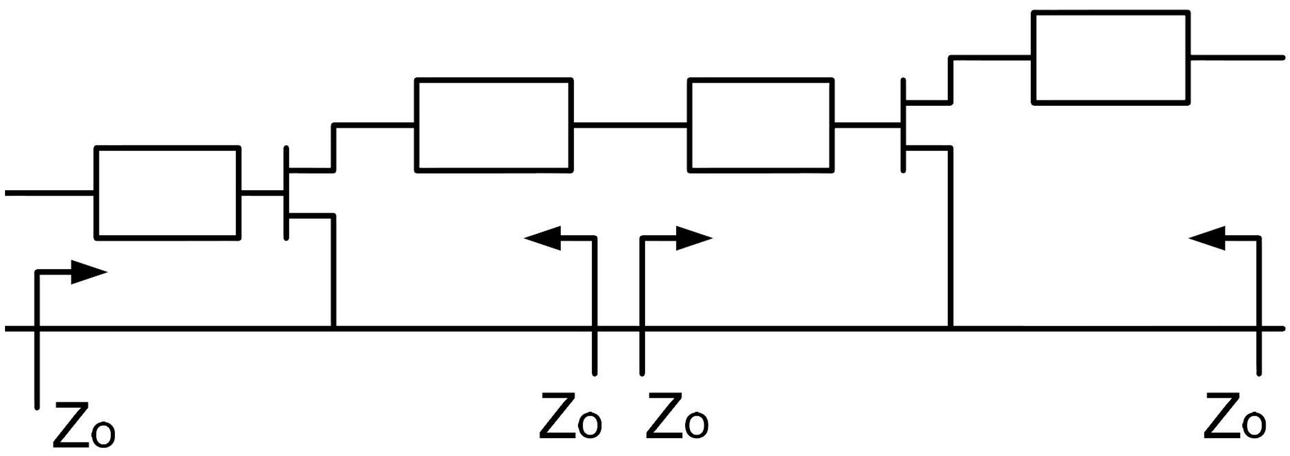
Term
Term1
Num=1
Z=50 Ohm
Ref
TLOC
TL3
Z=50.0 Ohm
E=137.9
F=5 GHz
TLIN
TL1
Z=50.0 Ohm
E=146.2
F=5 GHz
SnP
SnP2
File="D:\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=150
F=5 GHz
SnP
SnP1
File="D:\f341433a.s2p"
TLIN
TL2
Z=50.0 Ohm
E=116.8
F=5 GHz
Ref
TLOC
TL4
Z=50 Ohm
E=159.3
F=5 GHz
Term
Term2
Num=2
Z=50 Ohm



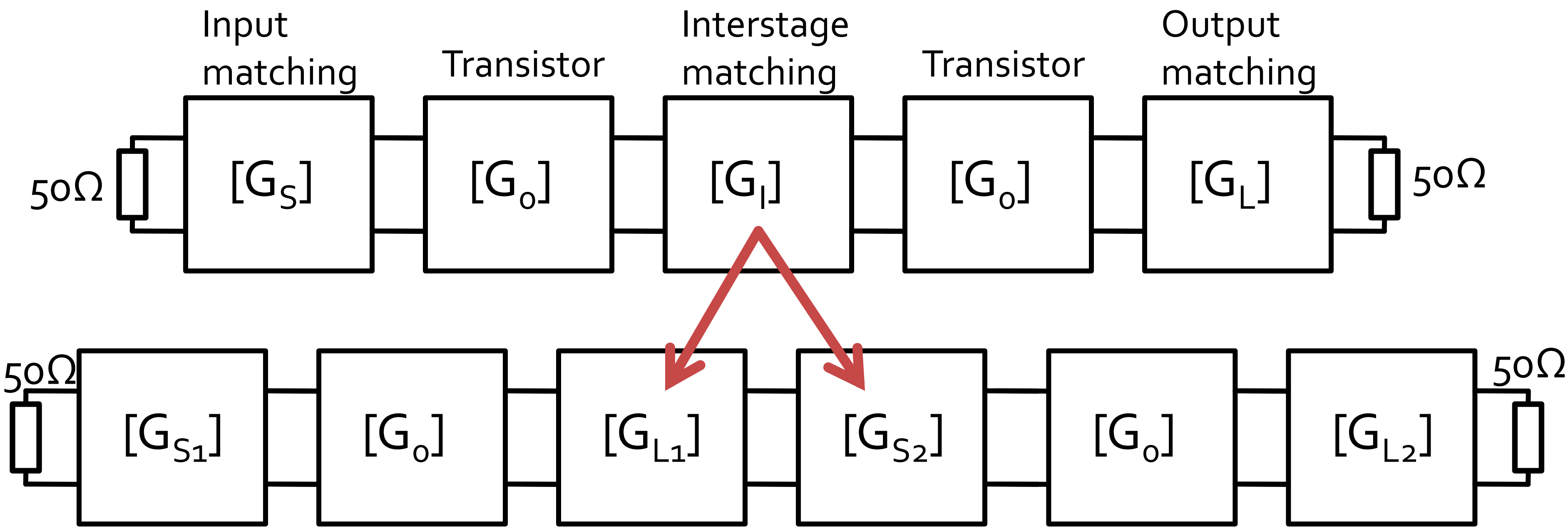
ADS
nf(2)
3.5
3.0
2.5
2.0
1.5
1.0
0.5
m2
freq=5.000GHz
nf(2)=0.593
m2
4.0 4.2 4.4 4.6 4.8 5.0 5.2 5.4 5.6 5.8 6.0
freq, GHz

Interstage matching 2

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



Interstage matching 2



- Instead of a single match design we have to design two matching networks
- However both matching networks are anchored to a fixed point (50Ω, Γ=0) so we can use design **formulas** (Impedance Matching with Stubs)
- Also, due to the presence of multiple networks, we can target **precise** positions (reflection coefficients) on both stages

Multistage amplifier

- Power gain

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

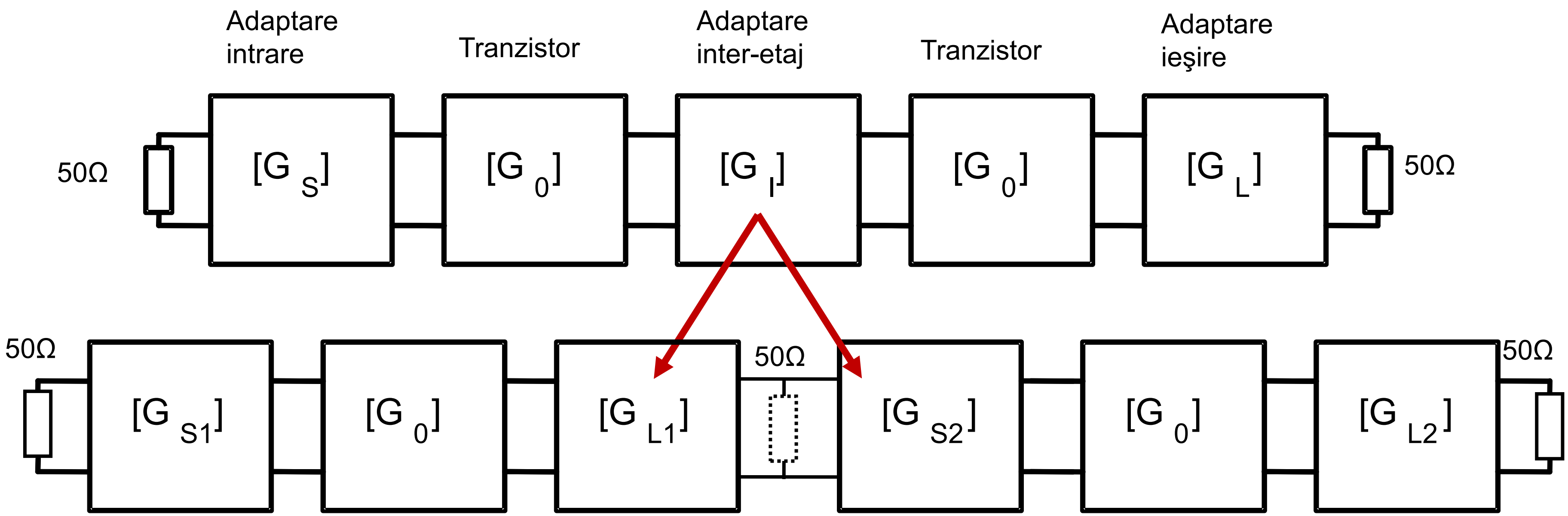
$$G_T[dB] = 1\ dB + 10\ dB + G_{L1}[dB] + G_{S2}[dB] + 10\ dB + 0.2\ dB$$

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

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

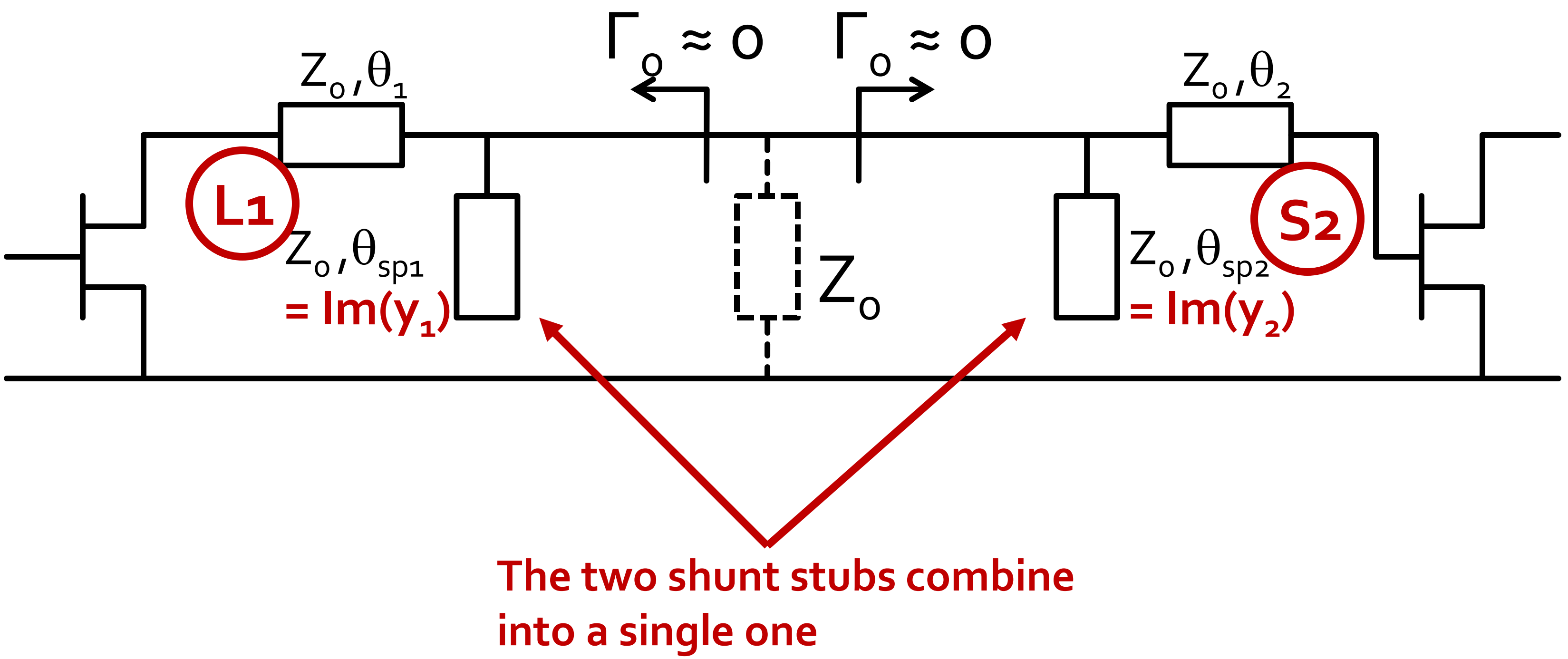
Interstage matching 2

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



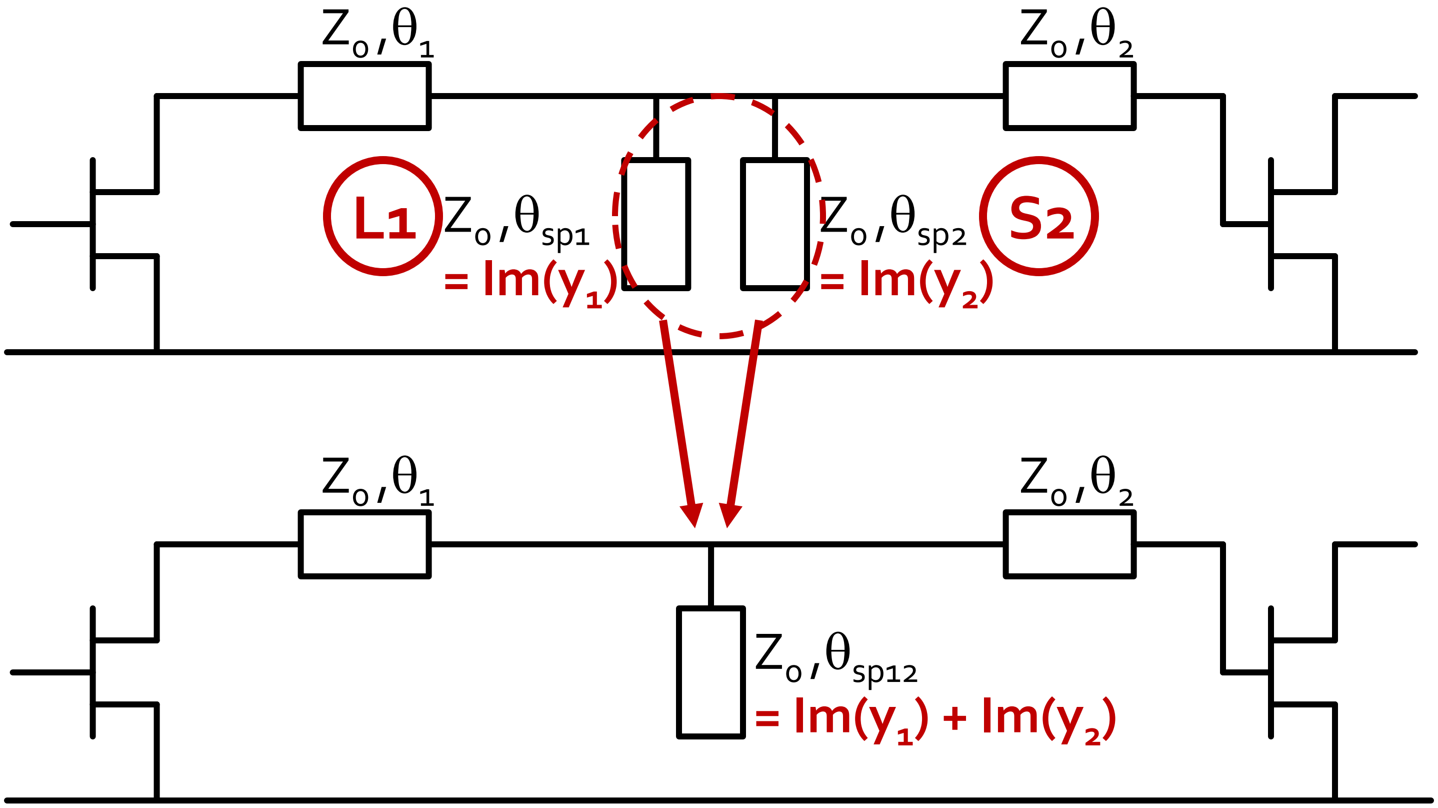
Interstage matching 2

- One of the stages **creates** through its matching network a reflection coefficient Γ=0 towards which the other stage is matched



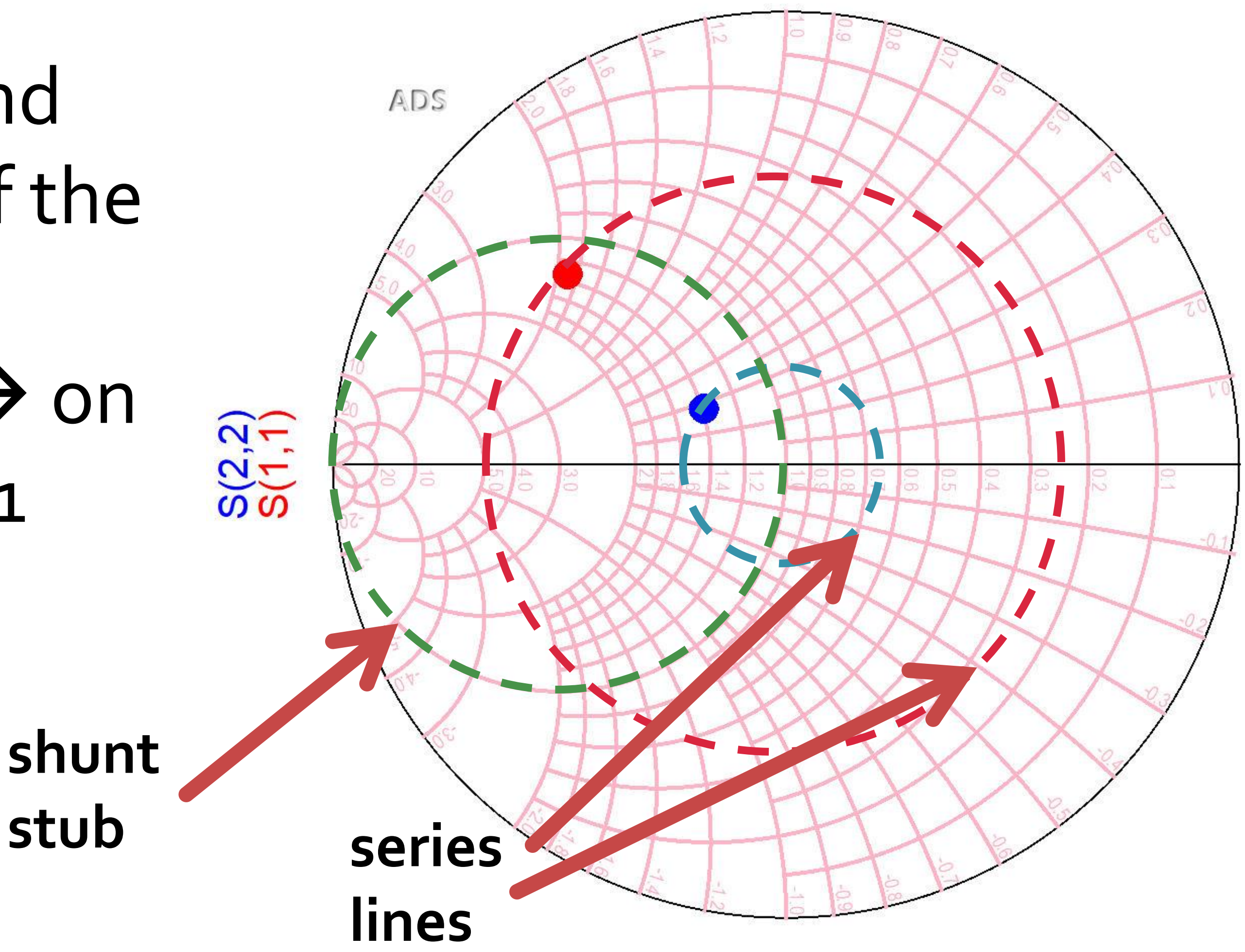
Interstage matching 2

- The two shunt stubs combine into a single one



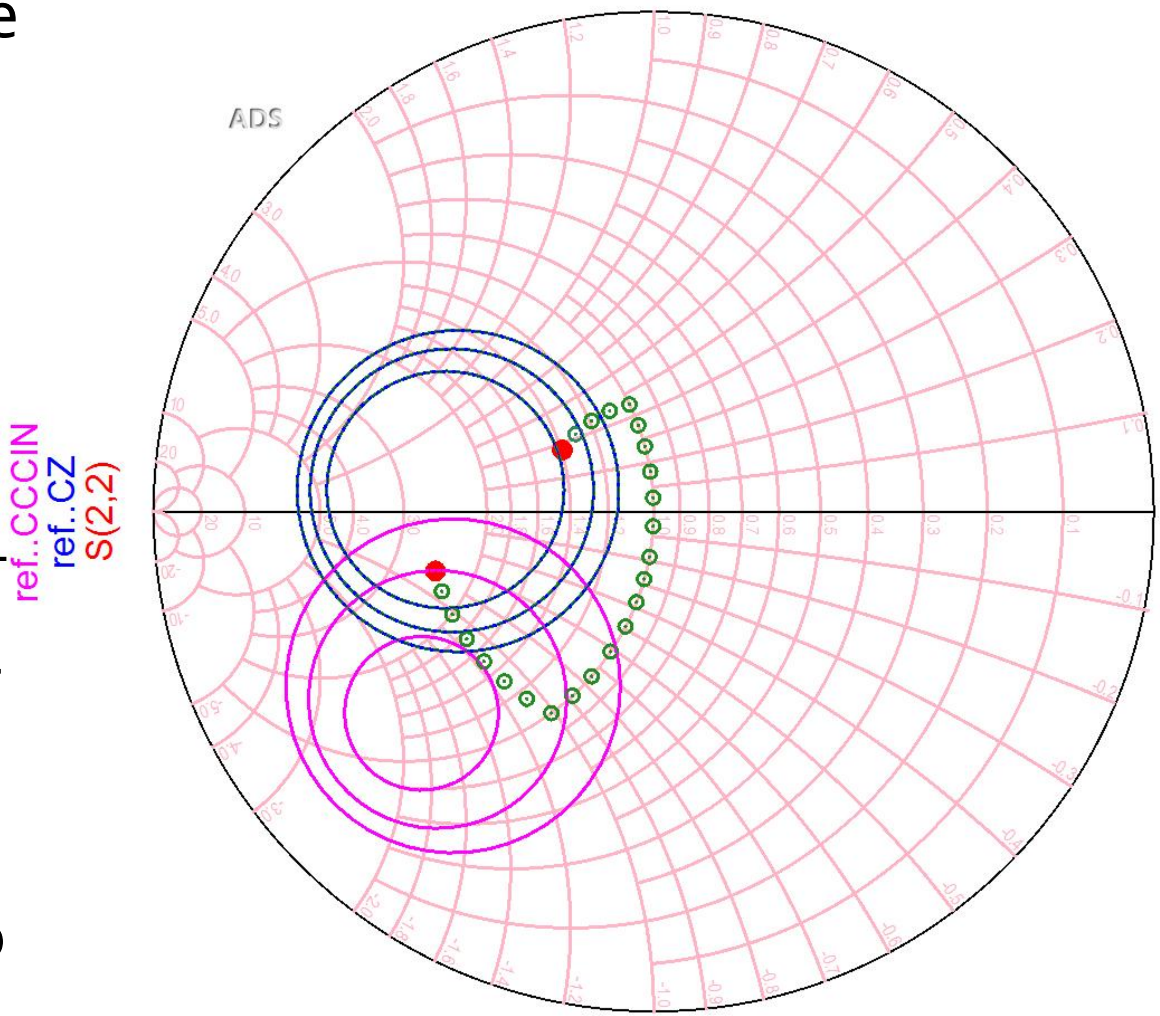
Interstage matching 2

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



Interstage matching 2

- 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 will then combine into one



Output matching stage 1 (L1)

- **G_{L1}** (we use the same point <- output L2), **0.2dB**

$$\Gamma_L = 0.186\angle -132.9^\circ \qquad |\Gamma_L| = 0.186; \quad \varphi = -132.9^\circ$$

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

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

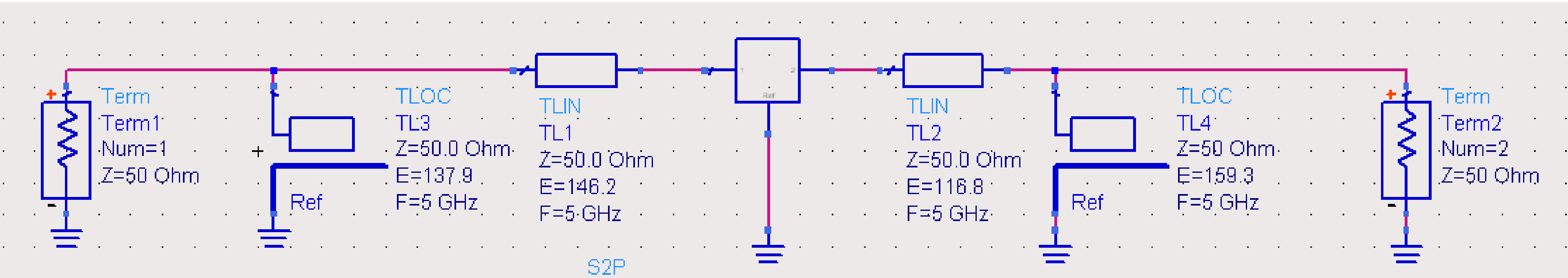
- the length of the shunt stub θ_{sp} is not calculated because it is **not** needed

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

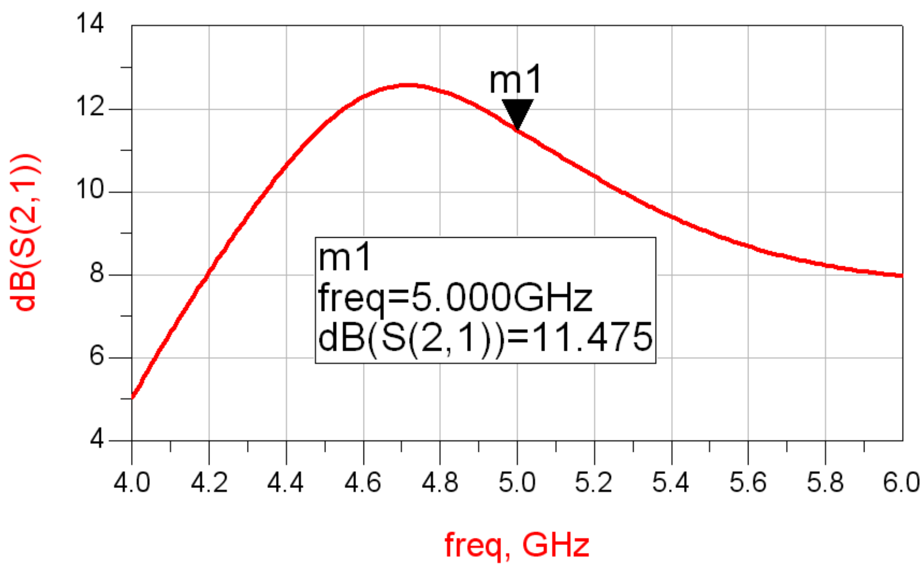
Output matching stage 1 (L1)

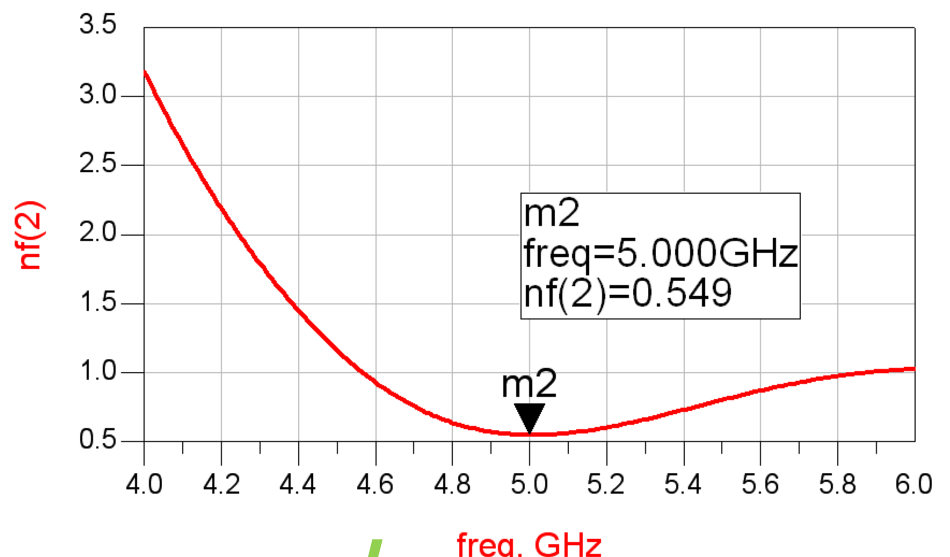
Equation	Solution L1A	Solution L1B
$\Phi+2\theta$	$+100.72°$	$-100.72°$
θ	$116.8°$	$16.1°$
$\mathrm{Im}[y(\theta)]$	-0.379	$+0.379$

Verify stage 1



G_{S1} G_o G_{L1}

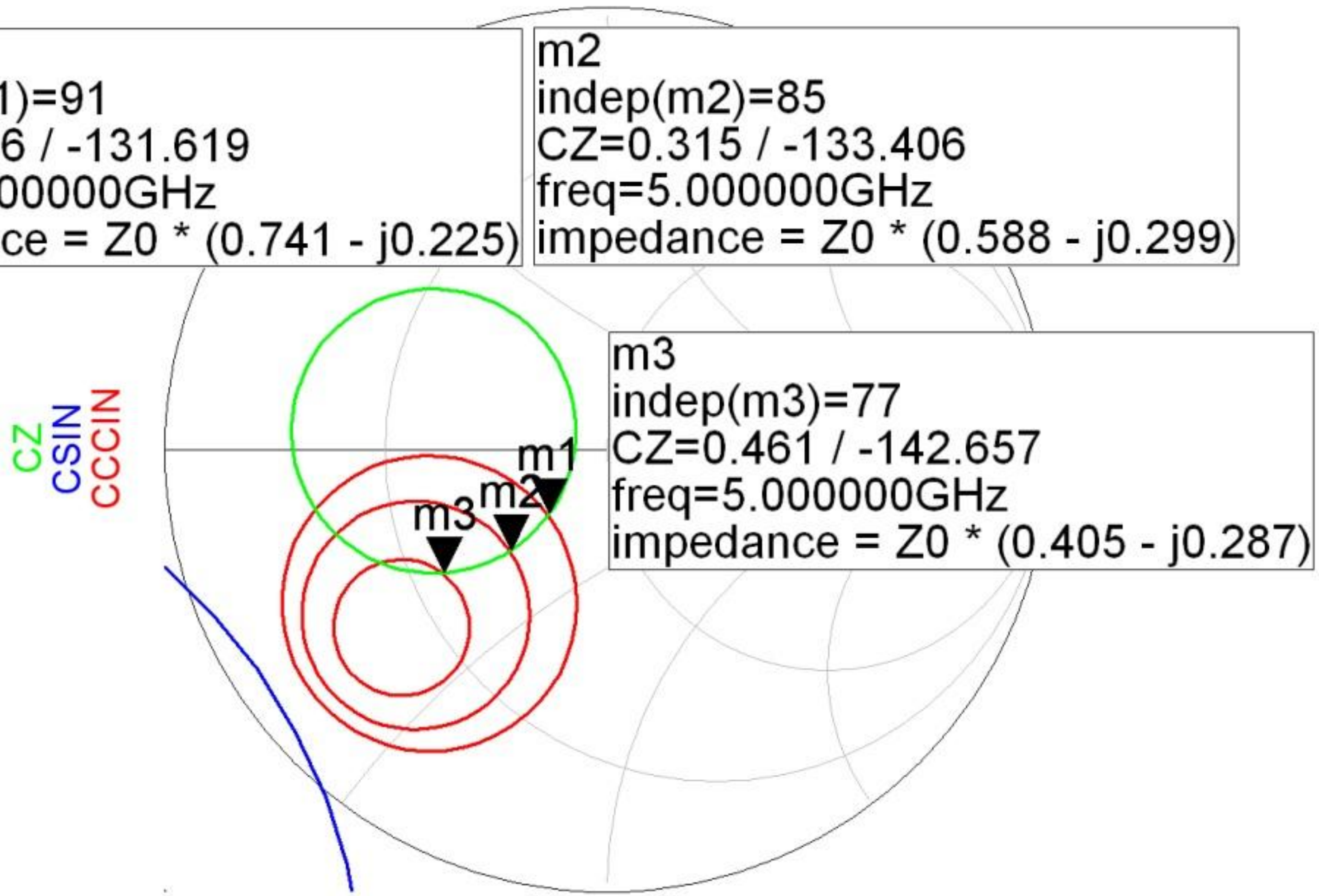




$F_1 < 0.7$ dB, $G_1 > 9$ dB √

Input matching stage 2 (S2)

- G_{S_2} (moving from Γ_{S_2} we choose towards complex plane origin – **m3** – gain 2dB)



Input matching stage 2 (S2)

- **G_{S_2}** (going from m3 towards origin), **2dB**

$$\Gamma_{S2} = 0.461\angle -142.66^\circ \qquad |\Gamma_{S2}| = 0.461; \quad \varphi = -142.66^\circ$$

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

$$\cos(\varphi + 2\theta) = -0.461 \Rightarrow \quad (\varphi + 2\theta) = \pm 117.45^\circ$$

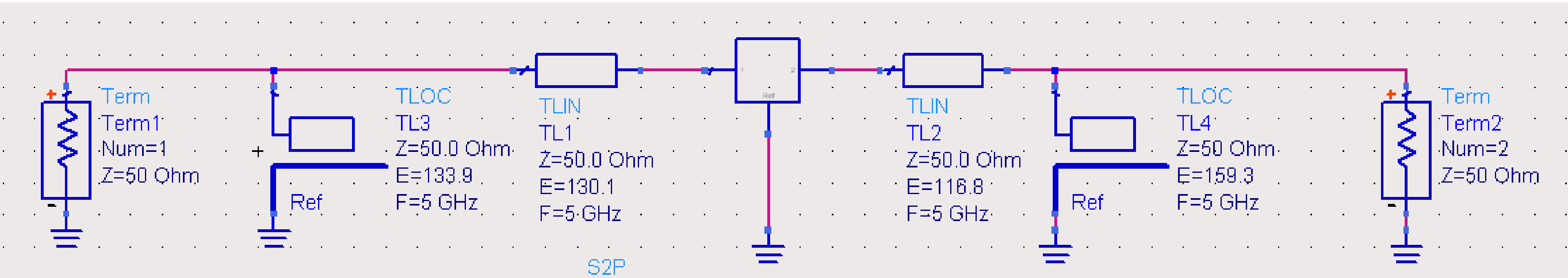
- the length of the shunt stub θ_{sp} is not calculated because it is **not** needed

$$(\varphi + 2\theta) = \begin{cases} +117.45^\circ \\ -117.45^\circ \end{cases} \quad \theta = \begin{cases} 130.1^\circ \\ 12.6^\circ \end{cases} \quad \mathrm{Im}[y_{S2}(\theta)] = \begin{cases} -1.039 \\ +1.039 \end{cases}$$

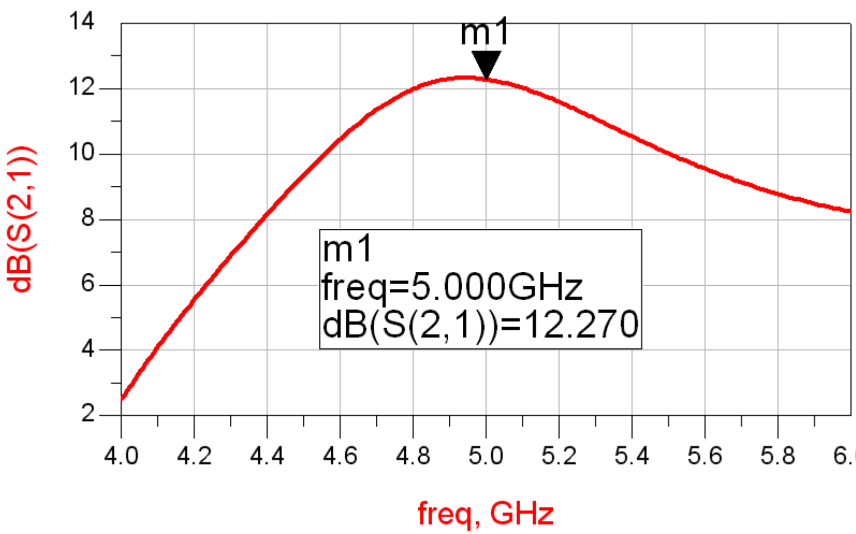
Input matching stage 2 (S2)

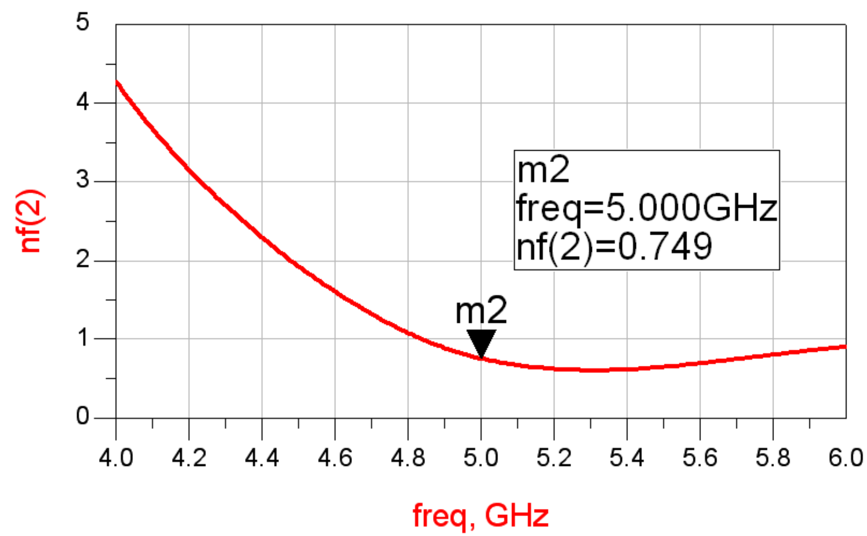
Equation	Solution S2A	Solution S2B
$\Phi+2\theta$	$+117.45°$	$-117.45°$
θ	$130.1°$	$12.6°$
$\mathrm{Im}[y(\theta)]$	-1.039	$+1.039$

Verify stage 2



G_{S1} G_o G_{L1}

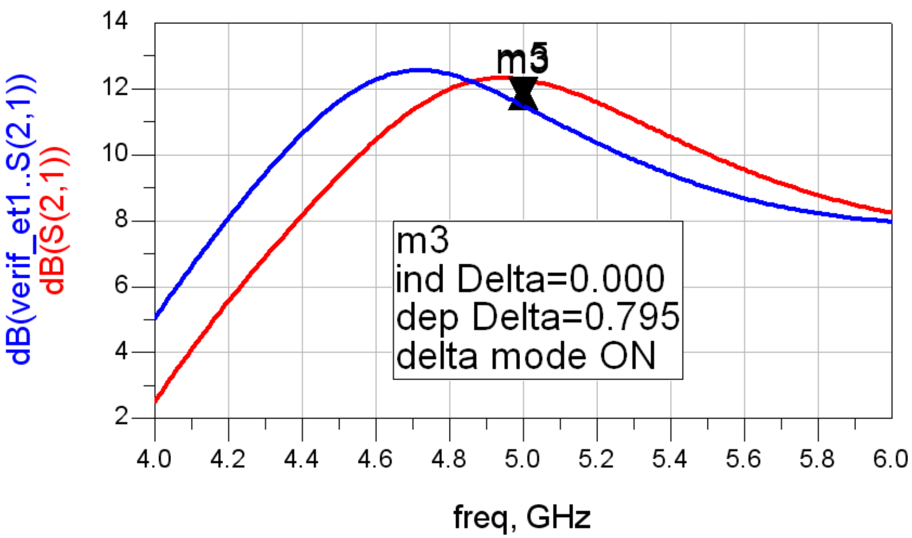


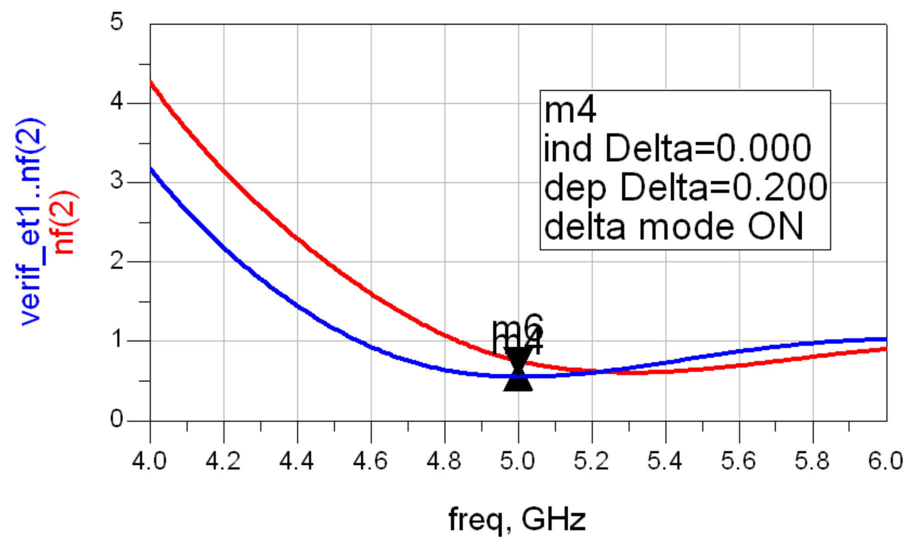


F2 < 1.2 dB, G2 < 13 dB X G1 = 11.5dB,G2 = 12.3dB,G1+G2 > 22dB √

Stage 1/2

- According to the conclusions of the Friis formula, the second stage obtains a higher gain because a higher noise is acceptable.





Merging the two shunt stubs

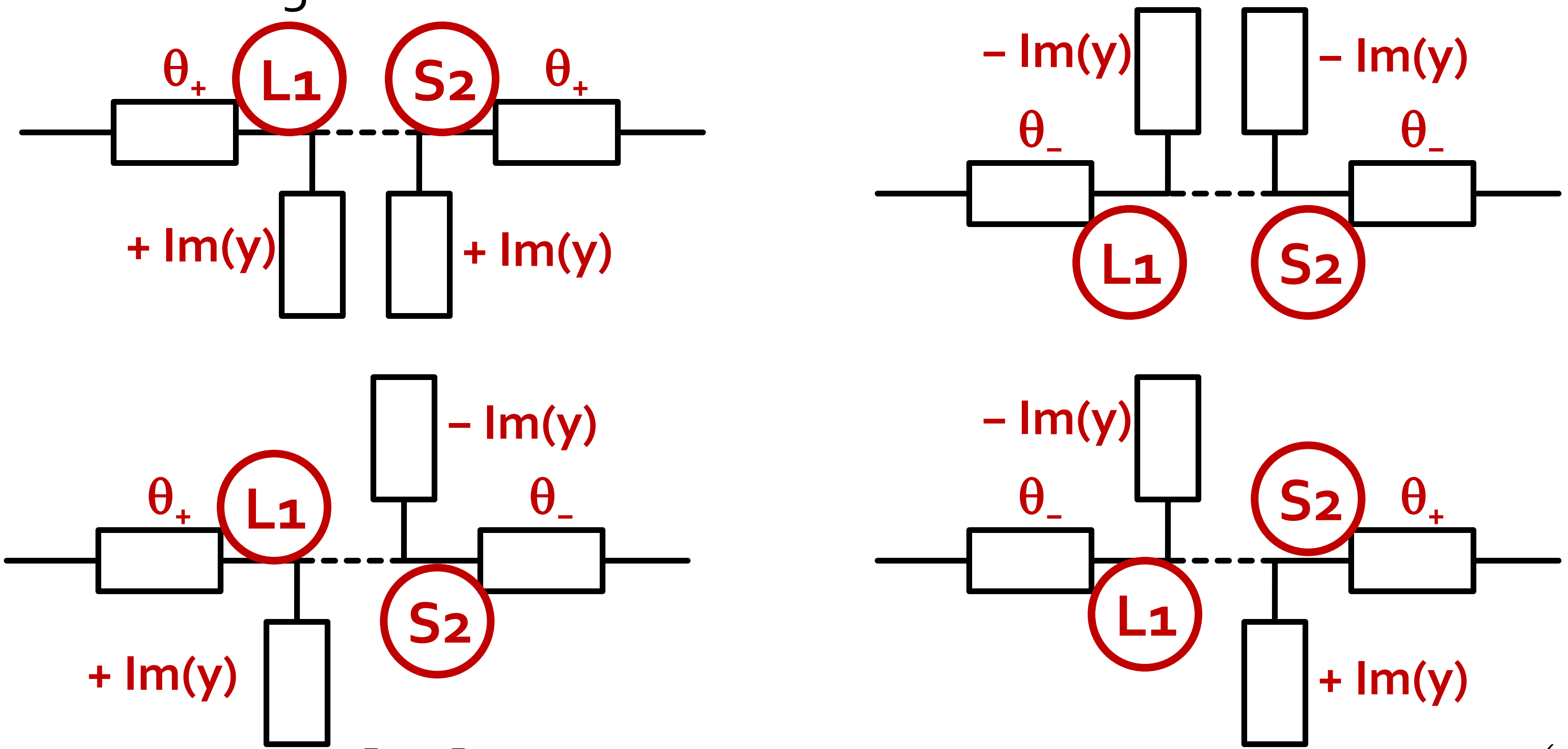
- **The two shunt stubs merge into a single one**
- There are **4 possible combinations** depending on how we chose the electrical length for the two series lines
  - for each chosen electric length (θ) the corresponding Im[y(θ)] must be used
- Ex:

$$\theta_{L1} = 116.8^\circ \quad \theta_{S2} = 130.1^\circ \qquad \mathrm{Im}[y_{sp}] = \mathrm{Im}[y_{L1}(\theta)] + \mathrm{Im}[y_{S2}(\theta)] = -1.418$$

$$\theta_{sp} = \tan^{-1}\left(\mathrm{Im}[y_{sp}]\right) \qquad \theta_{sp} = 125.2^\circ$$

Merging the two shunt stubs

- 4 possible combinations
  - the **admittances** are in parallel and **add** up, not the electrical lengths



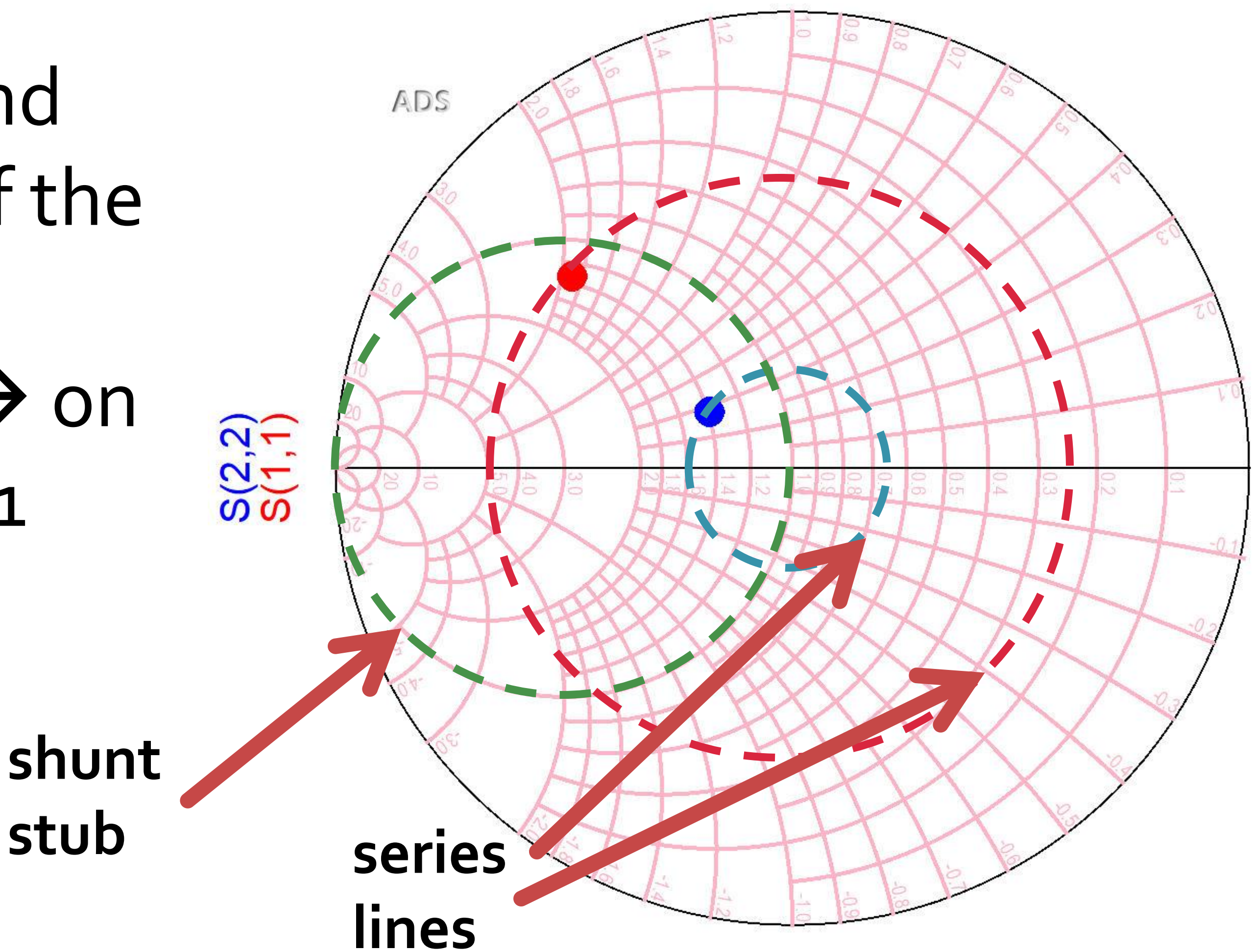
$$\mathrm{Im}[y_{sp}] = \mathrm{Im}[y_{L1}(\theta)] + \mathrm{Im}[y_{S2}(\theta)] \qquad \theta_{sp} = \tan^{-1}(\mathrm{Im}[y_{sp}])$$

Merging the two shunt stubs

		Solution S2A	Solution S2B
		$\theta = 130.1°$ $Im[y(\theta)] = -1.039$	$\theta = 12.6°$ $Im[y(\theta)] = +1.039$
Solution L1A	$\theta = 116.8°$ $Im[y(\theta)] = -0.379$	$\theta_{L1} = 116.8°$ $Im[y(\theta)] = -1.418$ $\theta_p = 125.2°$ $\theta_{S2} = 130.1°$	$\theta_{L1} = 116.8°$ $Im[y(\theta)] = +0.66$ $\theta_p = 33.4°$ $\theta_{S2} = 12.6°$
Solution L1B	$\theta = 16.1°$ $Im[y(\theta)] = +0.379$	$\theta_{L1} = 16.1°$ $Im[y(\theta)] = -0.66$ $\theta_p = 146.6°$ $\theta_{S2} = 130.1°$	$\theta_{L1} = 16.1°$ $Im[y(\theta)] = 1.418$ $\theta_p = 54.8°$ $\theta_{S2} = 12.6°$

Smith Chart

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



Merge 1, Smith Chart

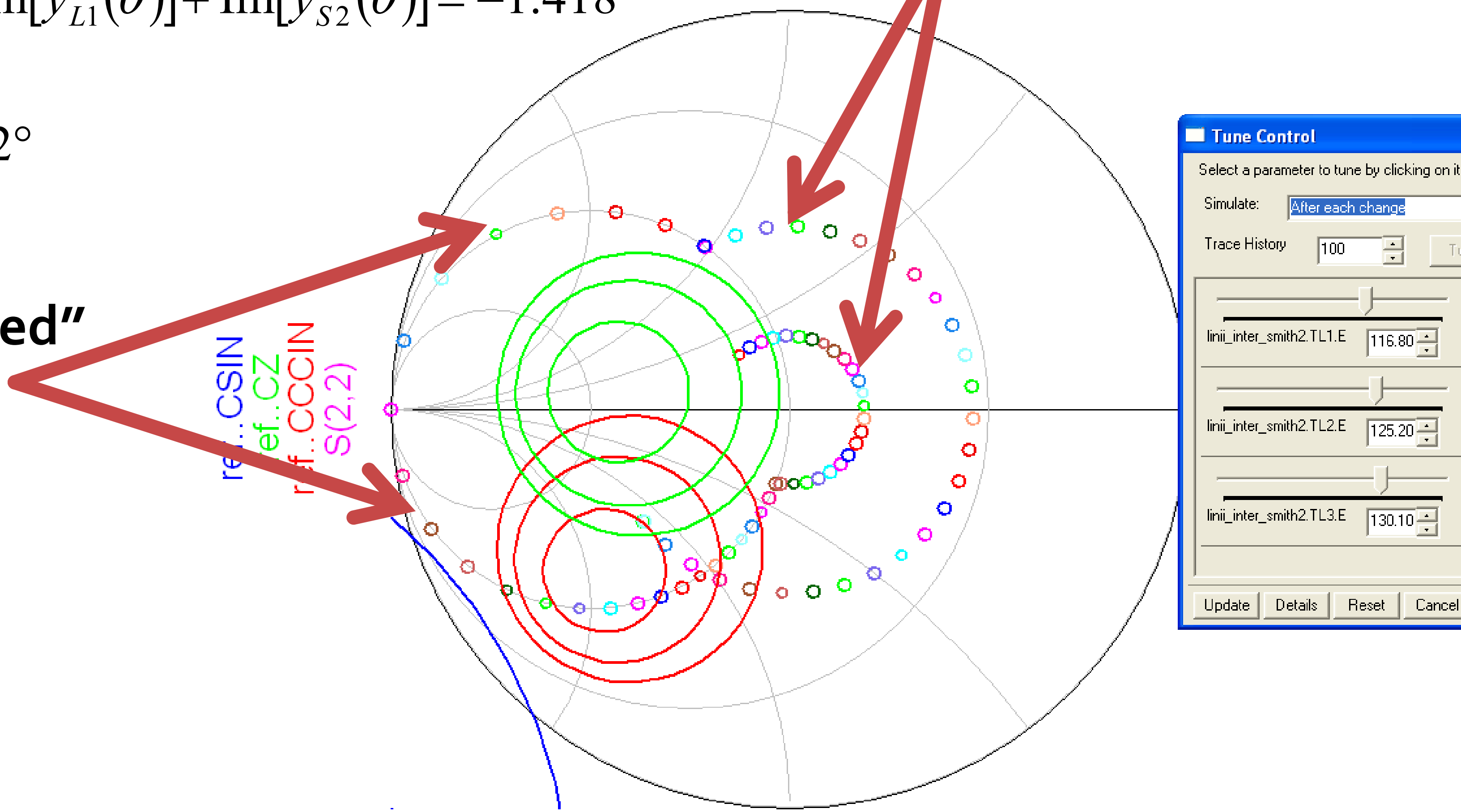
$$\theta_{L1} = 116.8° \qquad \theta_{S2} = 130.1°$$

$$\mathrm{Im}\left[y_{sp}\right] = \mathrm{Im}\left[y_{L1}(\theta)\right] + \mathrm{Im}\left[y_{S2}(\theta)\right] = -1.418$$

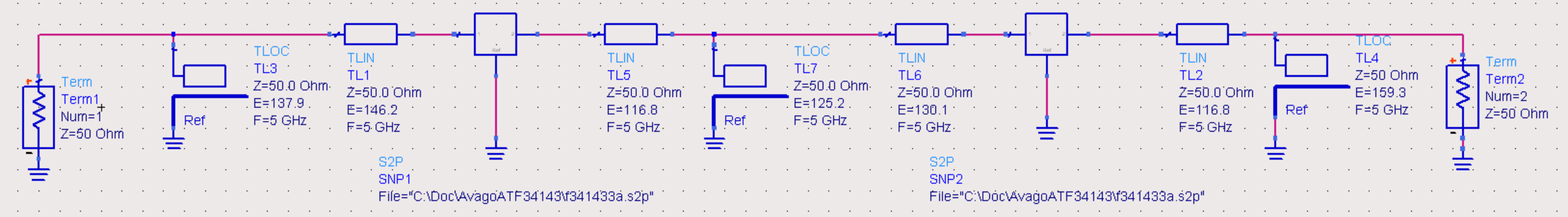
$$\theta_{sp} = 125.2°$$

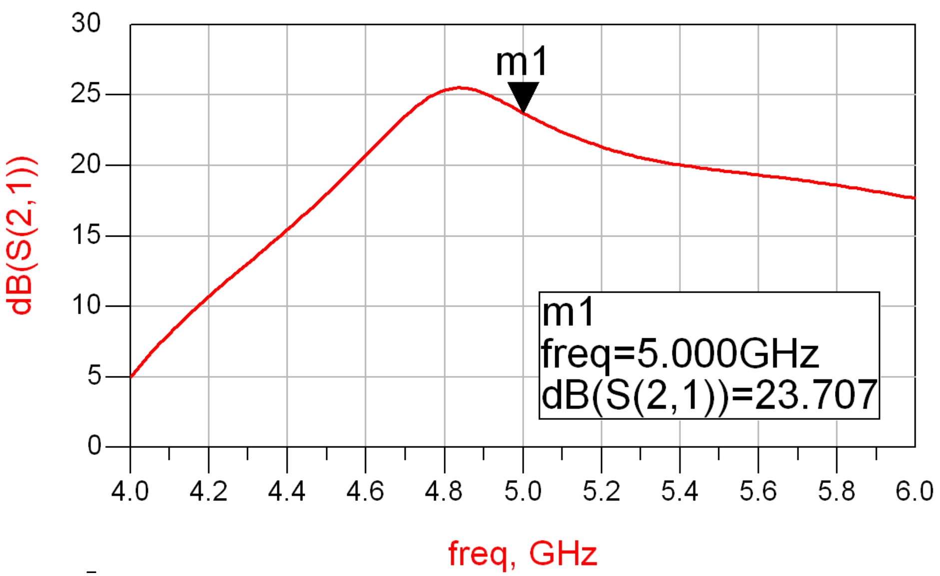
series lines

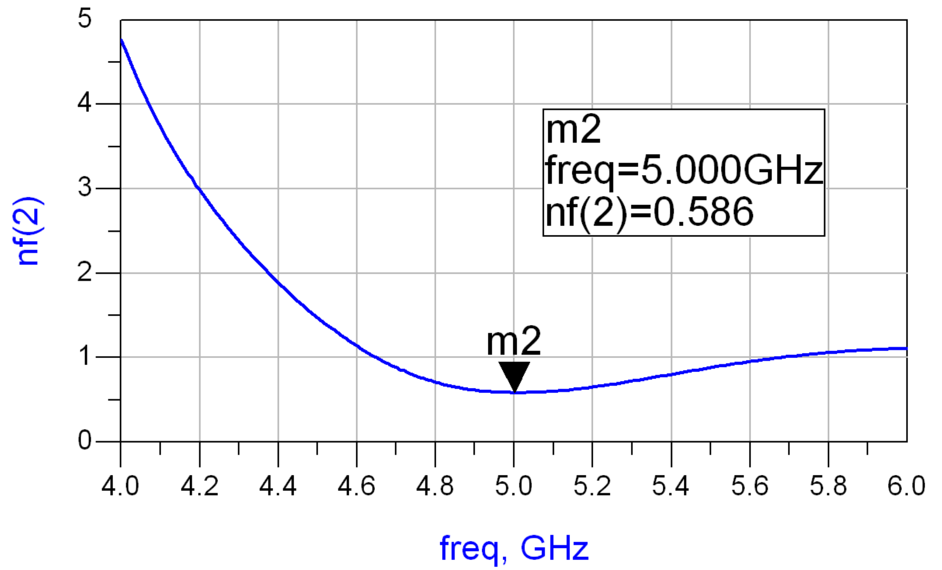
"combined" stub



Merge 1, 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
SNP1
File="C:\Doc\AvagoATF34143\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=116.8
F=5 GHz
Ref
TLOC
TL7
Z=50.0 Ohm
E=125.2
F=5 GHz
TLIN
TL6
Z=50.0 Ohm
E=130.1
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))=23.707
dB(S(2,1))
0
5
10
15
20
25
30
4.0
4.2
4.4
4.6
4.8
5.0
5.2
5.4
5.6
5.8
6.0
freq, GHz


m2
freq=5.000GHz
nf(2)=0.586
m2
nf(2)
0
1
2
3
4
5
4.0
4.2
4.4
4.6
4.8
5.0
5.2
5.4
5.6
5.8
6.0
freq, GHz

Merge 2, Smith Chart

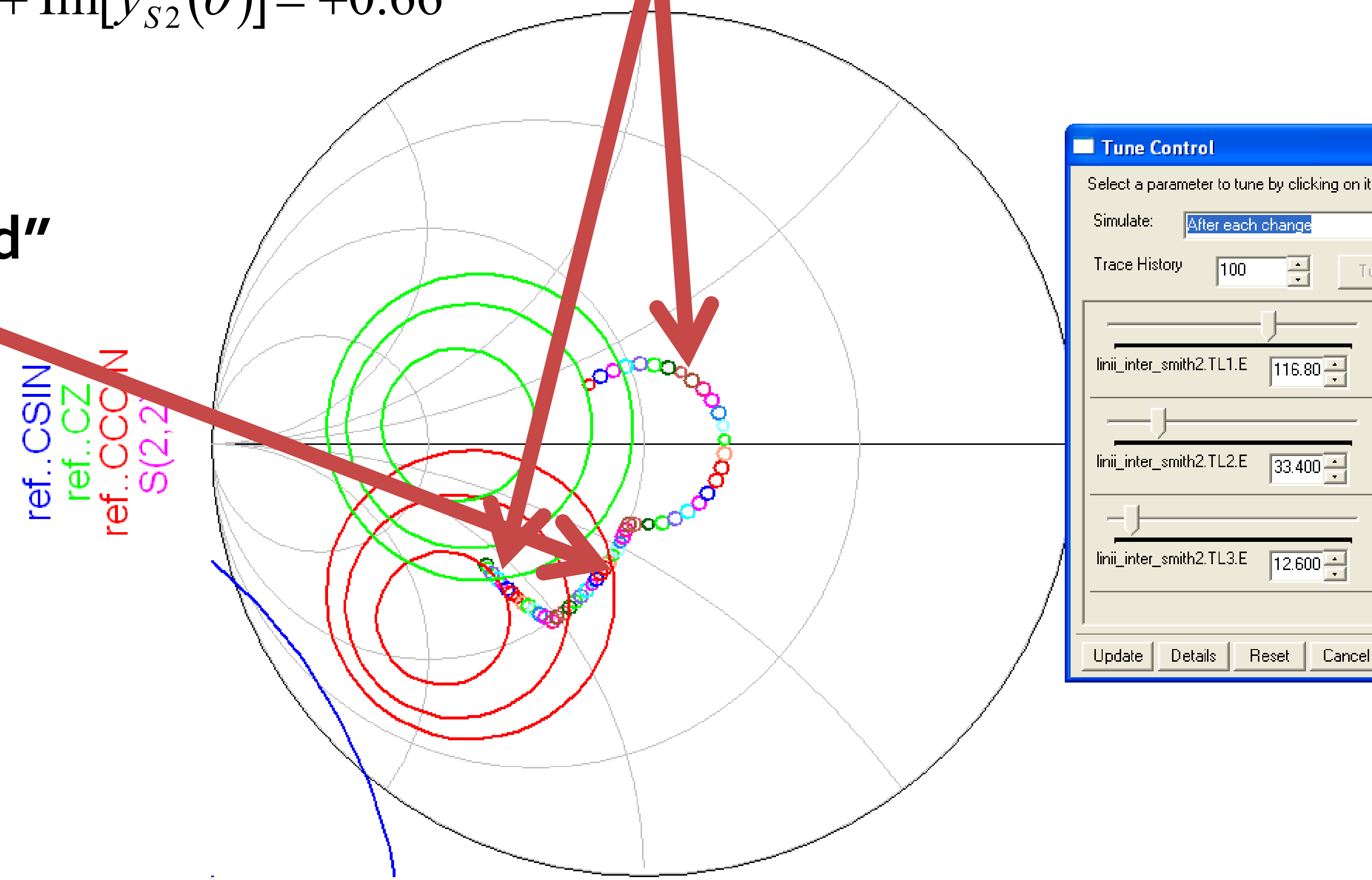
$\theta_{L1} = 116.8°$ $\theta_{S2} = 12.6°$

$\mathrm{Im}[y_{sp}] = \mathrm{Im}[y_{L1}(\theta)] + \mathrm{Im}[y_{S2}(\theta)] = +0.66$

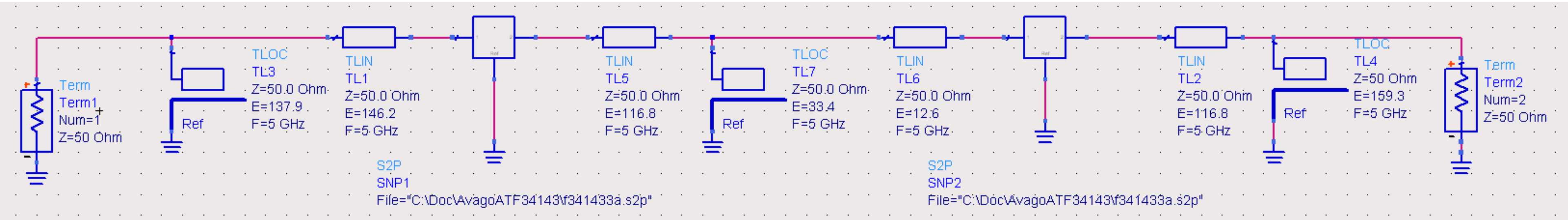
$\theta_{sp} = 33.4°$

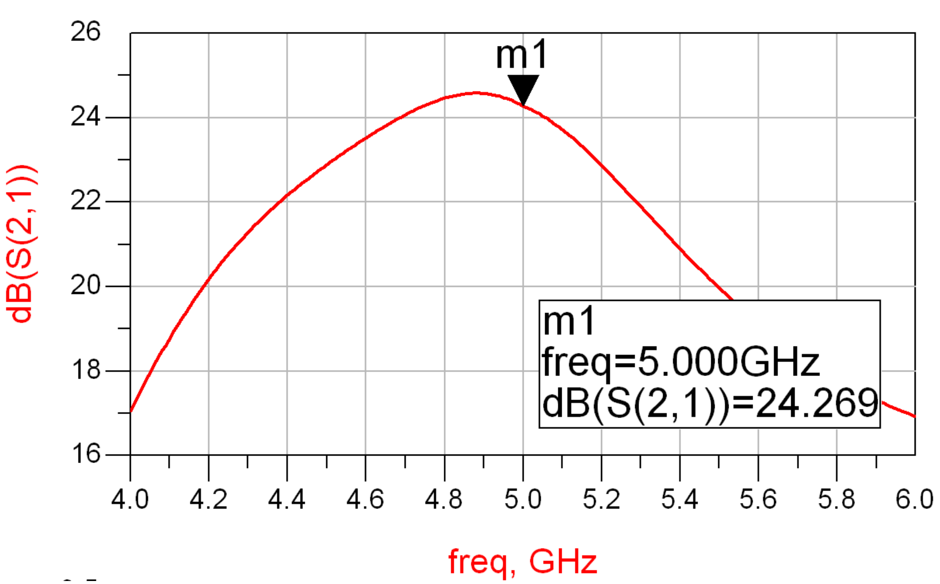
“combined” stub

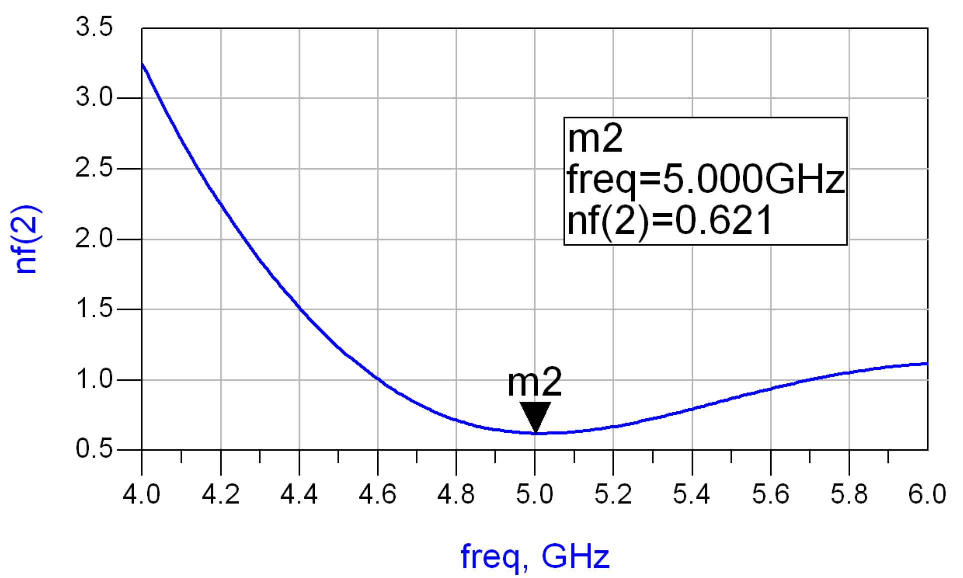
series lines



Merge 2, 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
SNP1
File="C:\Doc\AvagoATF34143\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=116.8
F=5 GHz
Ref
TLOC
TL7
Z=50.0 Ohm
E=33.4
F=5 GHz
TLIN
TL6
Z=50.0 Ohm
E=12.6
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


dB(S(2,1))
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
freq, GHz
m1
m1
freq=5.000GHz
dB(S(2,1))=24.269


nf(2)
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
freq, GHz
m2
m2
freq=5.000GHz
nf(2)=0.621

Merge 3, Smith Chart

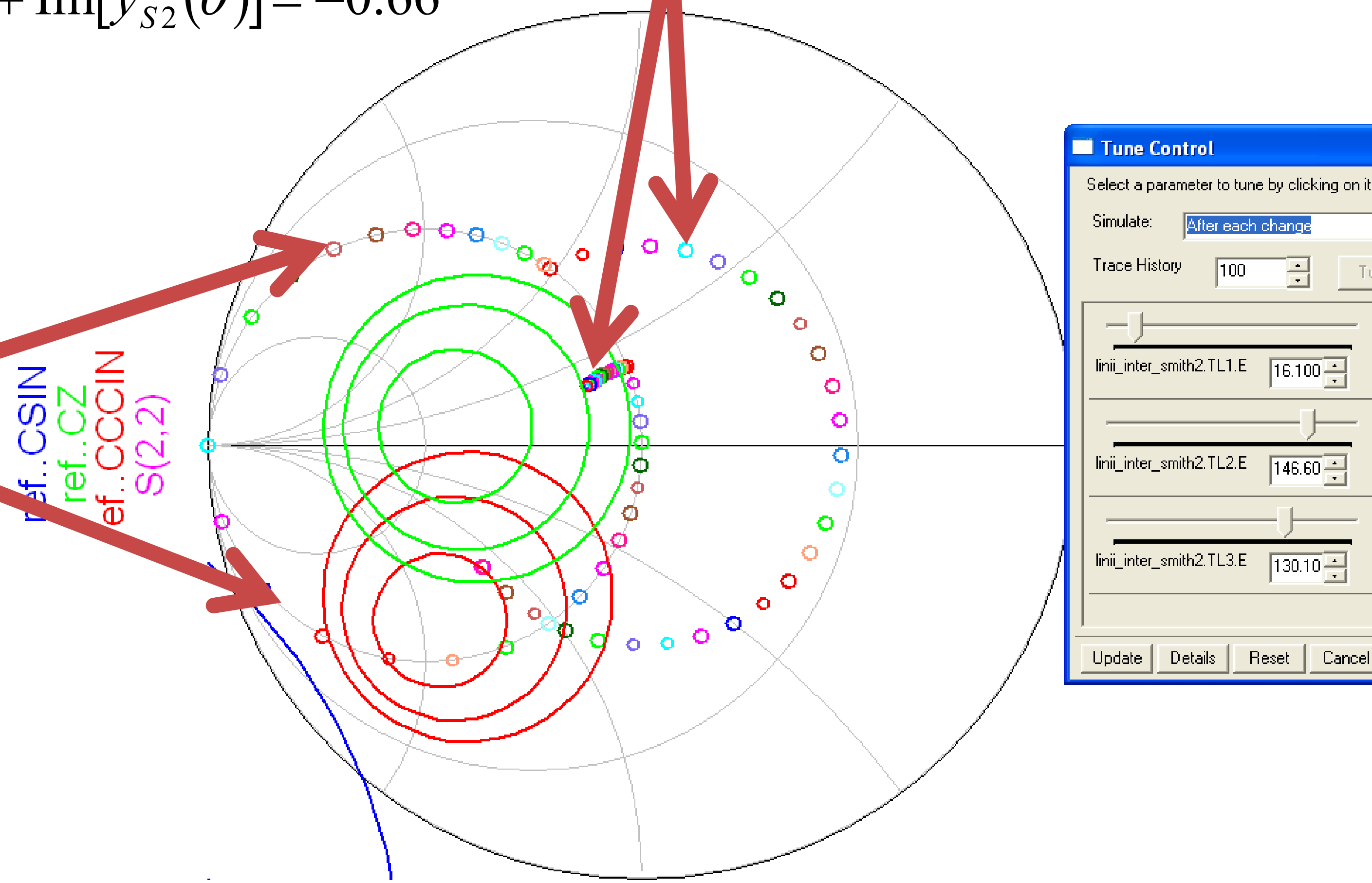
$$\theta_{L1} = 16.1^\circ \qquad \theta_{S2} = 130.1^\circ$$

$$\mathrm{Im}\left[y_{sp}\right] = \mathrm{Im}\left[y_{L1}(\theta)\right] + \mathrm{Im}\left[y_{S2}(\theta)\right] = -0.66$$

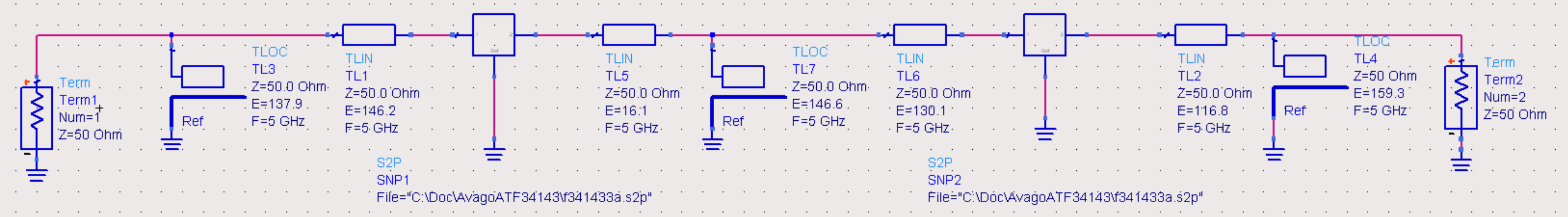
$$\theta_{sp} = 146.6^\circ$$

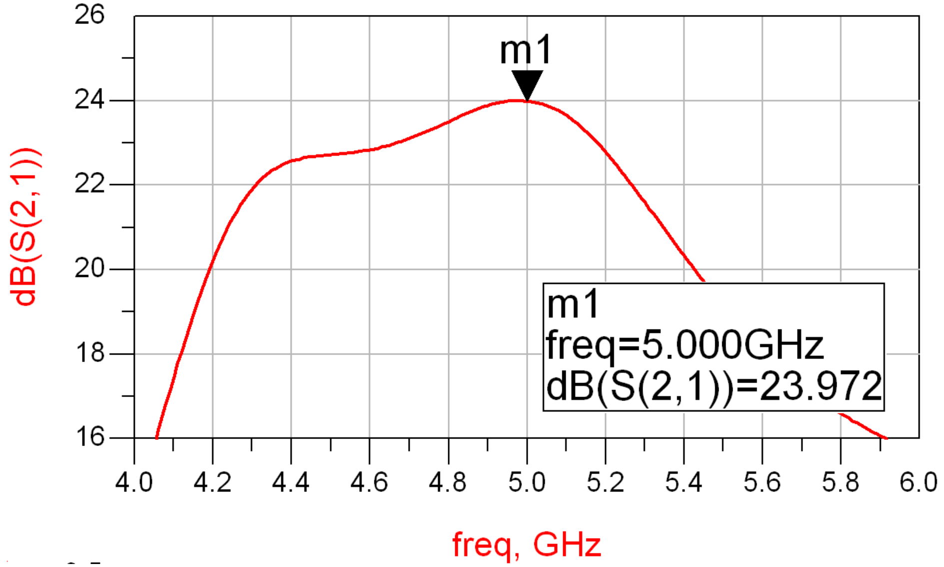
series lines

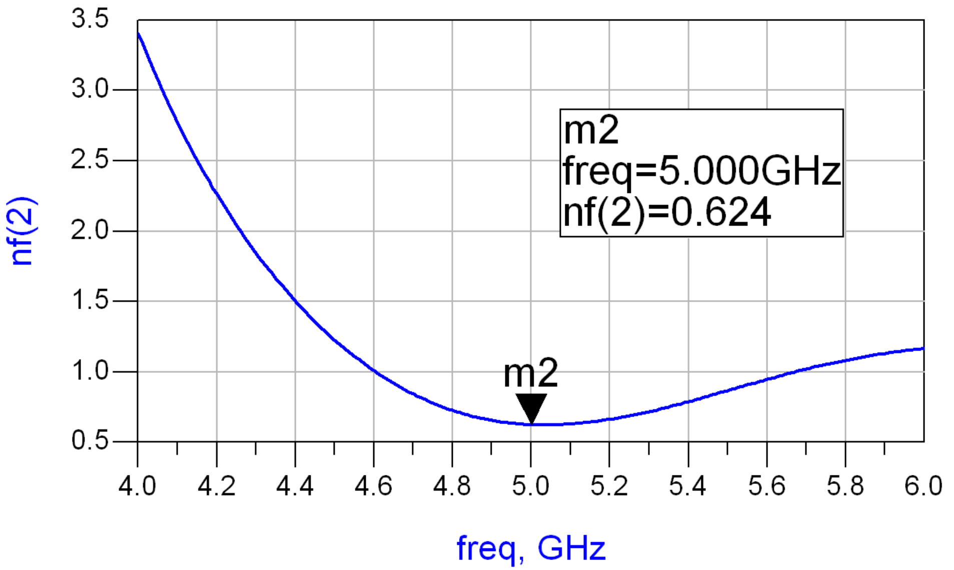
“combined” stub



Merge 3, 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
SNP1
File="C:\Doc\AvagoATF34143\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=16.1
F=5 GHz
Ref
TLOC
TL7
Z=50.0 Ohm
E=146.6
F=5 GHz
TLIN
TL6
Z=50.0 Ohm
E=130.1
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))=23.972
dB(S(2,1))
freq, GHz
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)=0.624
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

Merge 4, Smith Chart

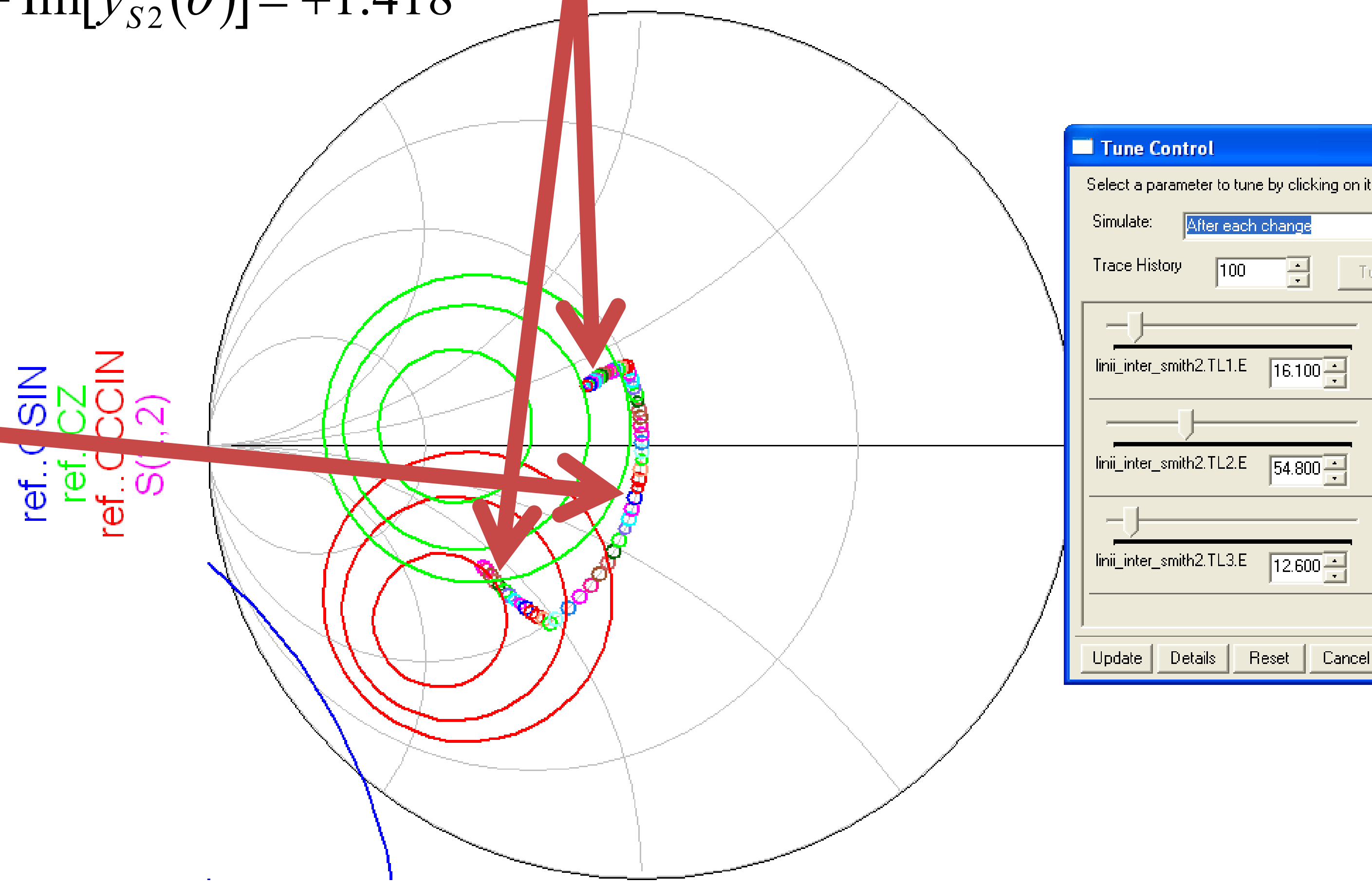
$\theta_{L1} = 16.1°$ $\theta_{S2} = 12.6°$

$$\mathrm{Im}[y_{sp}] = \mathrm{Im}[y_{L1}(\theta)] + \mathrm{Im}[y_{S2}(\theta)] = +1.418$$

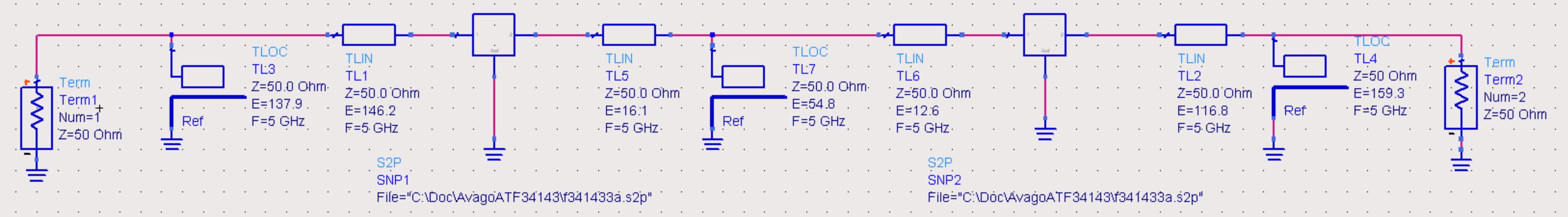
$\theta_{sp} = 54.8°$

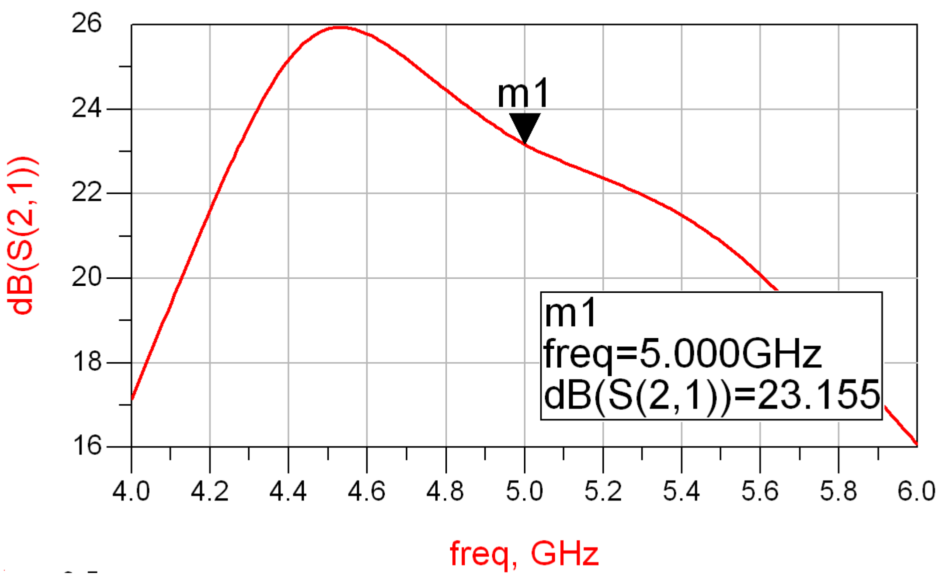
series lines

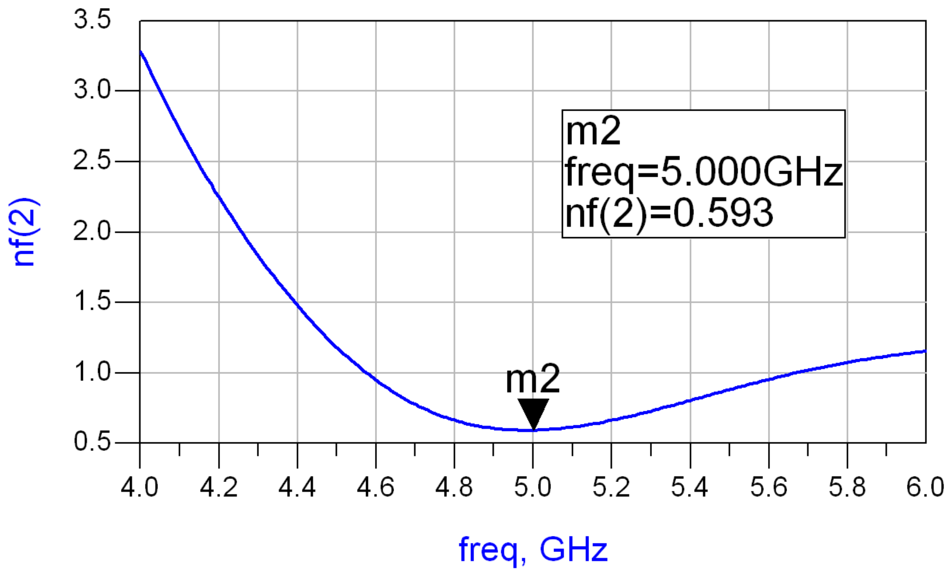
"combined" stub



Merge 4, 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
SNP1
File="C:\Doc\AvagoATF34143\f341433a.s2p"
TLIN
TL5
Z=50.0 Ohm
E=16.1
F=5 GHz
Ref
TLOC
TL7
Z=50.0 Ohm
E=54.8
F=5 GHz
TLIN
TL6
Z=50.0 Ohm
E=12.6
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


dB(S(2,1))
26
24
22
20
18
16
m1
m1
freq=5.000GHz
dB(S(2,1))=23.155
4.0
4.2
4.4
4.6
4.8
5.0
5.2
5.4
5.6
5.8
6.0
freq, GHz


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

- All the combinations obtained meet the target conditions for gain and noise
- Choose a convenient one depending on:
  - the physical dimensions of the lines $l = \dfrac{\theta}{360°} \cdot \lambda$
  - frequency bandwidth/flatness
  - stability
  - performance (noise/gain)
  - input and output reflection
  - etc.

Supplement Mini Project

Implementation in microstrip technology

- microstrip lines
  - dielectric layer
  - plane metallization (ground plane)
  - traces which will control:
    - characteristic impedance
    - physical/electrical length

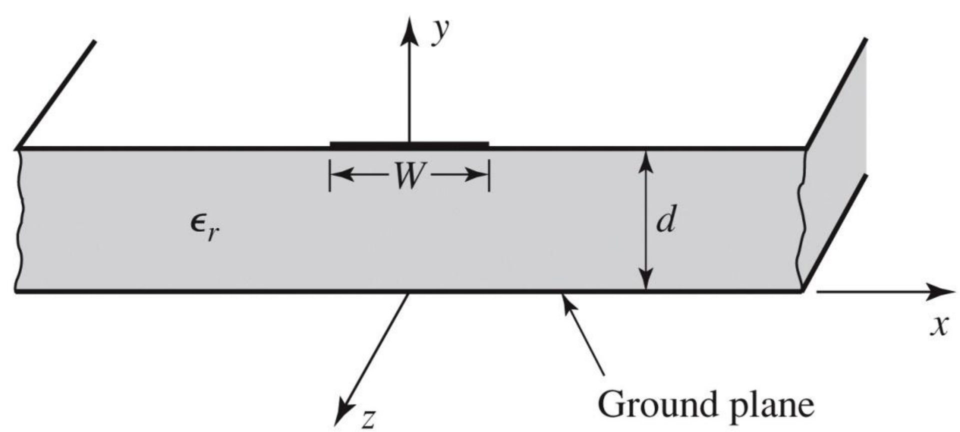


Figure 3.25a

Implementation in microstrip technology

- quasi TEM line

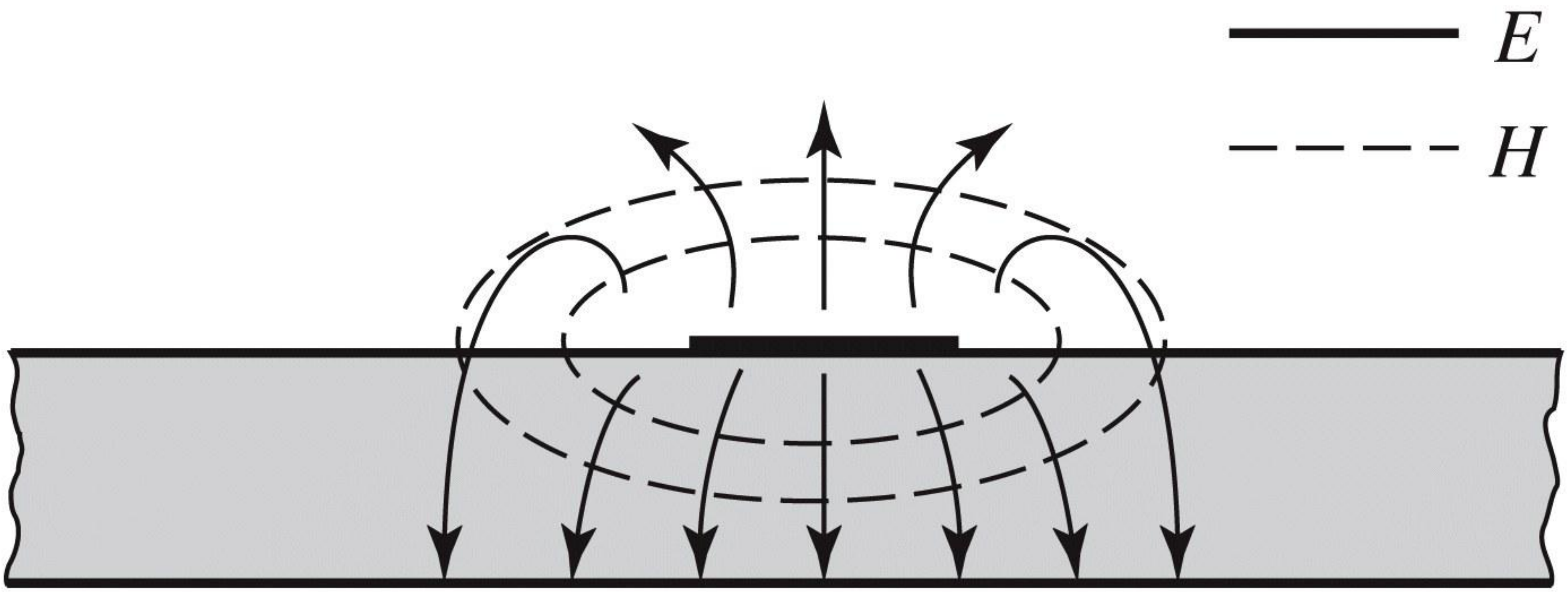
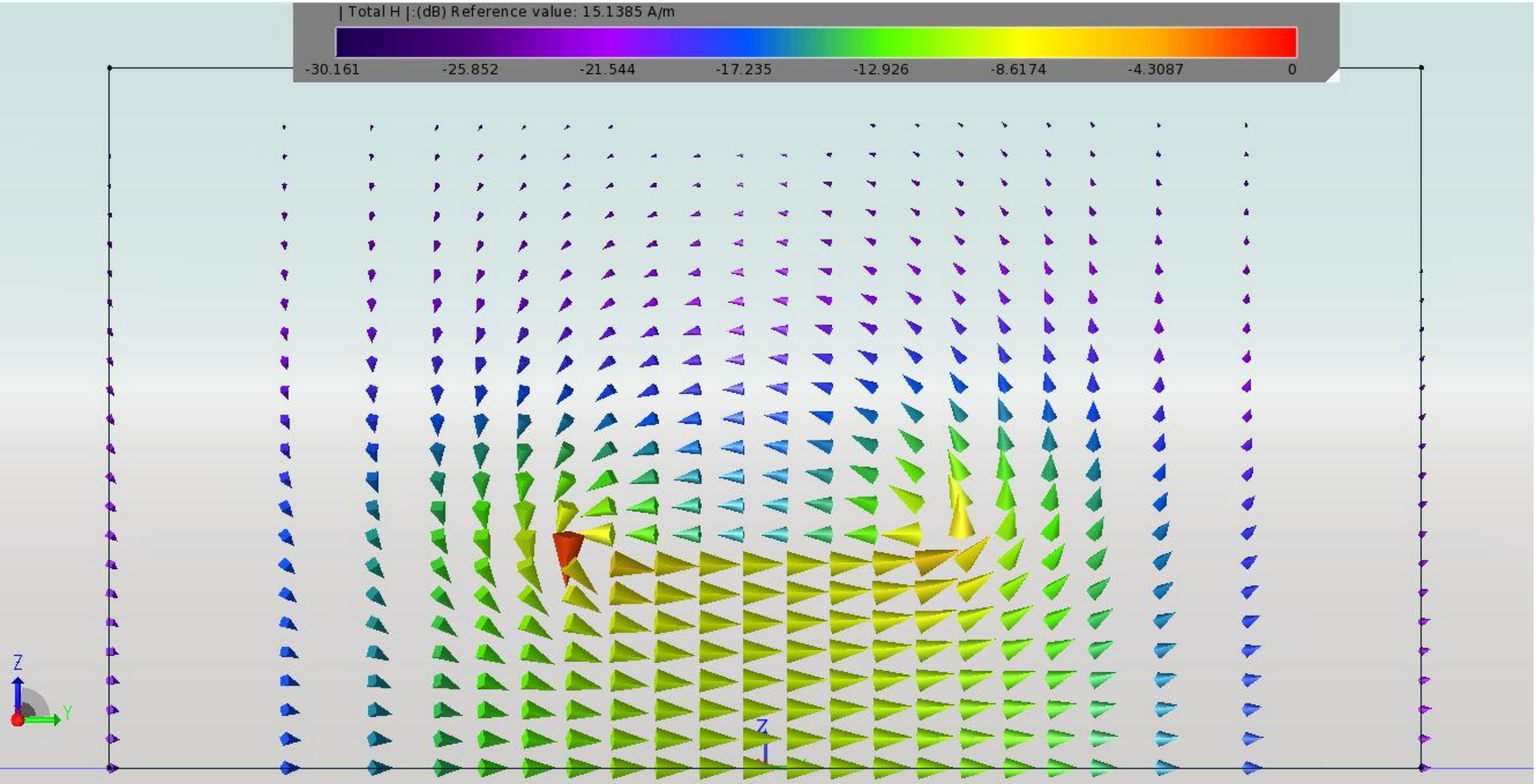


Figure 3.25b

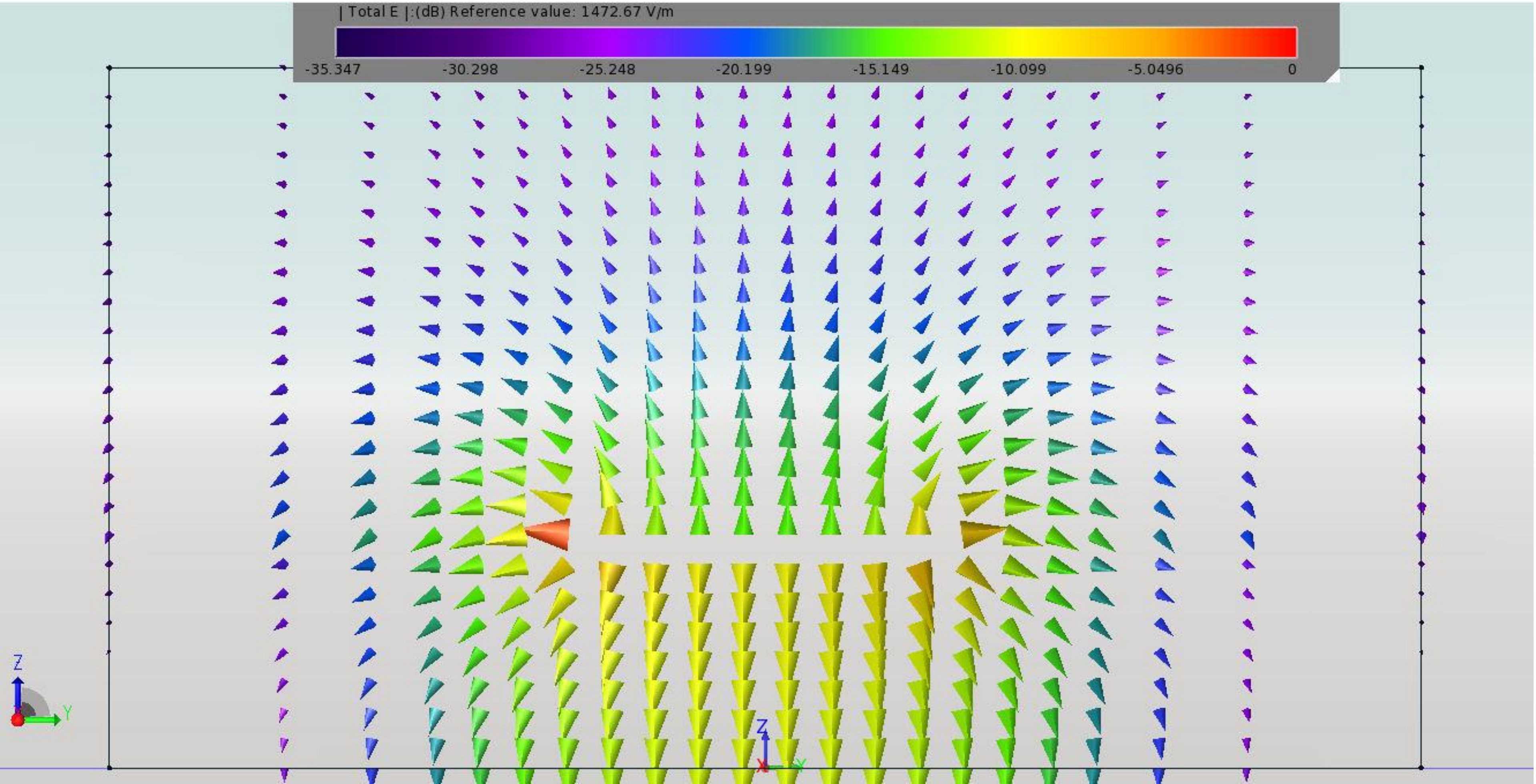
Implementation in microstrip technology

- quasi TEM line, EmPro



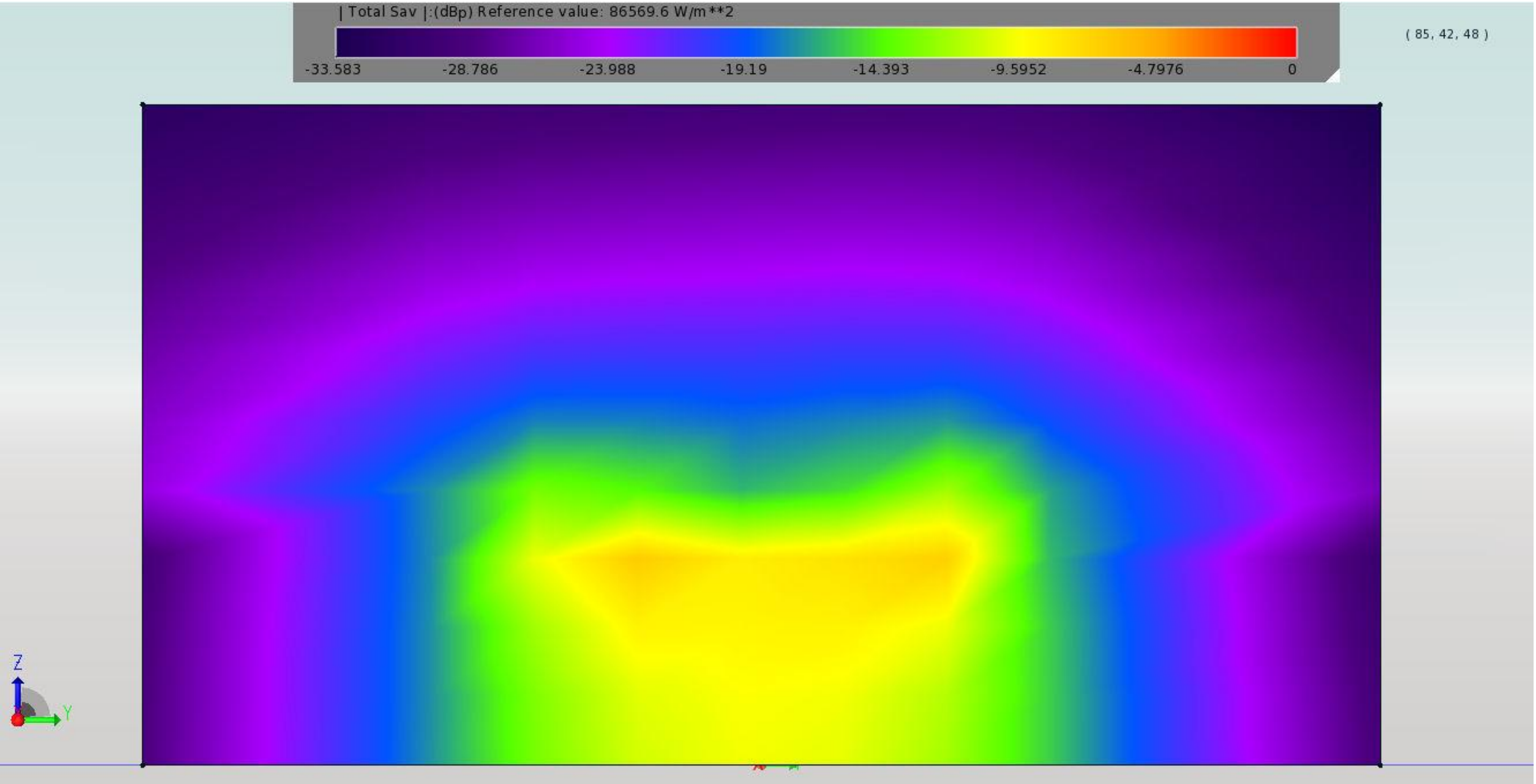
Implementation in microstrip technology

- quasi TEM line, EmPro



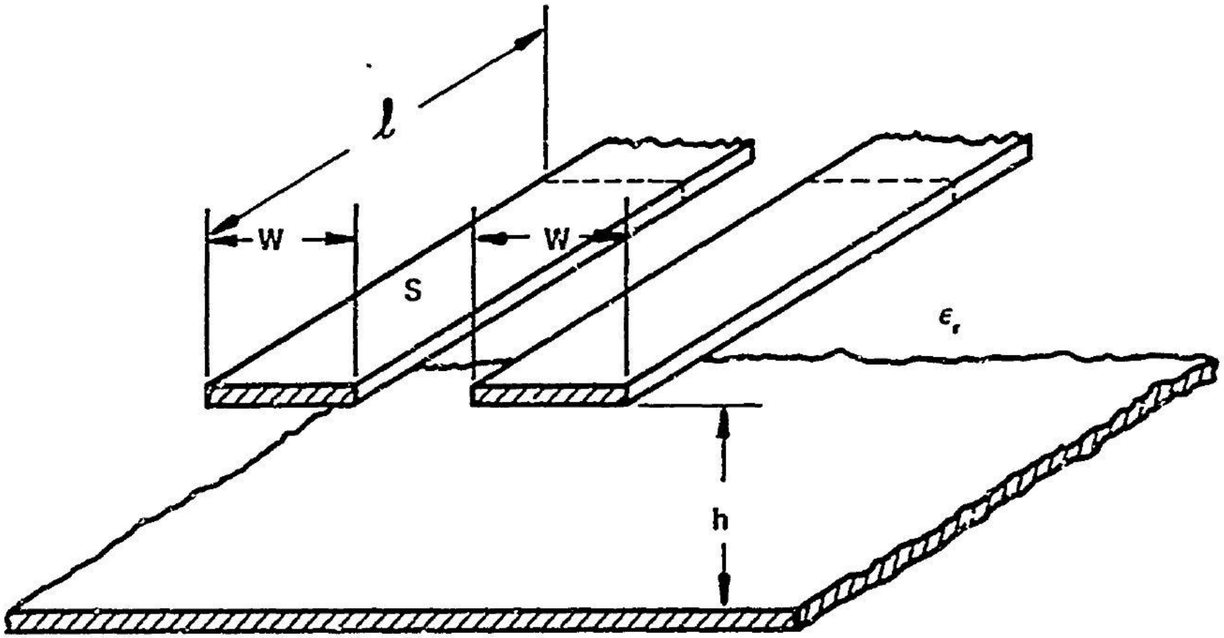
Implementation in microstrip technology

- quasi TEM line, EmPro



Implementation in microstrip technology

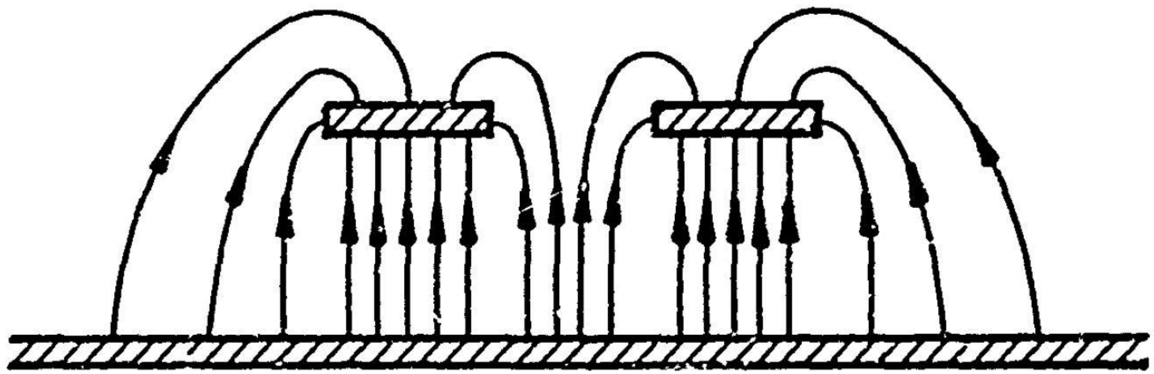
- ~ quasi TEM



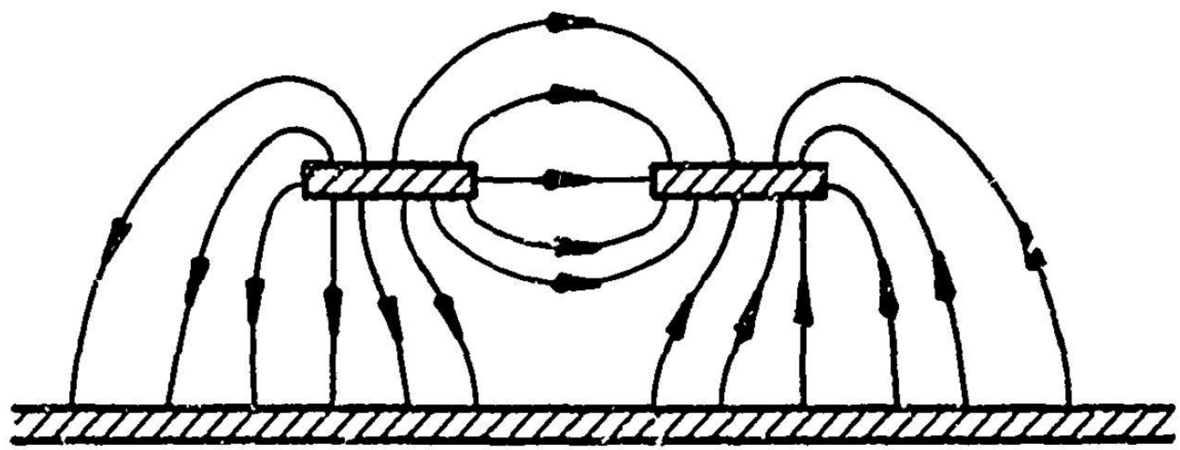
a) COUPLED STRIP GEOMETRY

Implementation in microstrip technology

- ~ quasi TEM



b) EVEN MODE ELECTRIC FIELD PATTERN (SCHEMATIC)



c) ODD MODE ELECTRIC FIELD PATTERN (SCHEMATIC)

Implementation in microstrip technology

- Equivalent geometry of a quasi-TEM microstrip line with effective dielectric constant homogeneous medium

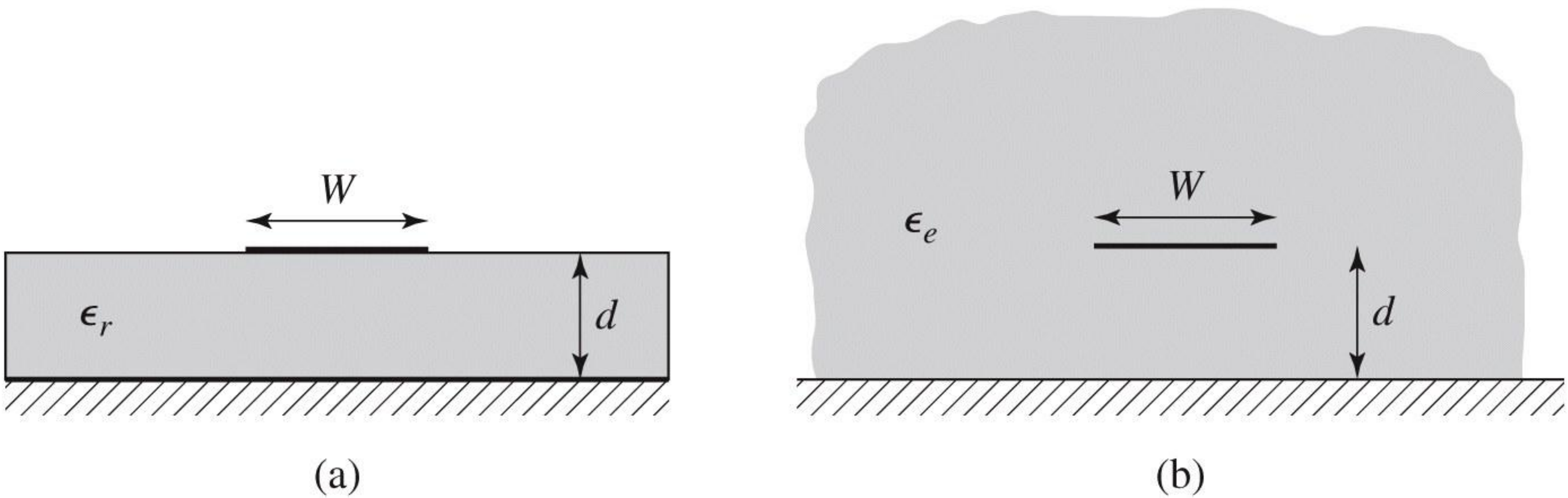


Figure 3.26

Design

- Empirical formulas

$$v_p = \frac{c}{\sqrt{\epsilon_e}},$$

$$\beta = k_0 \sqrt{\epsilon_e},$$

$$\epsilon_e = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \frac{1}{\sqrt{1 + 12d/W}}.$$

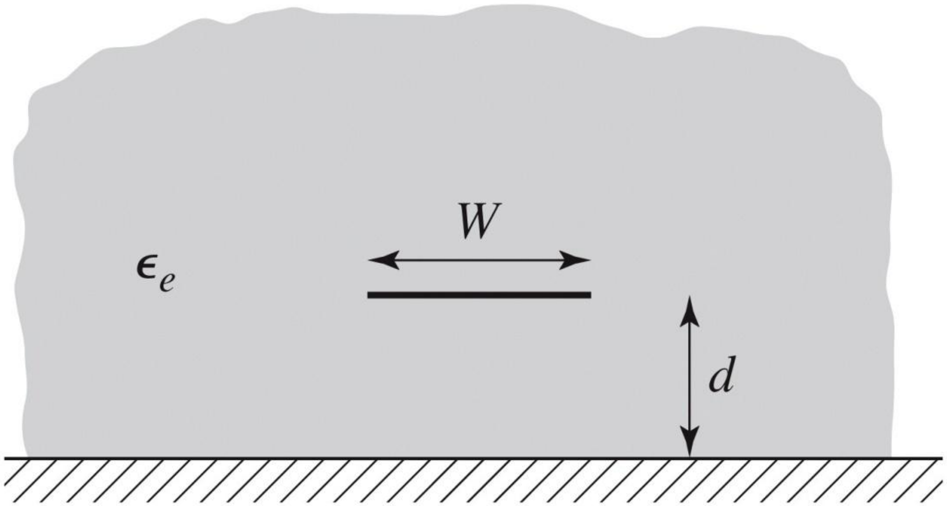


Figure 3.26b

$$Z_0 = \begin{cases} \dfrac{60}{\sqrt{\epsilon_e}} \ln\left(\dfrac{8d}{W} + \dfrac{W}{4d}\right) & \text{for } W/d \leq 1 \\ \dfrac{120\pi}{\sqrt{\epsilon_e}\left[W/d + 1.393 + 0.667 \ln\left(W/d + 1.444\right)\right]} & \text{for } W/d \geq 1. \end{cases}$$

Design

- Empirical formulas

$$A = \frac{Z_0}{60}\sqrt{\frac{\epsilon_r + 1}{2}} + \frac{\epsilon_r - 1}{\epsilon_r + 1}\left(0.23 + \frac{0.11}{\epsilon_r}\right)$$

$$B = \frac{377\pi}{2Z_0\sqrt{\epsilon_r}}.$$

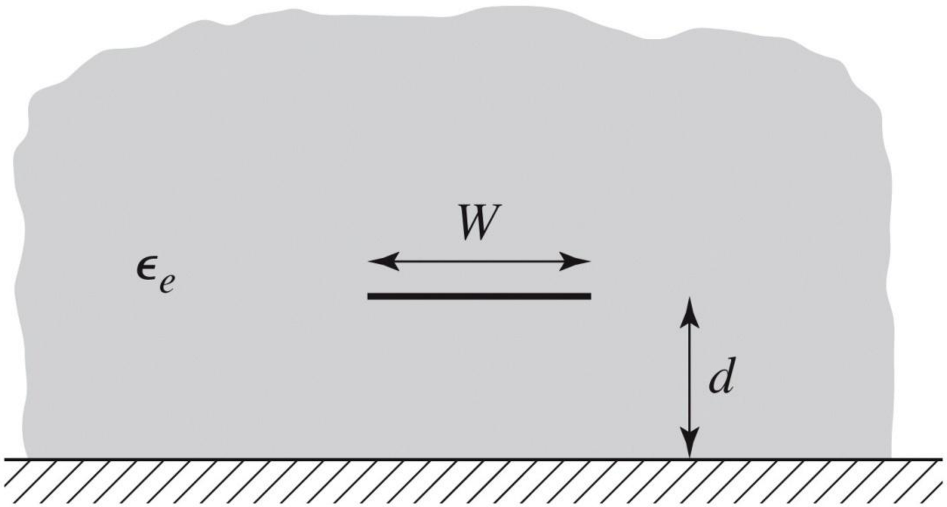
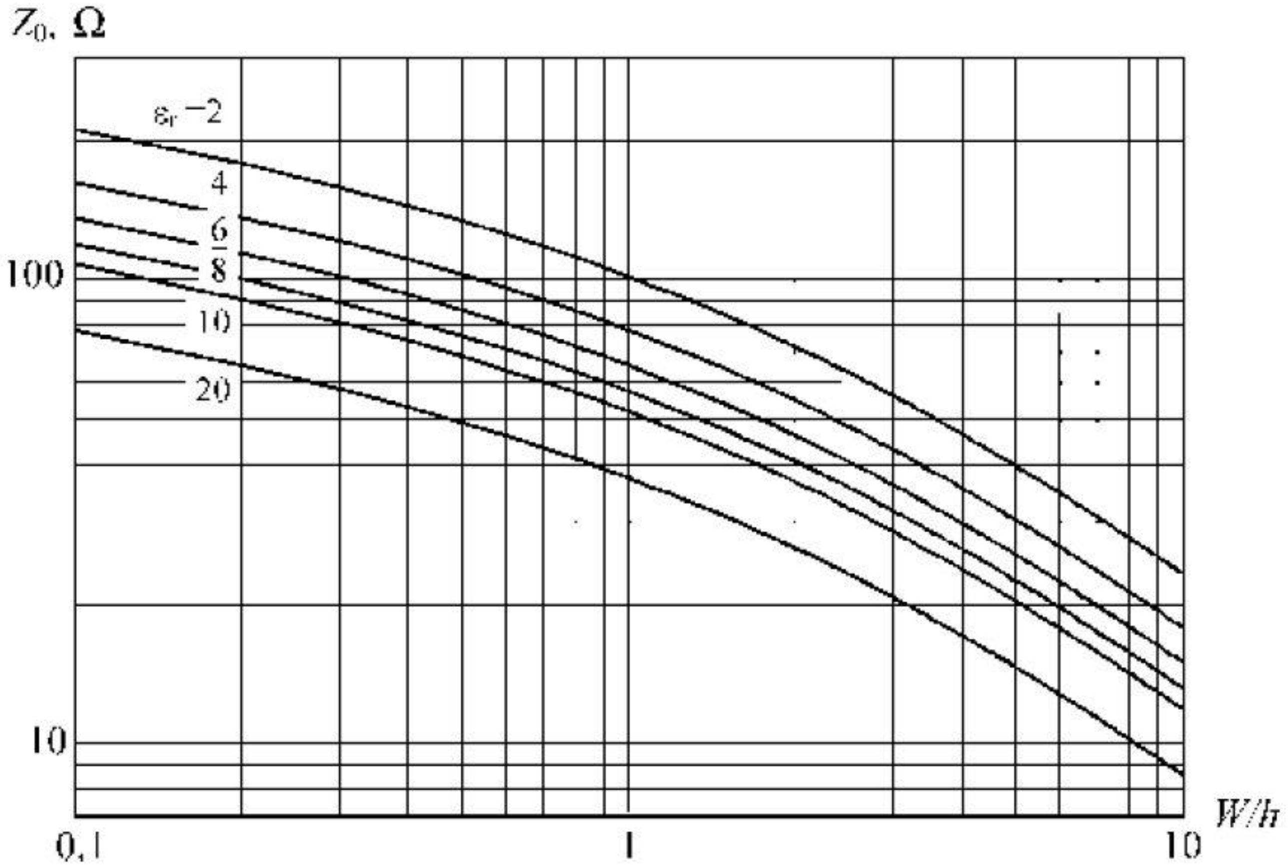


Figure 3.26b

$$\frac{W}{d} = \begin{cases} \dfrac{8e^A}{e^{2A} - 2} & \text{for } W/d < 2 \\ \dfrac{2}{\pi}\left[B - 1 - \ln(2B - 1) + \dfrac{\epsilon_r - 1}{2\epsilon_r}\left\{\ln(B - 1) + 0.39 - \dfrac{0.61}{\epsilon_r}\right\}\right] & \text{for } W/d > 2, \end{cases}$$

Characteristic impedance

- **Large impedances** require **narrow traces**
- **Small impedances** require **wide traces**



$$k_0 = \frac{2\pi f}{c}$$

$$\beta \ell = \sqrt{\epsilon_e} k_0 \ell,$$

Microstrip standardization

- Standardization
  - dimensions in **mil**
  - 1 mil = 10^{-3} inch
  - 1 inch = 2.54 cm
- Trace thickness
  - based on the weight of the deposited copper
  - oz/ft^2
  - 1oz=28.35g and 1ft=30.48cm

Weight of the deposited copper		Trace thickness	
oz/ft^2	g/ft^2	inch	mm
0.5	14.175	0.0007	0.0178
1.0	28.35	0.0014	0.0356
2.0	56.7	0.0028	0.0712

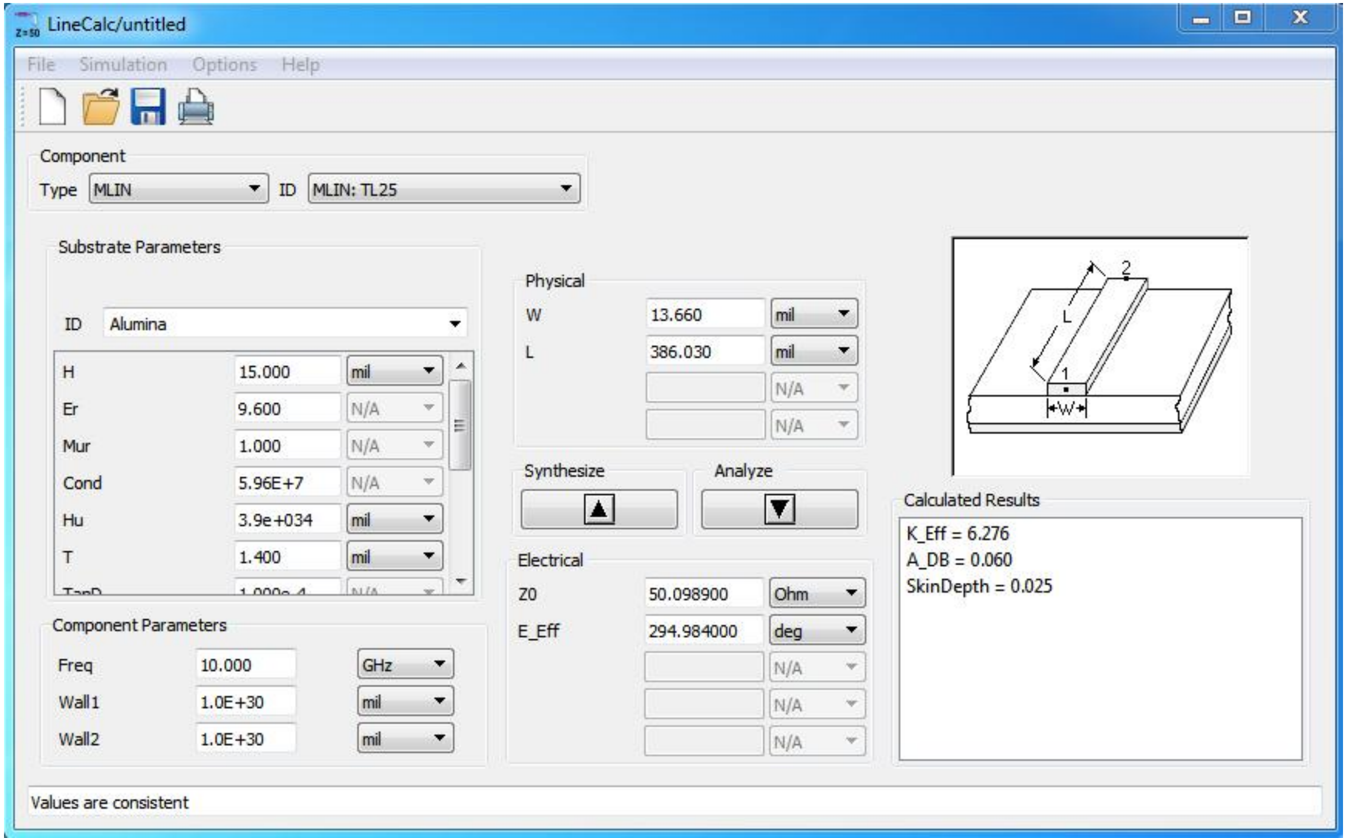
Microstrip standardization

- Typically the height of the dielectric layers is also standardized in mil

Standard Thickness
RO4003C: 0.008" (0.203mm), 0.012 (0.305mm), 0.016" (0.406mm), 0.020" (0.508mm) 0.032" (0.813mm), 0.060" (1.524mm) **RO4350B**: *0.004" (0.101mm), 0.0066" (0.168mm) 0.010" (0.254mm), 0.0133 (0.338mm), 0.0166 (0.422mm), 0.020" (0.508mm) 0.030" (0.762mm), 0.060" (1.524mm)

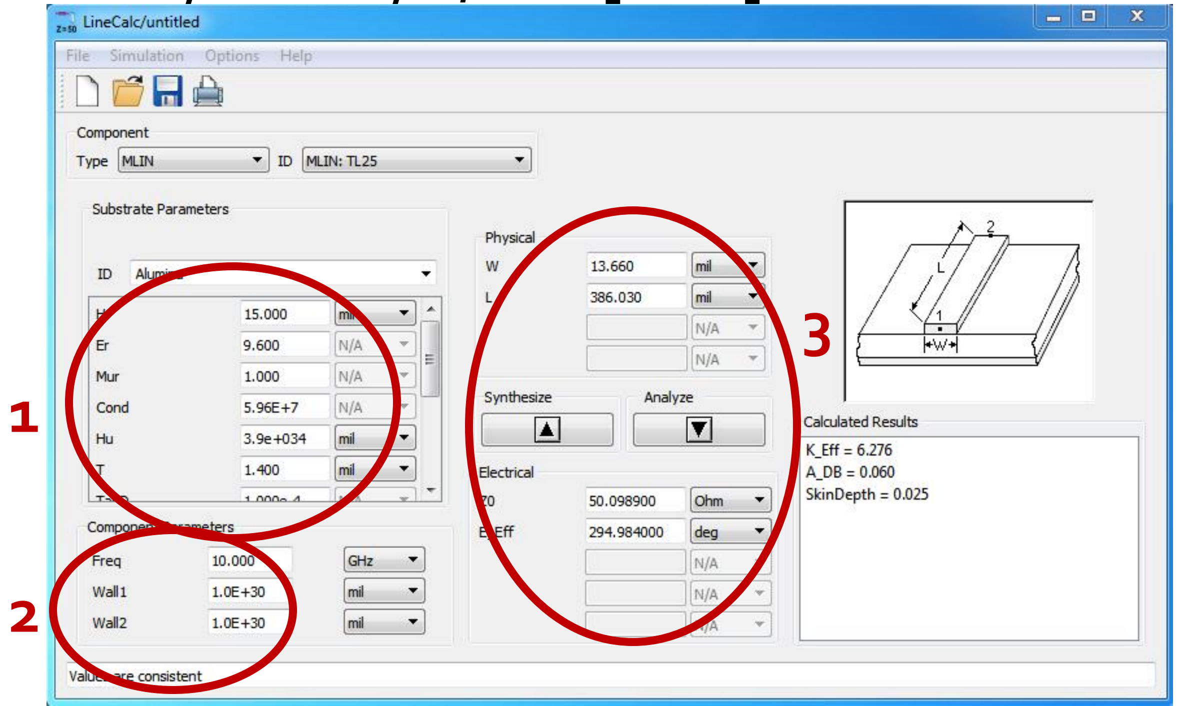
ADS linecalc

- In schematics: >Tools>LineCalc>Start
- for Microstrip lines >Tools>LineCalc>Send to Linecalc



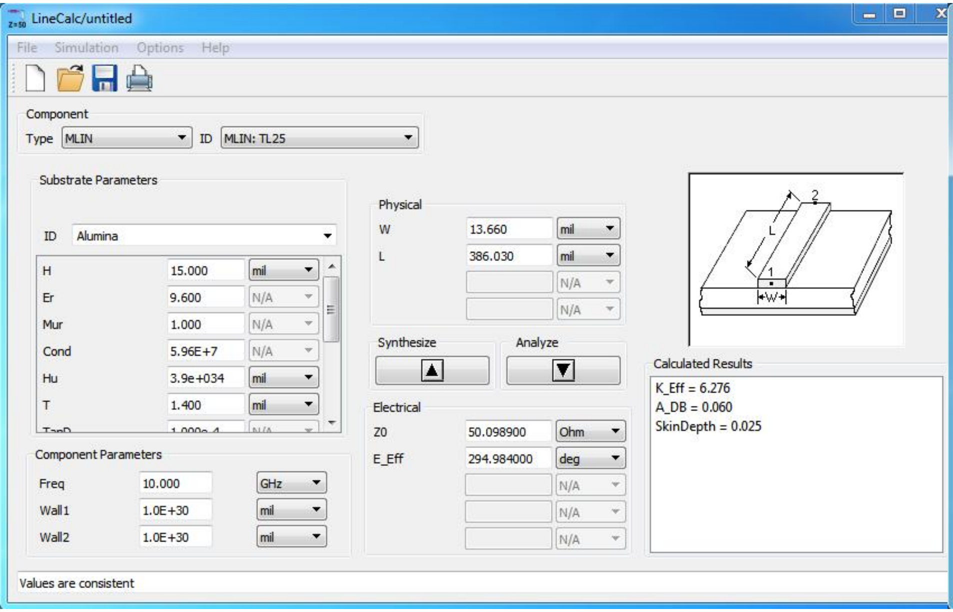
ADS linecalc

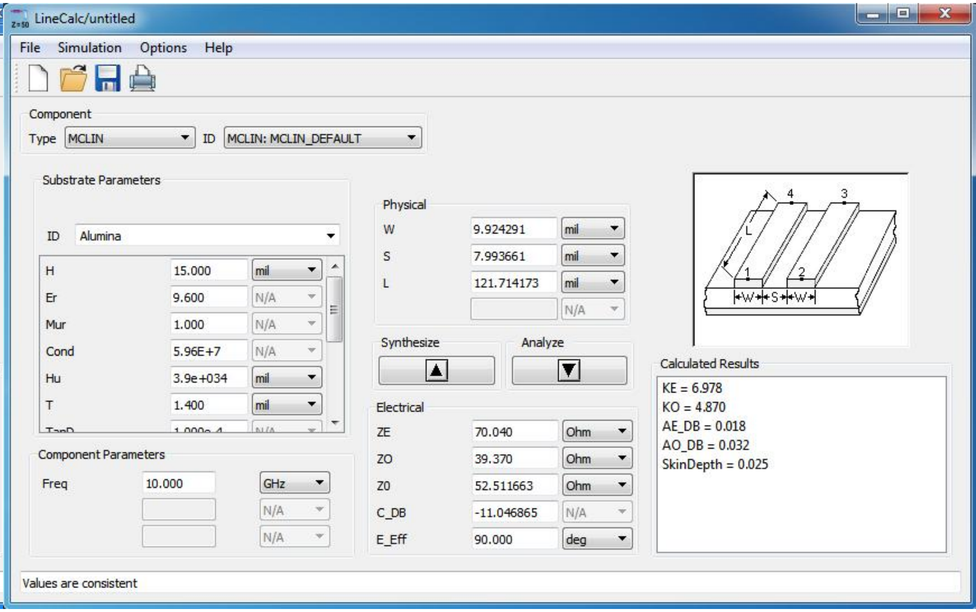
- 1. Define substrate (receive from schematic)
- 2. Insert frequency
- 3. Insert input data
  - Analyze: W,L → Zo,E or Ze,Zo,E / at f [GHz]
  - Synthesis: Zo,E → W,L / at f [GHz]



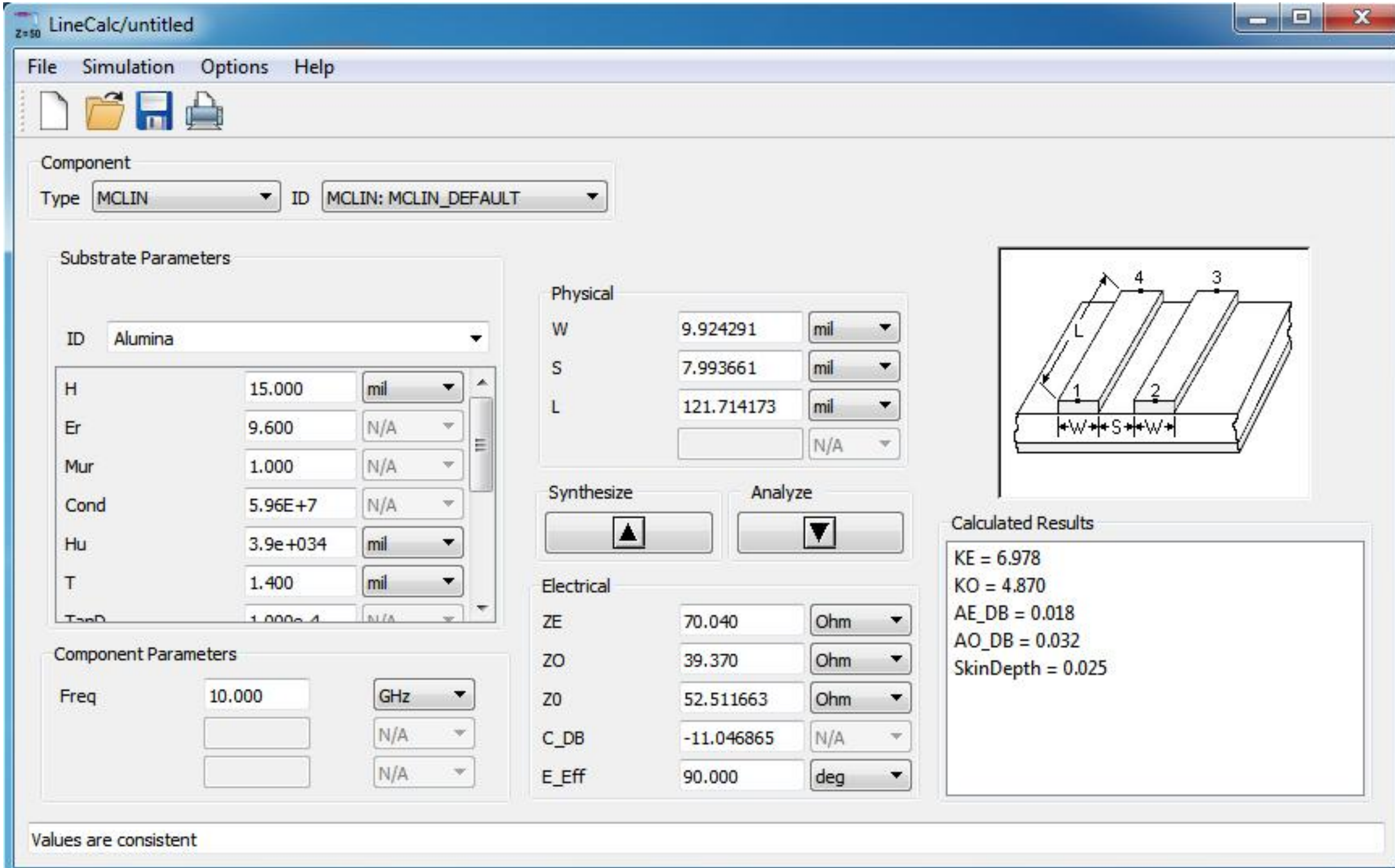
ADS linecalc

- Can be used for:
  - microstrip lines MLIN: W,L ⬄ Zo,E
  - microstrip coupled lines MCLIN: W,L,S ⬄ Ze,Zo,E





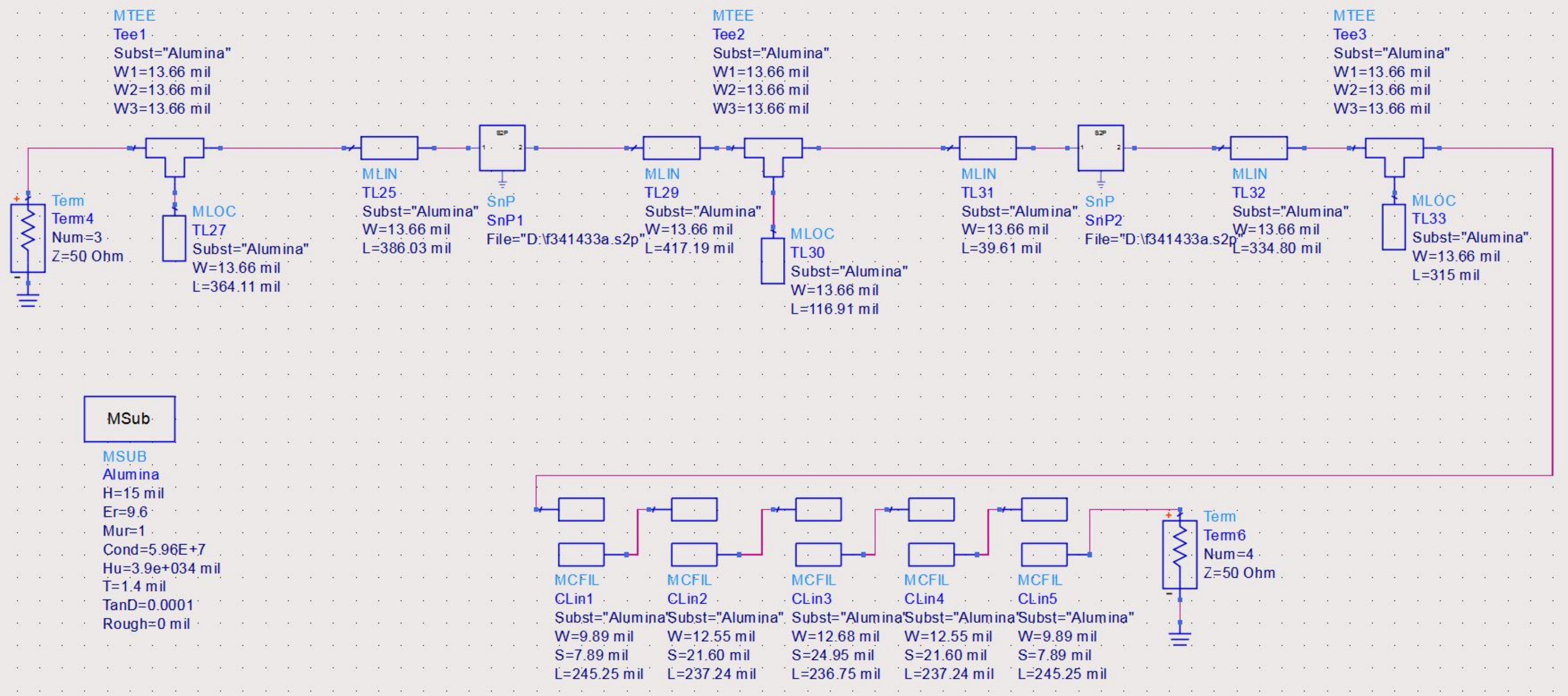
ADS linecalc



Transmission lines

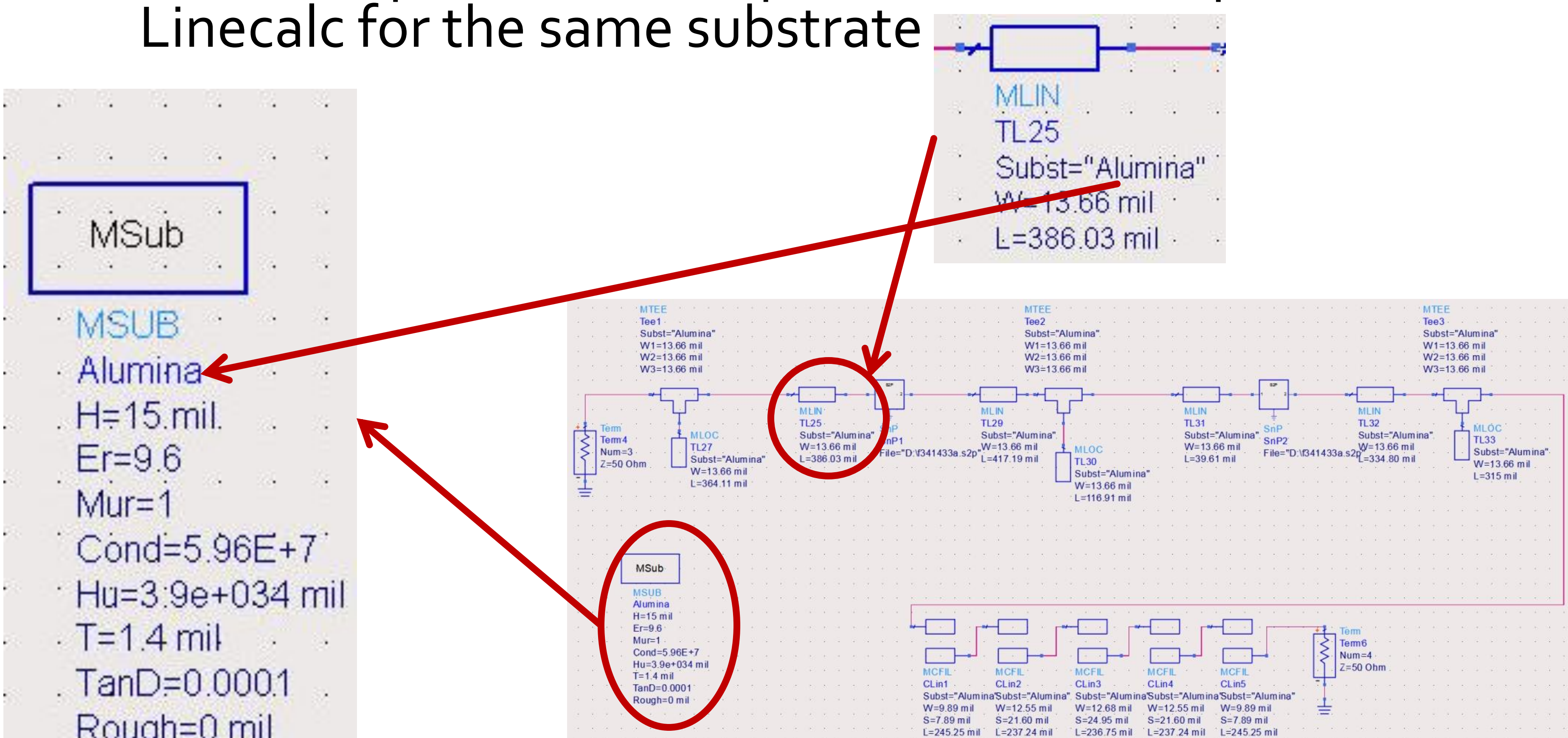
- http://rf-opto.etti.tuiasi.ro
- Transmission lines / Rogers
  - more precise formulas including
    - t, trace thickness
    - f, frequency
  - formulas for
    - microstrip
    - strip
    - coupled lines

Implementation in microstrip technology



Implementation in microstrip technology

- On all schematics you must have an substrate model/component
- Miscrostrip lines and coupled lines are computed in Linecalc for the same substrate

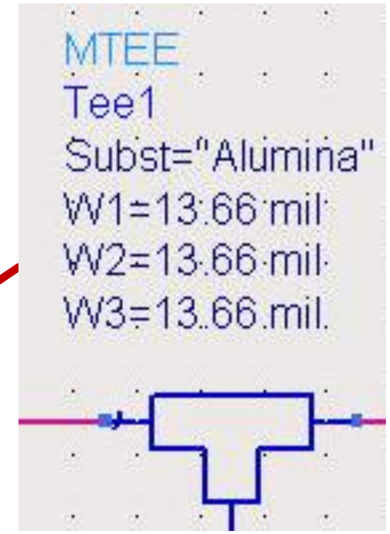


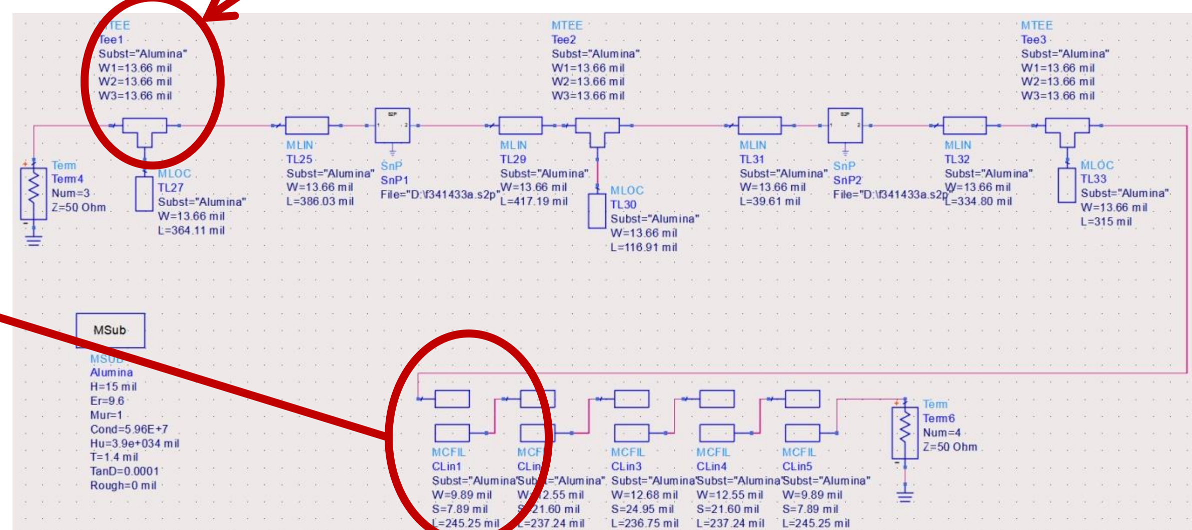
Implementation in microstrip technology

- We use components from the “Transmission Lines – Microstrip” pallete
  - MSUB - substrate
  - MLIN – series line
  - MLOC – open-circuit shunt stub
  - MTEE – modeling of T junction (shunt stub connection to main line)
  - MCFIL – coupled line filter section (more accurate model than MCLIN – takes into account that two adjacent sections are physically close)

Implementation in microstrip technology

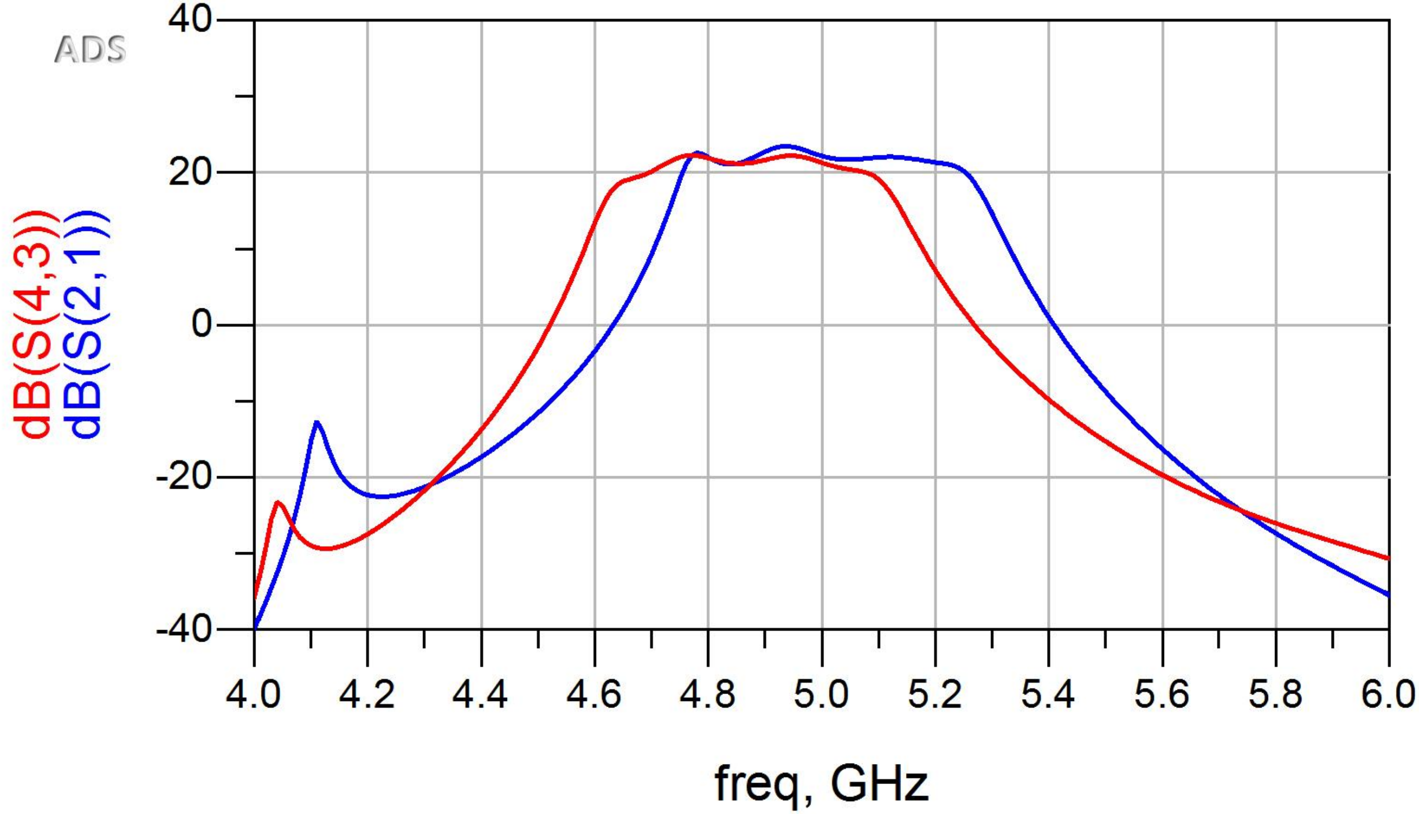
- Attention is required when inserting parameters for MTEE and MCFIL by checking in the schematic the width of the lines connected to each port.





Implementation in microstrip technology

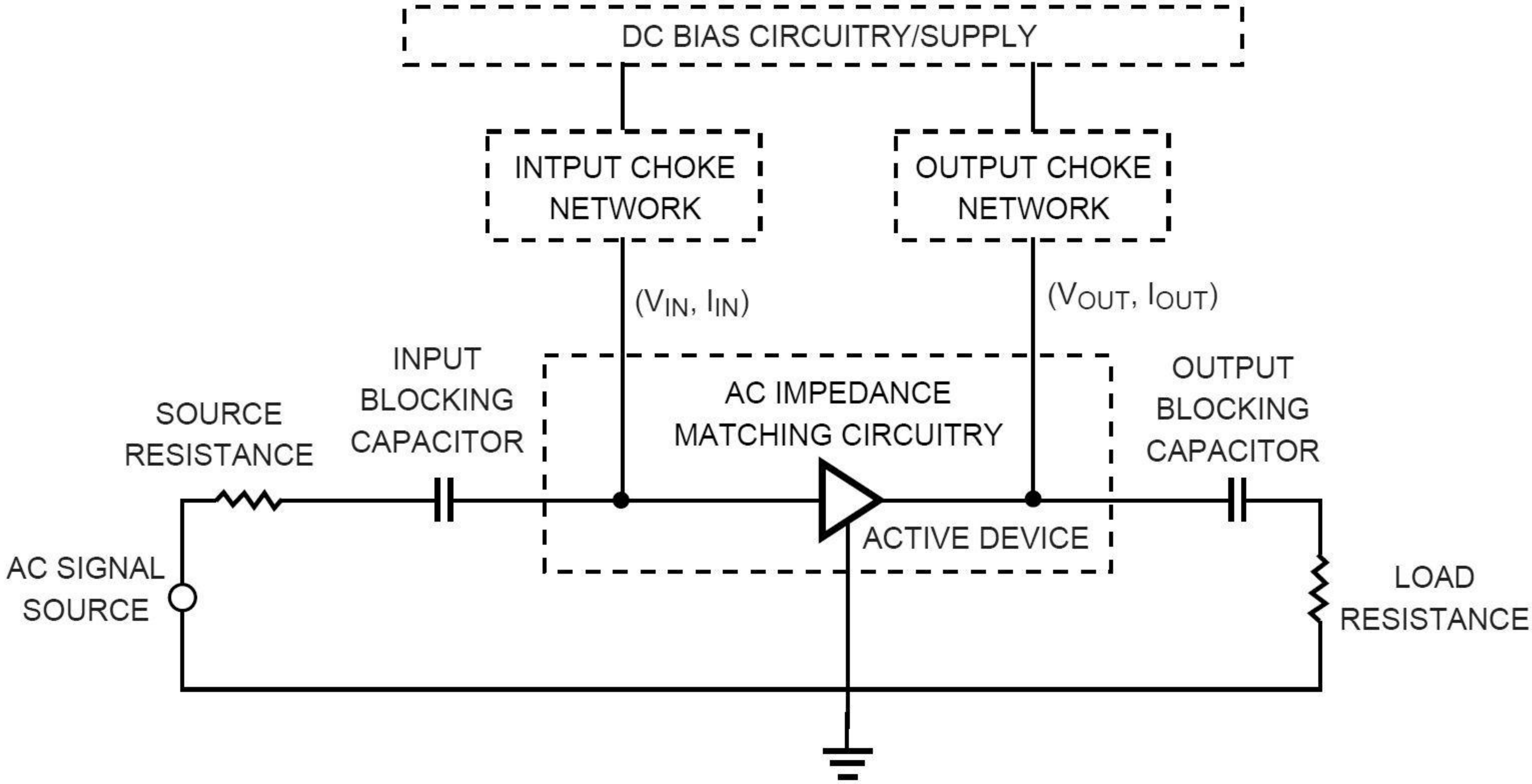
- Usually there is a shift of the transfer function (red) towards lower frequencies compared to the ideal model (blue)
  - due to the MCFIL/MCLIN difference
- Tune the length of filter elements to move the filter bandwidth around fo = 5GHz



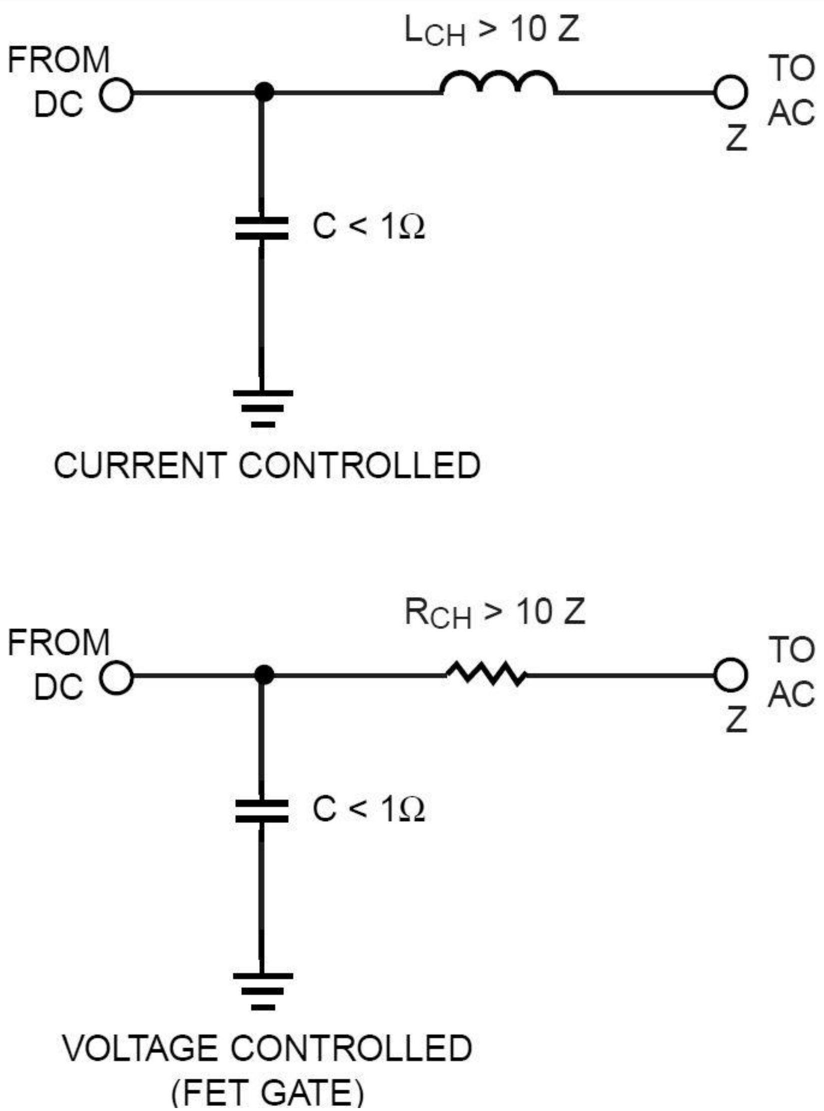
DC Bias

- https://rf-opto.etti.tuiasi.ro
- Agilent Application Notes
  - decoupling signal from DC Bias circuitry
  - DC Bias circuits for microwave transistors
- Appcad has tools for designing DC Bias circuits

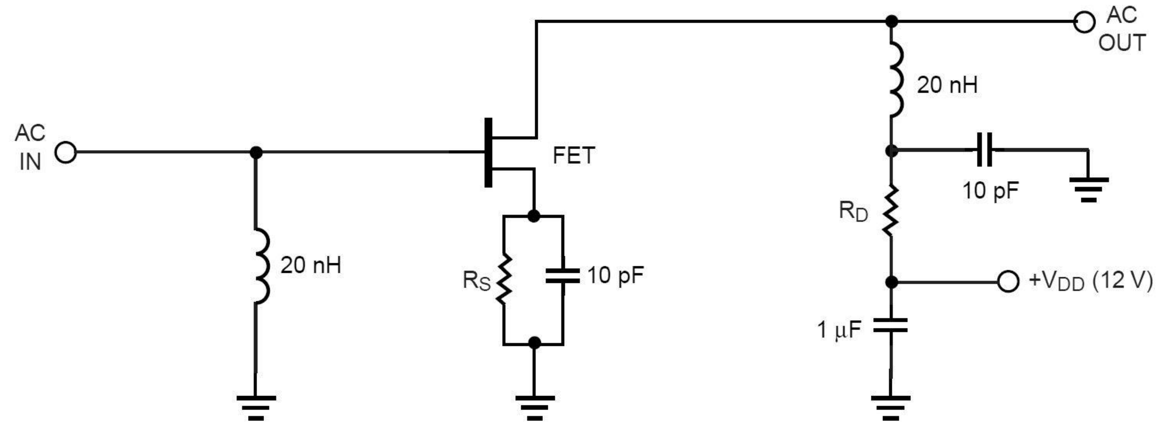
DC Bias

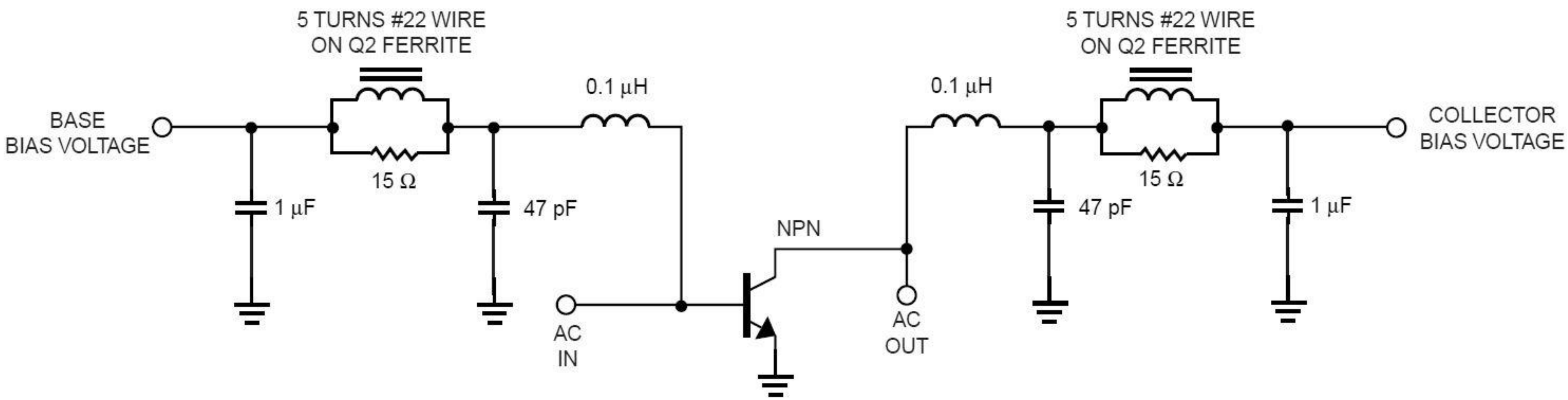

DC BIAS CIRCUITRY/SUPPLY
INTPUT CHOKE NETWORK
OUTPUT CHOKE NETWORK
(VIN, IIN)
(VOUT, IOUT)
INPUT BLOCKING CAPACITOR
SOURCE RESISTANCE
AC IMPEDANCE MATCHING CIRCUITRY
OUTPUT BLOCKING CAPACITOR
ACTIVE DEVICE
AC SIGNAL SOURCE
LOAD RESISTANCE

DC Bias, typical choke

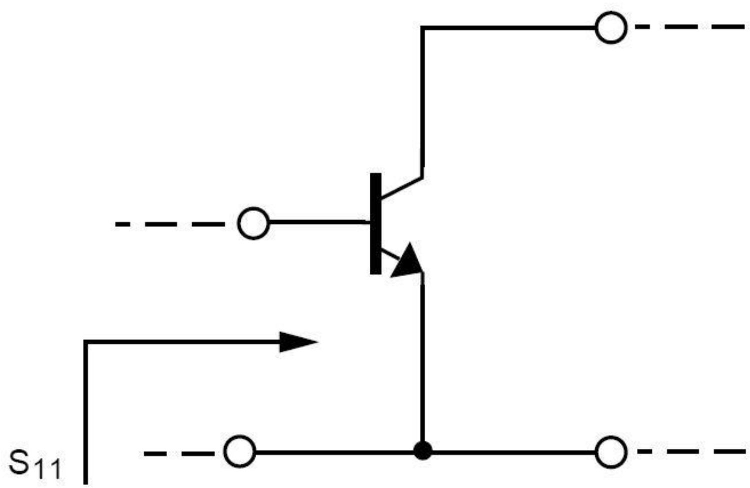


DC Bias, typical schematics/values



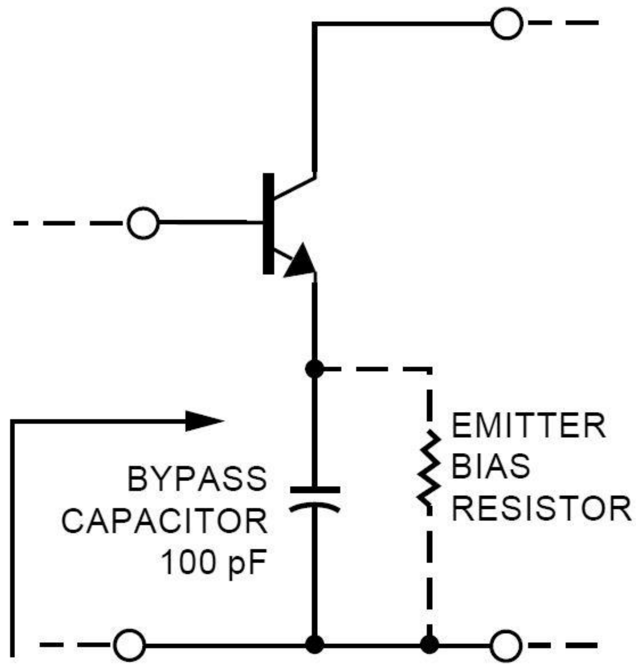


DC Bias, elements in E/S



S_{11} (AT 4 GHz) = 0.52 ∠ 154°

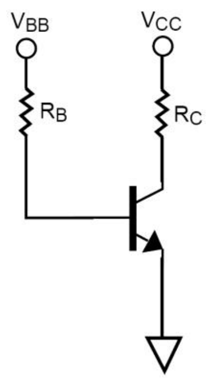
S_{11} (AT 0.1 GHz) = 0.901 ∠ –14.9°



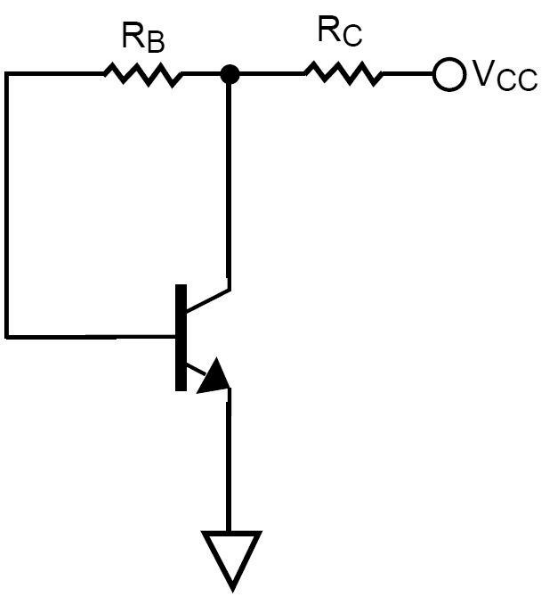
S'_{11} (AT 4 GHz) = 0.52 ∠ 154° UNCHANGED AT 4 GHz

S'_{11} (AT 0.1 GHz) = 1.066 ∠ -8.5° $|S_{11}|>1$ AT 0.1GHz

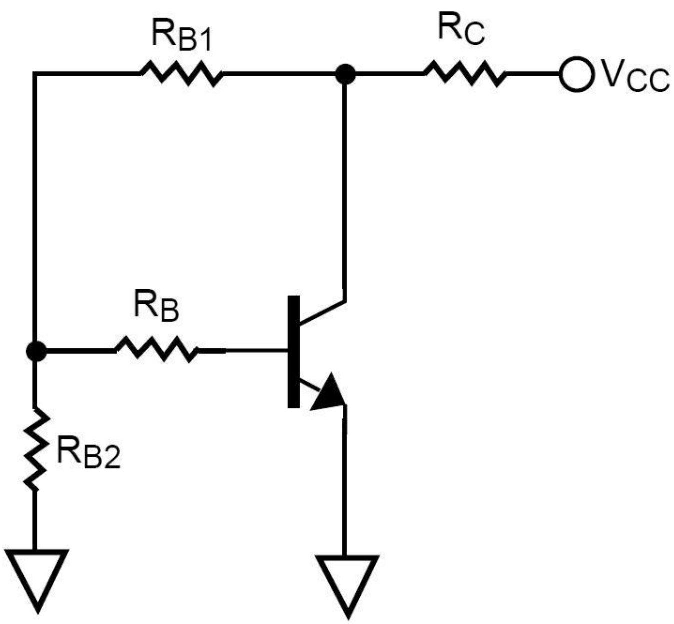
DC Bias, bipolar transistors



NON-STABILIZED



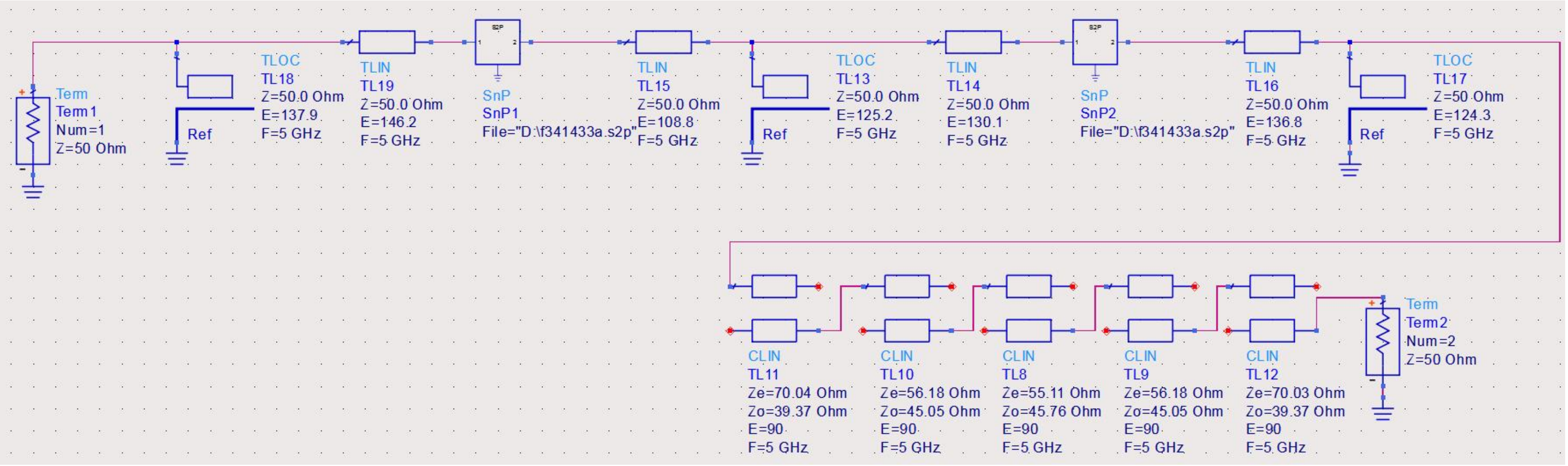
VOLTAGE FEEDBACK



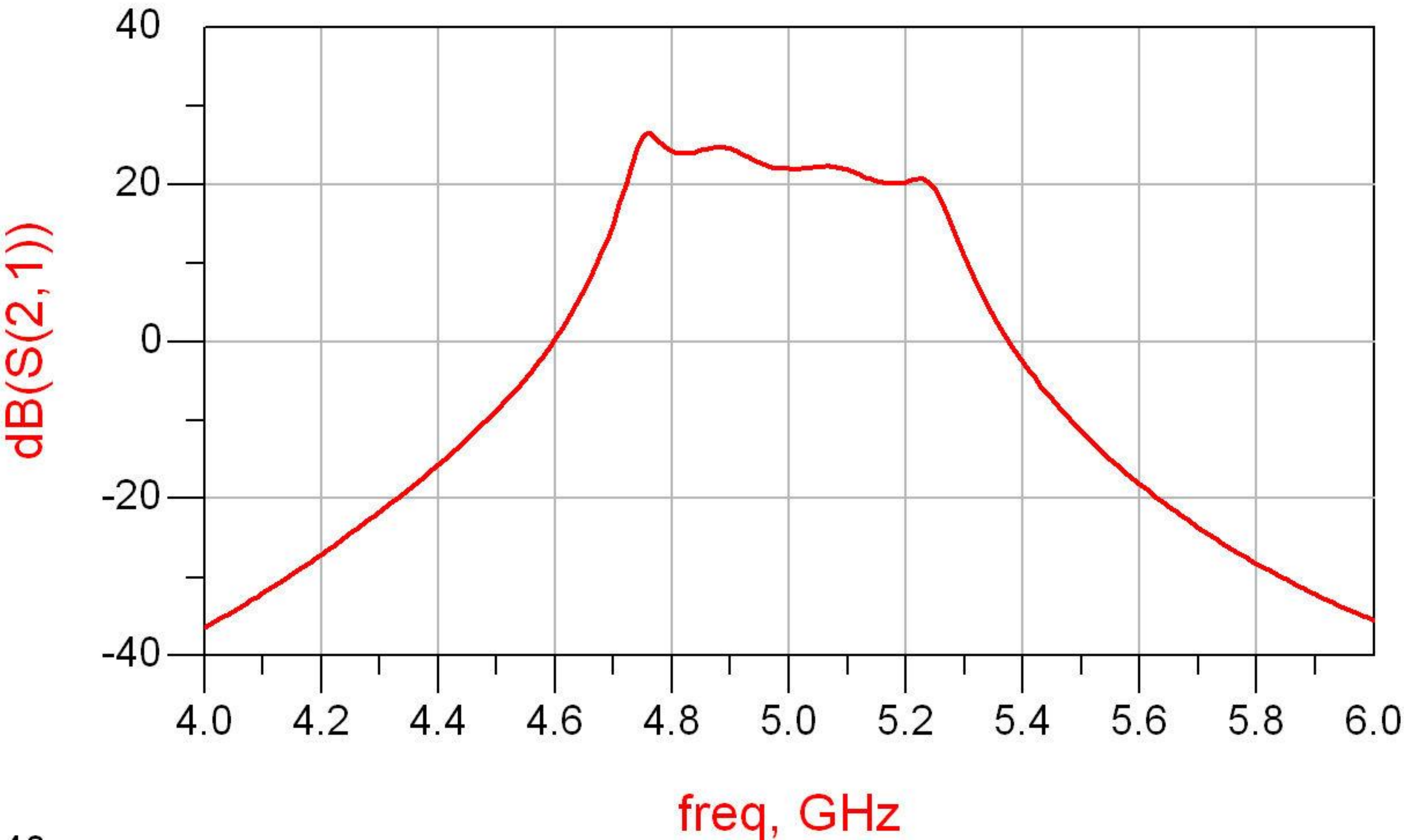
VOLTAGE FEEDBACK AND CONSTANT BASE CURRENT SOURCE

Example project

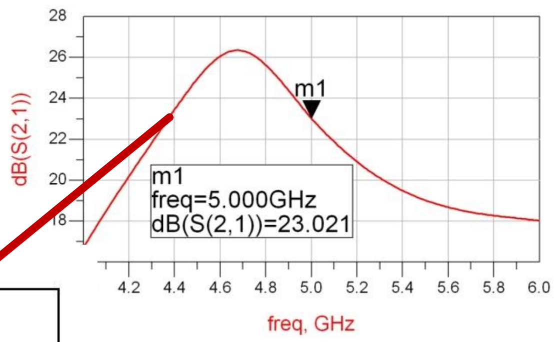
- Unify the two schematics
  - L10 – amplifier
  - L12 – filter

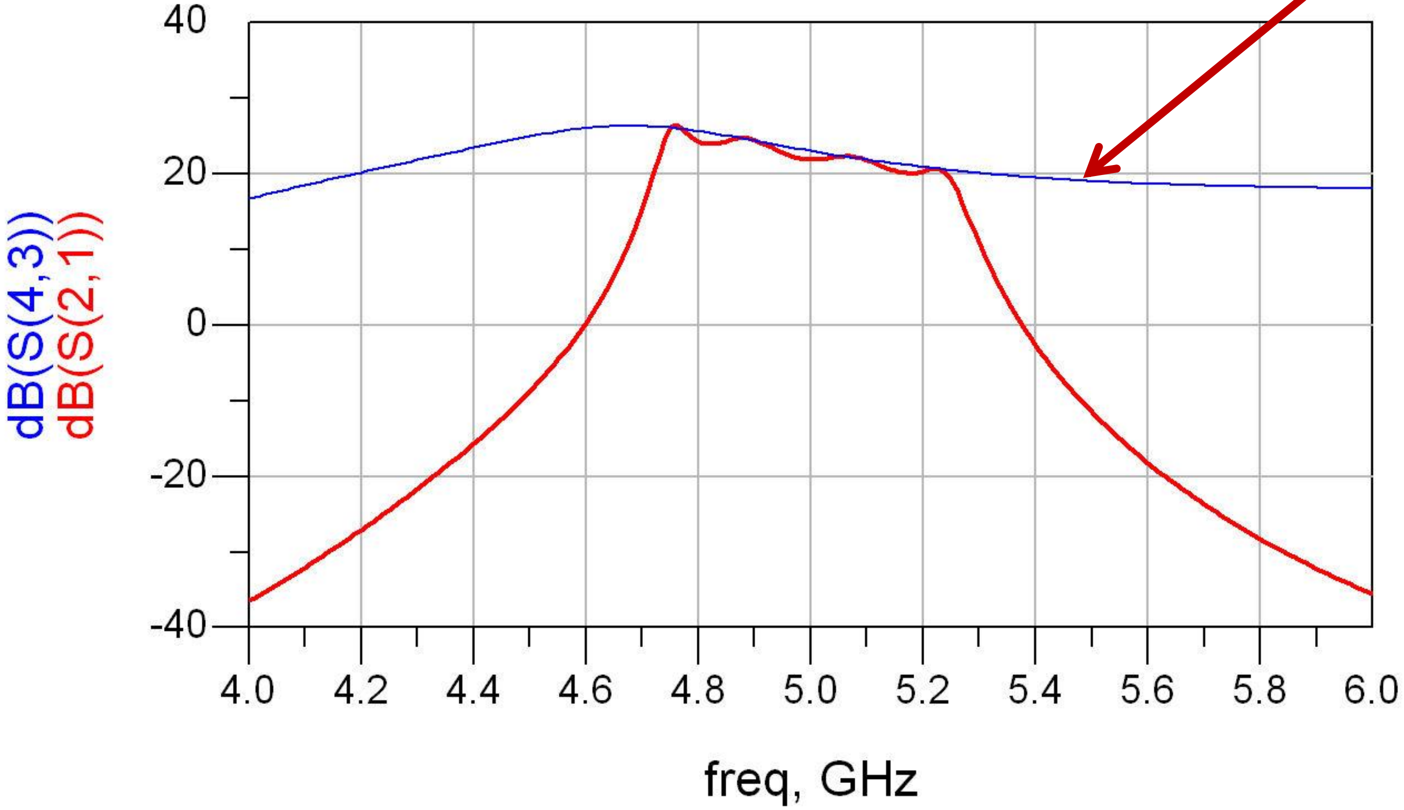


Result (unbalanced)

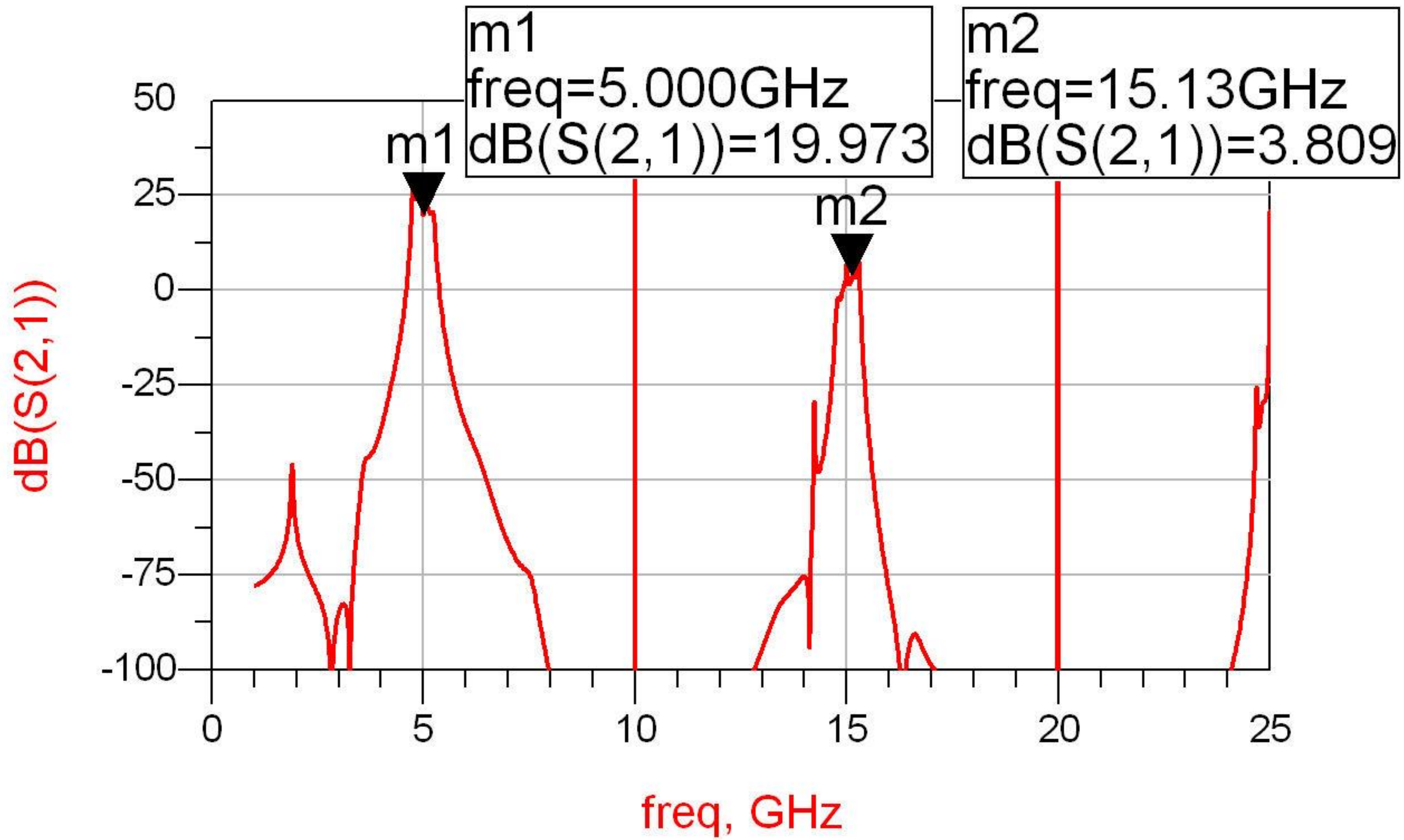


Result (unbalanced)



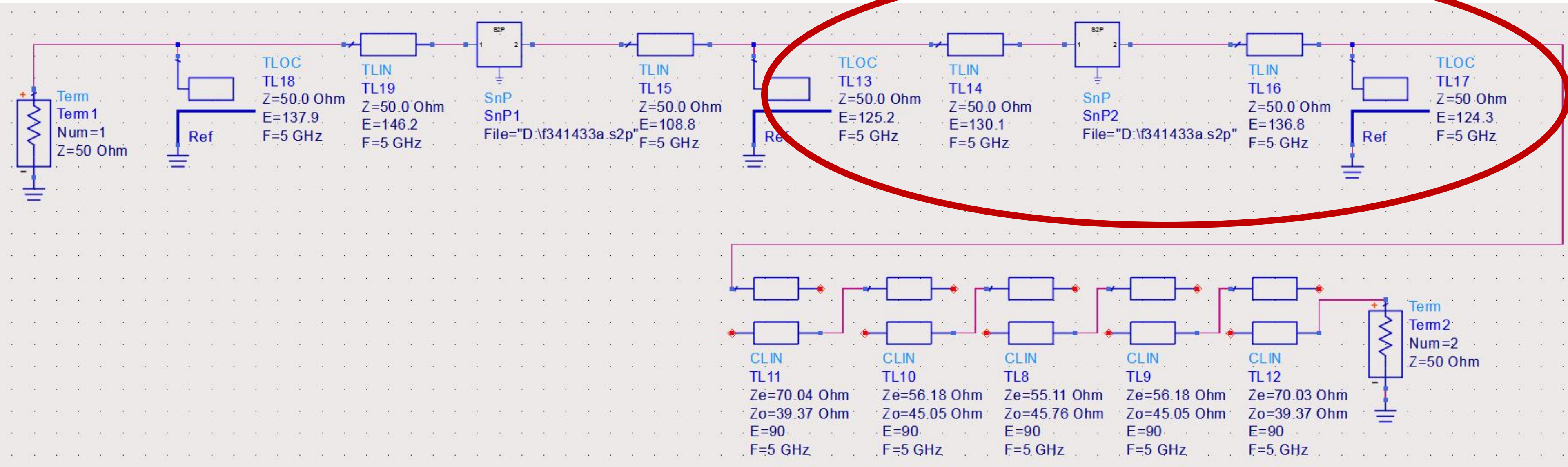


Result (~periodic in frequency)

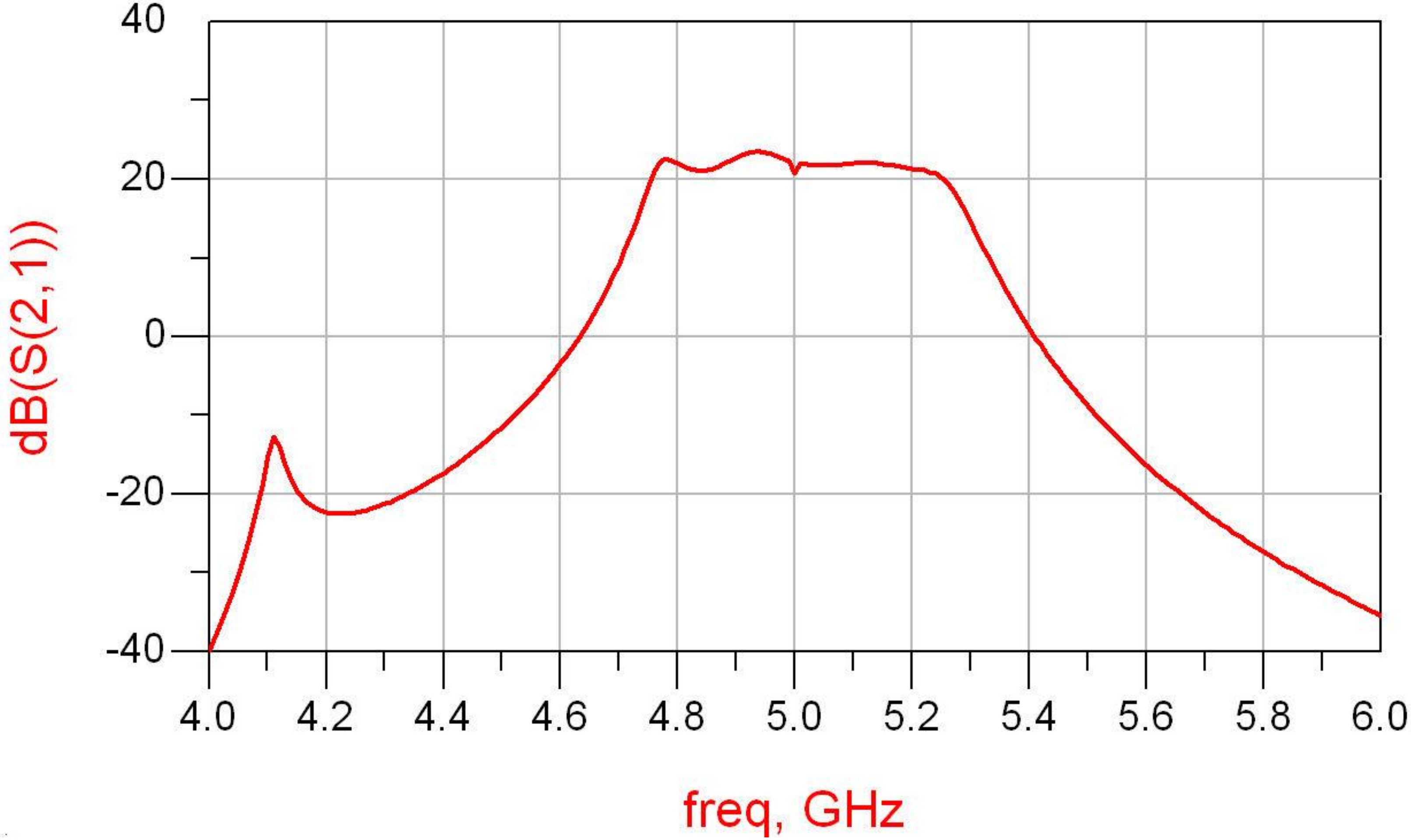


Tune -> balance

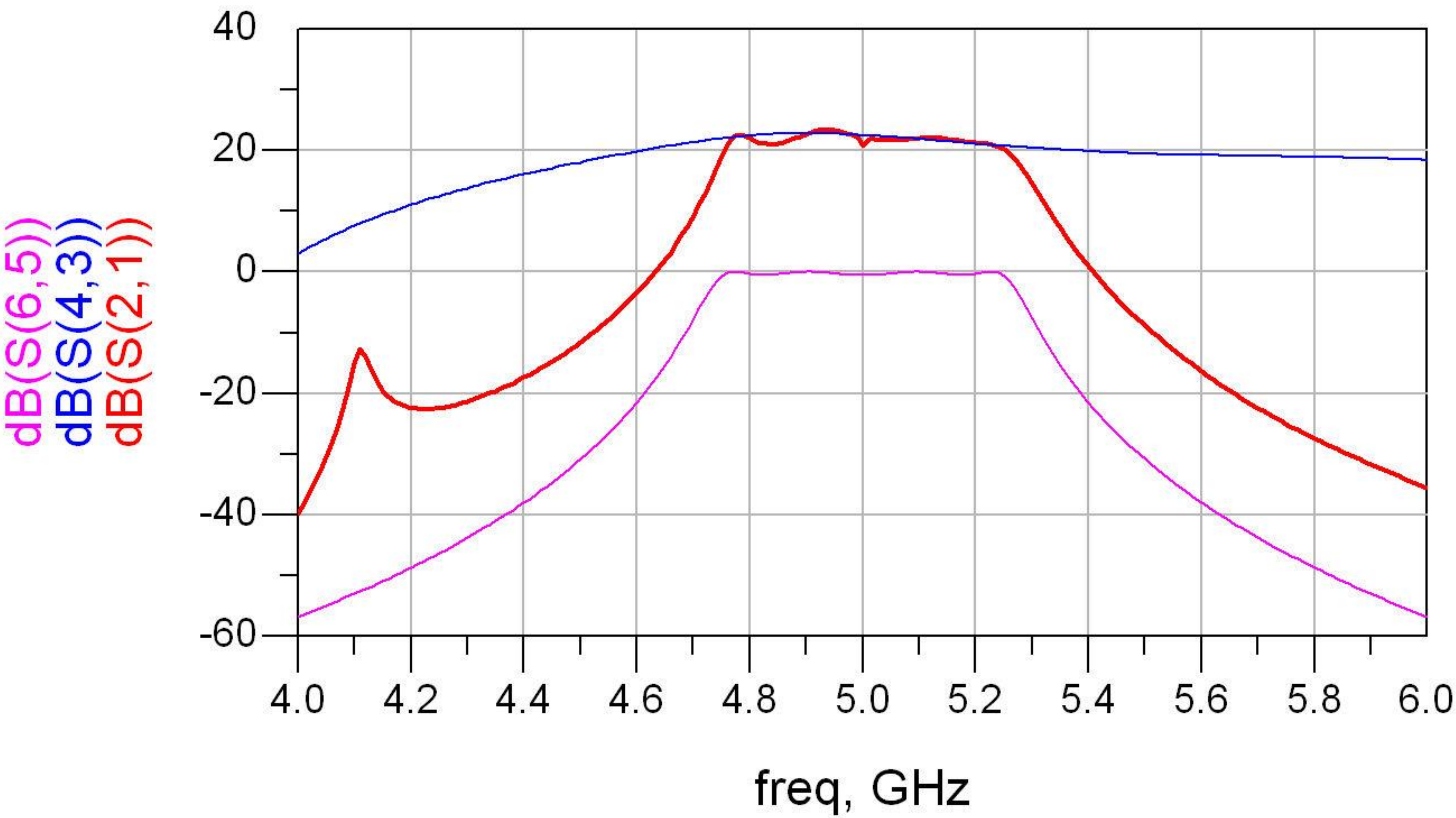
- purpose: balance the gain characteristic of the amplifier (maximum at design frequency)
  - favor tuning lines at the end of the amplifier
    - eliminate/minimize effect of the tune on noise



Tune -> balance, result

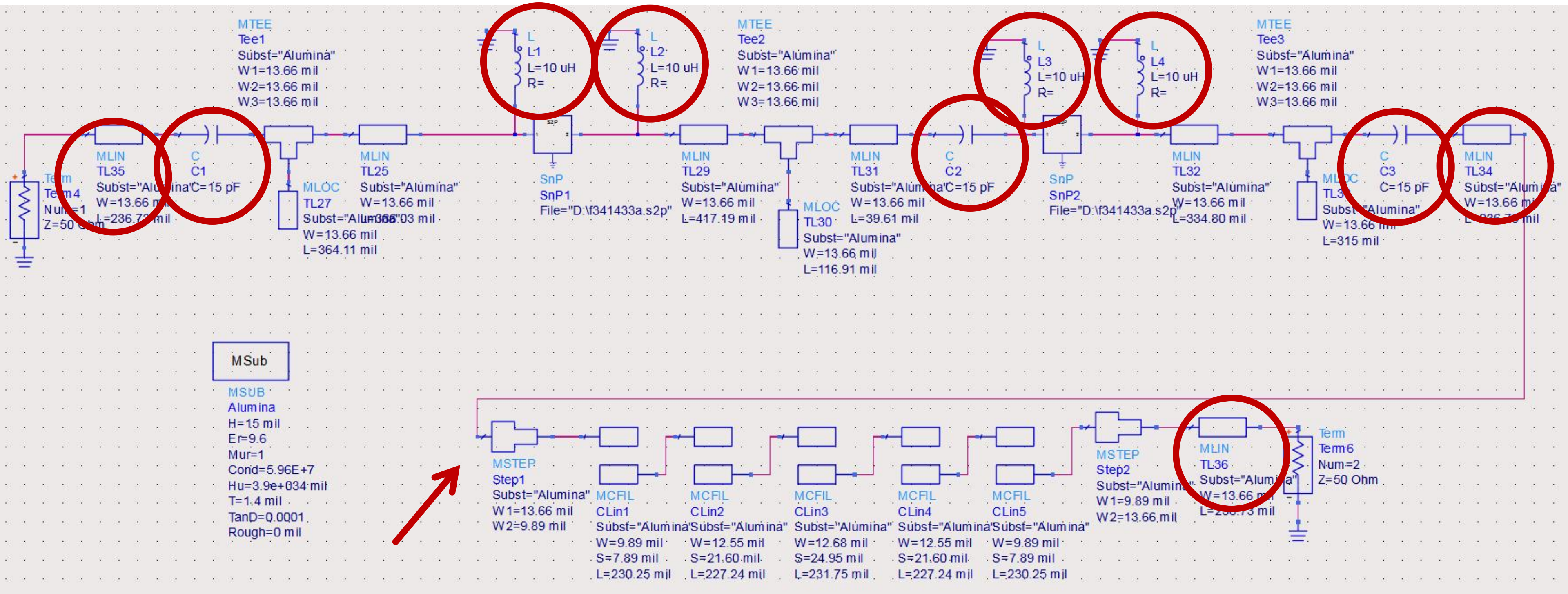


Amplifier, Filter, Total

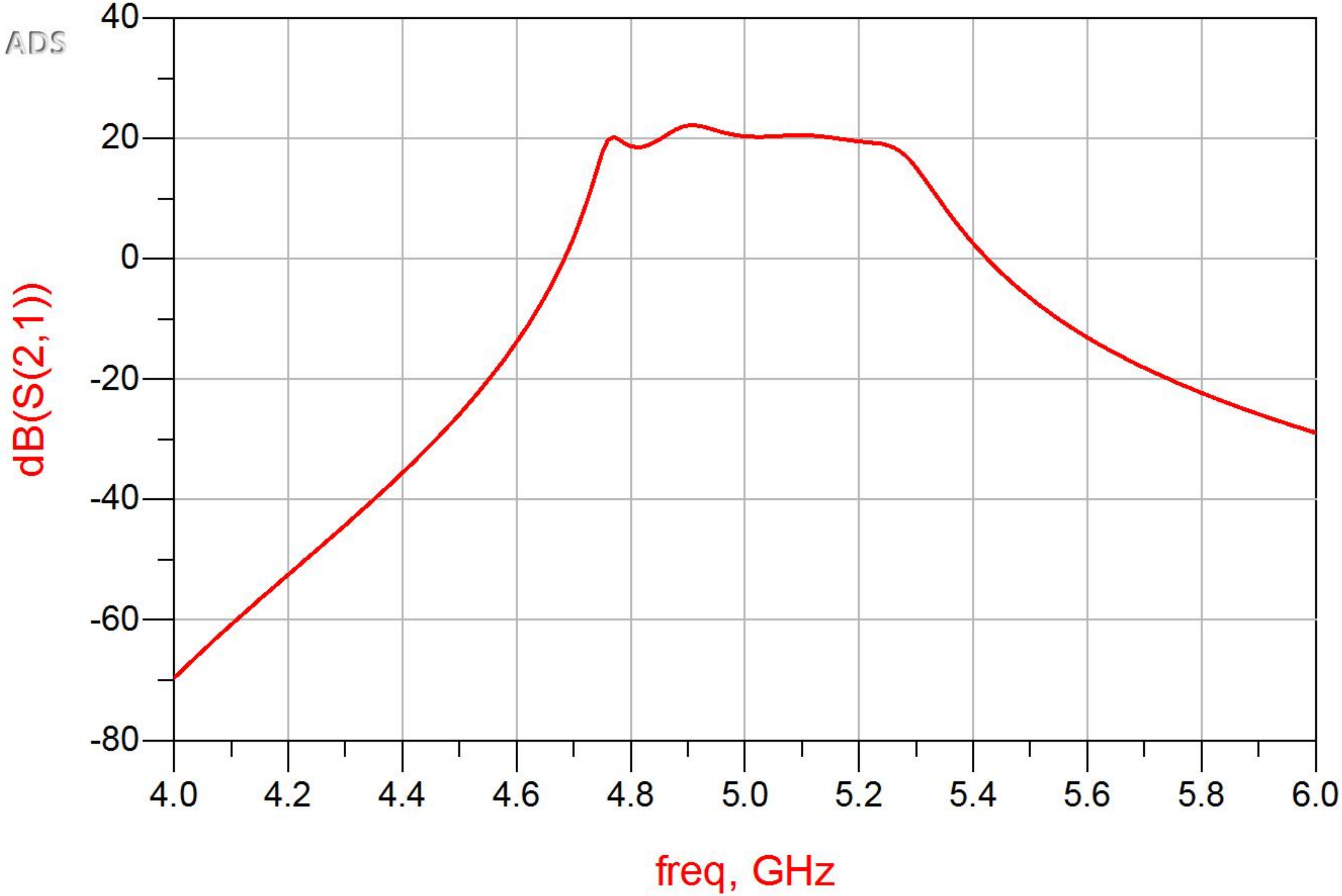


DC Bias elements in ADS schematic

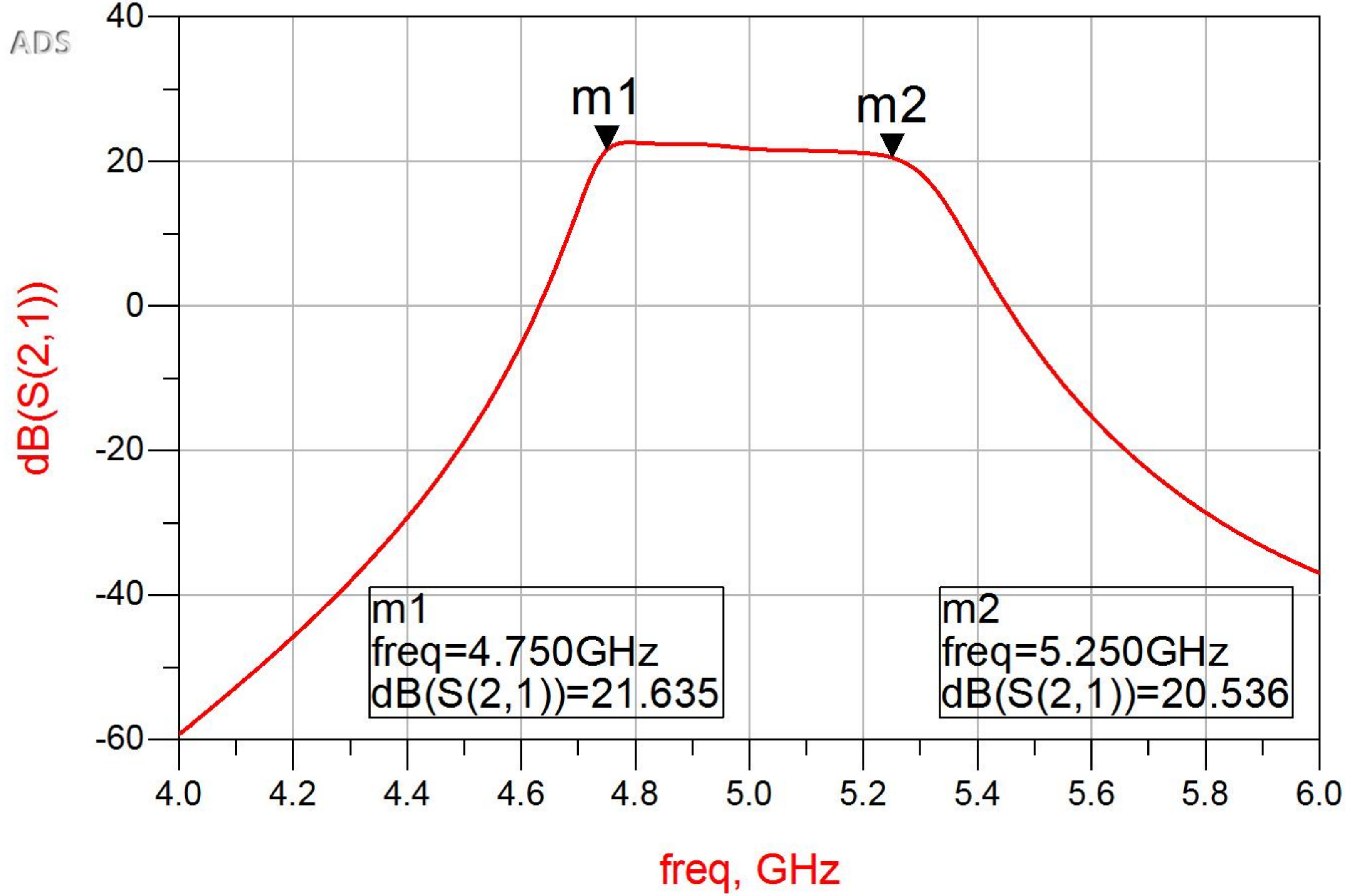
- Insert L (RF chokes) and C (decoupling)
- additional 50Ω connection lines
  - source
  - load
  - between blocks



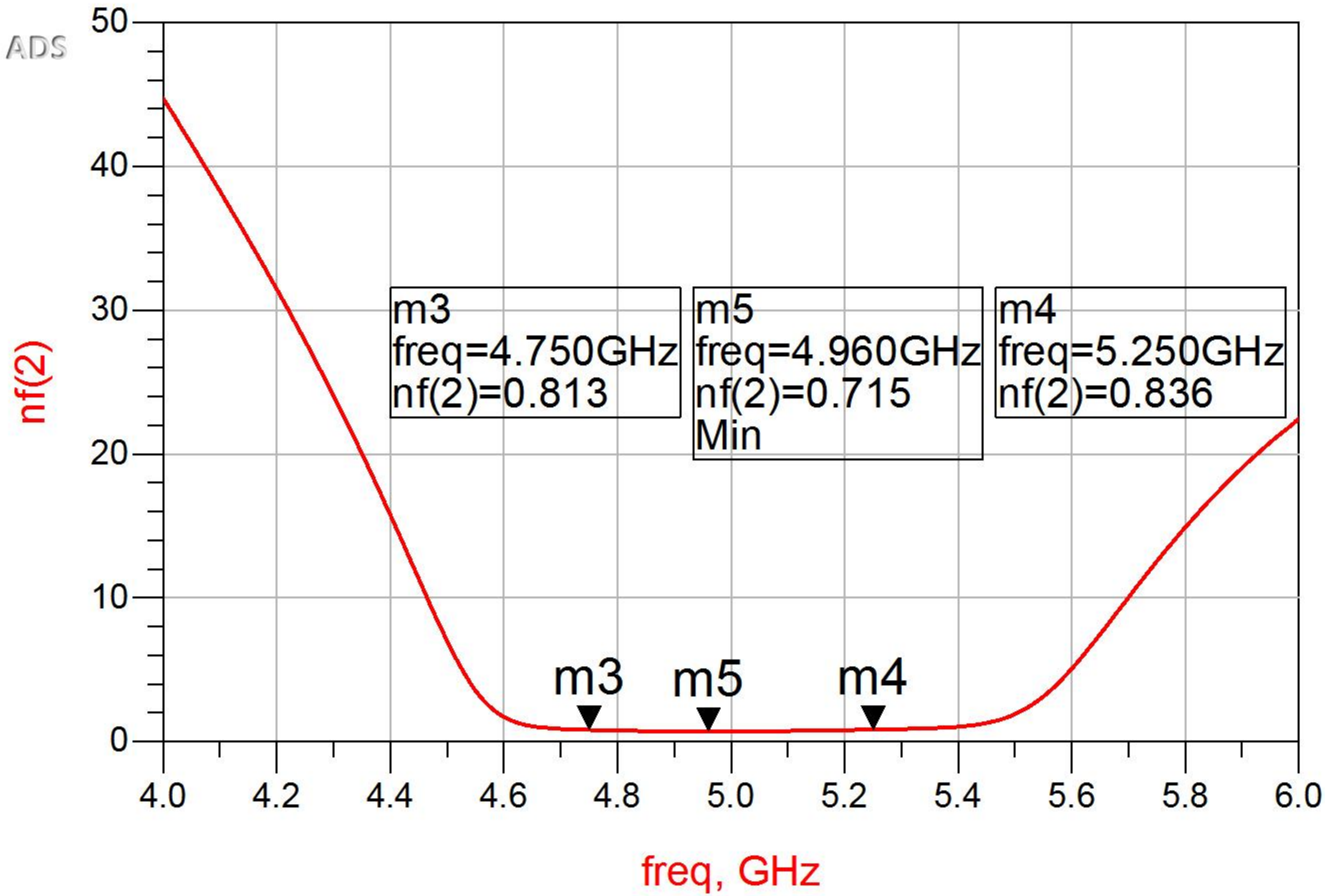
Gain -> Tune/Optimization



Final result (Gain)

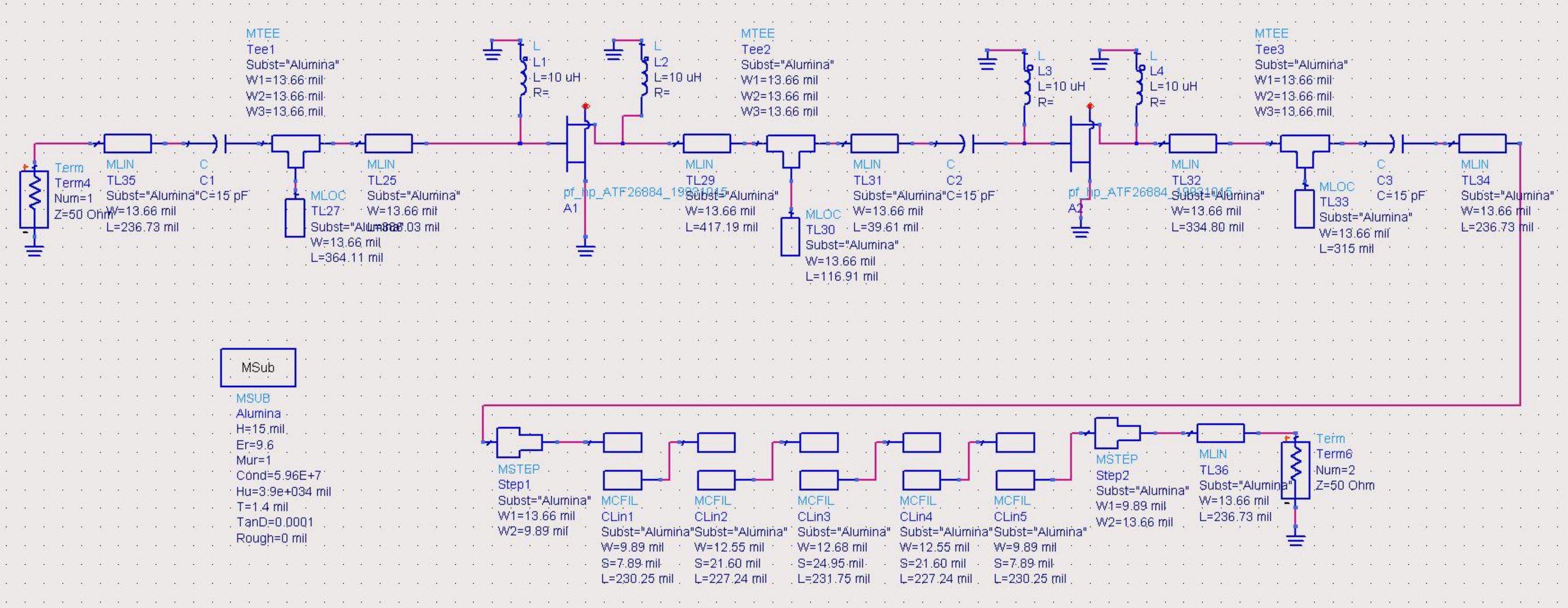


Final result (Noise)

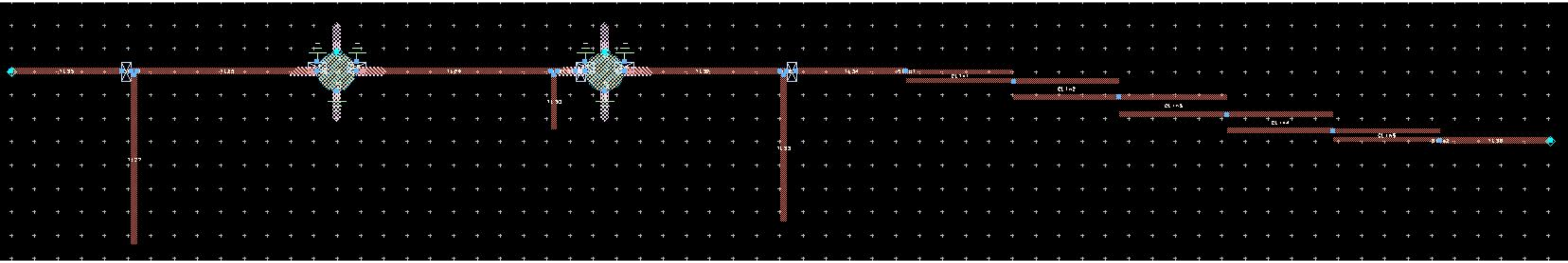


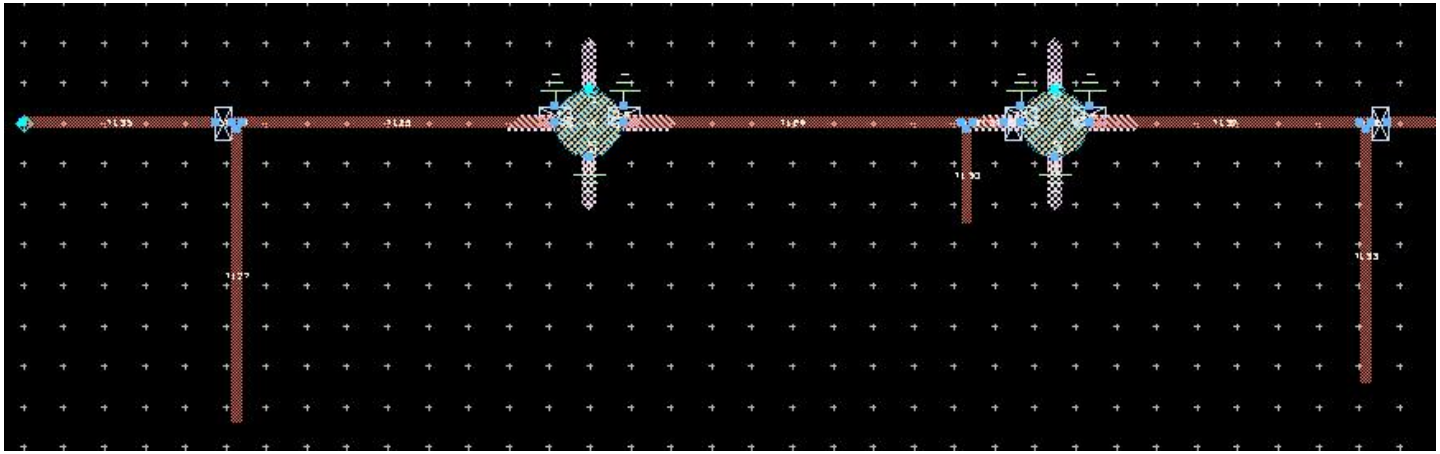
Layout (Example)

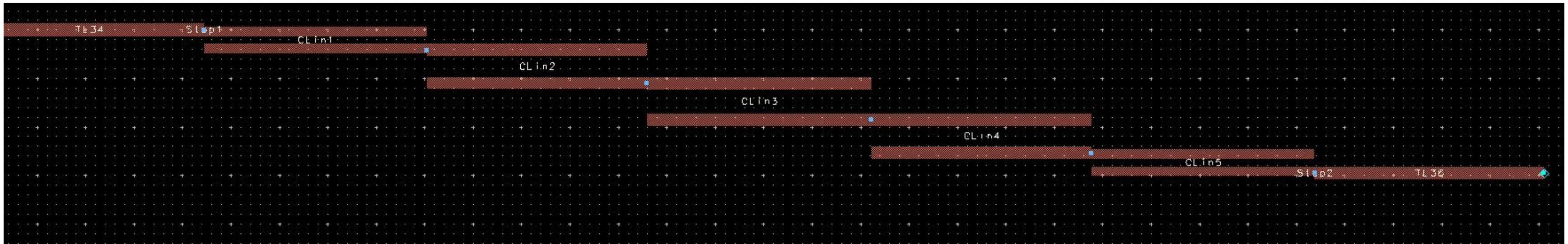
- Temporary replacement of the transistors and lumped elements (LC) with elements for which ADS has case information



Layout (Example)







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